Volume II

TELEVISION for RADIO MEN

EDWARD M. NOLL

TELEVISION FOR RADIOMEN

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TELEVISION FOR RADIOMEN

Volume II

by EDWARD M. NOLL

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Dedicated to my uncle and aunt
Mr. and Mrs. H. Roy Eberly

PREFACE TO REVISED EDITION

When the first edition of this book was published, I wrote of how much I appreciated and recognized the contributions, guidance, and research of the many who had contributed to the development of a practical television system and whose work is necessarily a part of this television text. Today I must extend my gratitude to many, many more organizations and individuals who have since shared in the refinement and colossal growth of television. Despite the impressive expansion of television broadcasting during the few years intervening, its present status represents only a small percentage of its potential.

This edition of *Television for Radiomen* has been revised thoroughly and enlarged. The book should be even more serviceable in revision because the author has included information on new developments in television and eliminated sections which have been little used. New sections have been added and two new chapters included that present comprehensive data on UHF and transistors. The increasingly important subject of color television occupies the entire, new, second part of the book.

It is hoped that the book will encourage the student of television to pursue his studies diligently and that he will not become discouraged over inability to understand all things quickly and completely. But few things in the universe are understood with complete finality. We must learn to work with those phases of television that we do not understand completely. As we work and observe results, our knowledge and understanding increase.

It is essential that the television student develop the habit of having his mind ready to pursue and accept new knowledge, even when he must interpret an old subject differently in the light of new disclosures. He must study continuously to keep abreast of new products designed to meet specified performance levels in various price ranges. In a field of so many variables and cost factors, a "best without reservation" is almost a non-existent product.

Since there are so many differing circuit techniques, as functions of cost and application, each variation cannot be detailed, even in a text of considerable length. Our task has been to present materials necessary to the student in building the solid foundation and gaining the wide, thorough background he will need if he is to progress with the science in the future.

In conclusion I want to thank Stanton Snyderman for preparation of the fine illustrations for the revised edition and Dorothy Meeder for her helpful assistance in stenographic work.

EDWARD M. NOLL

PREFACE TO FIRST EDITION

Television For Radiomen was written to serve as a television text for practical radiomen, and television students in the final semesters of their studies. The text assumes a rather thorough basic knowledge of radio circuits, and, if practicing radiomen have been away from study for some time, it is advisable to review a good radio text. The material is arranged for progressive, orderly study in a sequence which the author has found most effective for classroom or written-lesson presentation of television. Treatment can be mathematical or nonmathematical depending on the scope of the course or the needs of the individual studying the text. Chapter 14 is a lengthy chapter on the mathematical aspects of various television circuits and gives to those who have the necessary background that added understanding necessary to handle effectively television jobs of a more complex nature. At the same time, practicing television technicians who do not require an extensive mathematical background can progress through the book without being impeded by mathematical formulas, interpretations, and derivations. Reference is made at various points in the text to the mathematical presentations in Chap. 14. As a further aid to study, questions and a thorough bibliography follow each chapter.

A technical author must necessarily feel humble when he reflects on the creative labor, organization, and toil of the many persons who have contributed to such a comprehensive field as television and, therefore, have contributed to this book. And when he considers the numerous persons who have contributed to his own education and abilities, he feels that only a little of himself, excluding some tedious work, has gotten into the work. This author feels indebted to many individuals and organizations.

For willing cooperation and permission to use manufacturers' data special thanks is given to the following men and the manufacturers they represent: A. W. Bernsohn, A. Liebscher, E. L. Clark, R. S. Burnap of RCA, E. D. Lucas of Philco, W. L. Parkinson of GE, A. C. Lescarboura representing Dumont, J. F. Bigelow of Farnsworth, F. L. Uhrus of Motorola, A. B. Friedman of J.F.D., R. K. Arbogast of Hickok, L. J. A. van Lieshout of Norelco and S. Horbach of Brach.

I thank Mr. Byron Young for the very fine illustrations. For typing and proofreading, I appreciate the careful work of the Kerner sisters, and Mrs. Evelyn Bondy.

I am grateful for the help and encouragement of officials and members of the faculties of Temple University Technical Institute and Trenton Technical School, and to Mr. Avery Pitt of Old York Road Publishing Company.

It has been my good fortune to have the guidance of the fine group of editors who manage our radio trade journals—Lewis Winner of *Service,* Oliver Read of *Radio News,* Fred Shunaman of *Radio-Electronics,* and Sanford Cowan of *Radio Service Dealer.* I thank them for permission to use again material which the author prepared originally in article form for these journals.

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Chapter 10

LARGE-SCREEN RECEIVERS AND PROJECTION

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114. *Large Picture Receiver Description and Review*

The RCA chassis, Figs. 210 a, b, and c, is a typical modern receiver having a large screen direct-view picture tube. It consists of a 16-channel turret tuner—12 VHF and 4 UHF channels. In VHF position, the tuner employs a 6BQ7A cascode r-f stage, a 1N82 crystal mixer, a 6AF4 local oscillator, and a 6BQ7A cascode i-f stage. For UHF operation, a crystal mixer, 6AF4, local oscillator, and 6BQ7A cascode i-f stage are used.

In UHF position the r-f amplifier tube V1 does not function, acting only as an amplifier stage for VHF stations. Output from a 45-megacycle i-f stage, which is a part of the tuner, is applied to a five-stage intercarrier i-f amplifier. This amplifier has excellent bandwidth and has been trapped thoroughly to minimize adjacent channel interference. A crystal video detector with a low-pass filter output circuit transfers video to the grid of the single stage but high-gain video amplifier. The video signal with sync positive is direct-coupled to the cathode of the picture tube.

Sound is removed from the inductor of a series-resonant sound trap at the output of the video detector and is coupled to a two-stage 4½-megacycle sound i-f amplifier—the second stage of which functions as an amplitude limiter. A ratio detector and two-stage audio amplifier complete the sound system.

Composite video is also direct-coupled from the plate circuit of the video amplifier to separate vertical and horizontal sync channels. Each channel contains a separator and sync amplifier. A-G-C potential is obtained from the cathode of the horizontal sync amplifier and biases the grid of the keyed a-g-c amplifier tube. The plate of the a-g-c amplifier is keyed by positive pulses from the horizontal sweep output circuit. A-G-C bias is applied to the grids of the first two video i-f stages and that of the VHF r-f amplifier.

The vertical blocking oscillator and discharge tube excites the vertical sweep output tube which, in turn, drives the vertical deflection yoke. The horizontal sweep system consists of a syncro-guide control, a sawtooth-forming circuit,

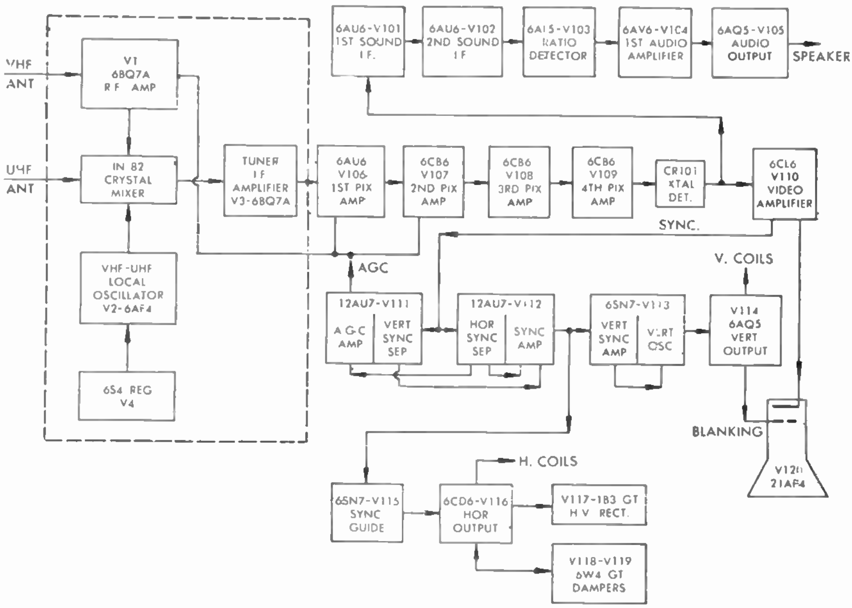


FIG. 210a Functional Block Diagram of Modern Receiver

and a horizontal sweep output tube to supply proper drive for the horizontal deflection coil. Dual parallel dampers and a high-voltage rectifier complete the horizontal system.

VHF-UHF TUNER

The RCA VHF-UHF tuner, Fig. 210b, has a 16-position turret containing 12 VHF inserts and 4 UHF inserts. In the VHF position, the antenna is attached at J2 terminals, and signals pass through the FM traps to a balance-to-unbalance transformer that is a part of each insert (transformer *T6* for VHF channels *2* through *4*). The FM traps are parallel-resonant circuits—capacitors *C1* and *C2* and inductors *L1* and *L2*. The secondaries of the antenna input transformers form series-resonant circuits in conjunction with capacitor *C5* (via terminals 4 and 5 of the turret) to supply signal to the input grid of the cascode r-f stage.

The plate of the first section of the cascode stage is direct-coupled to the cathode of the second section. An i-f trap of 43.5 megacycles (center of video i-f spectrum) connects from the plate to ground. Inductor *L6* and capacitor *C15* prevent re-active feedback into the input circuit of the cascode amplifier, keeping input impedance and termination constant.

A double-tuned transformer couples signal between the plate of the last section of the cascode stage and the input of the crystal mixer (terminals 8, 9, and 11 of the turret). A bandwidth capacitor *C40* is connected across the

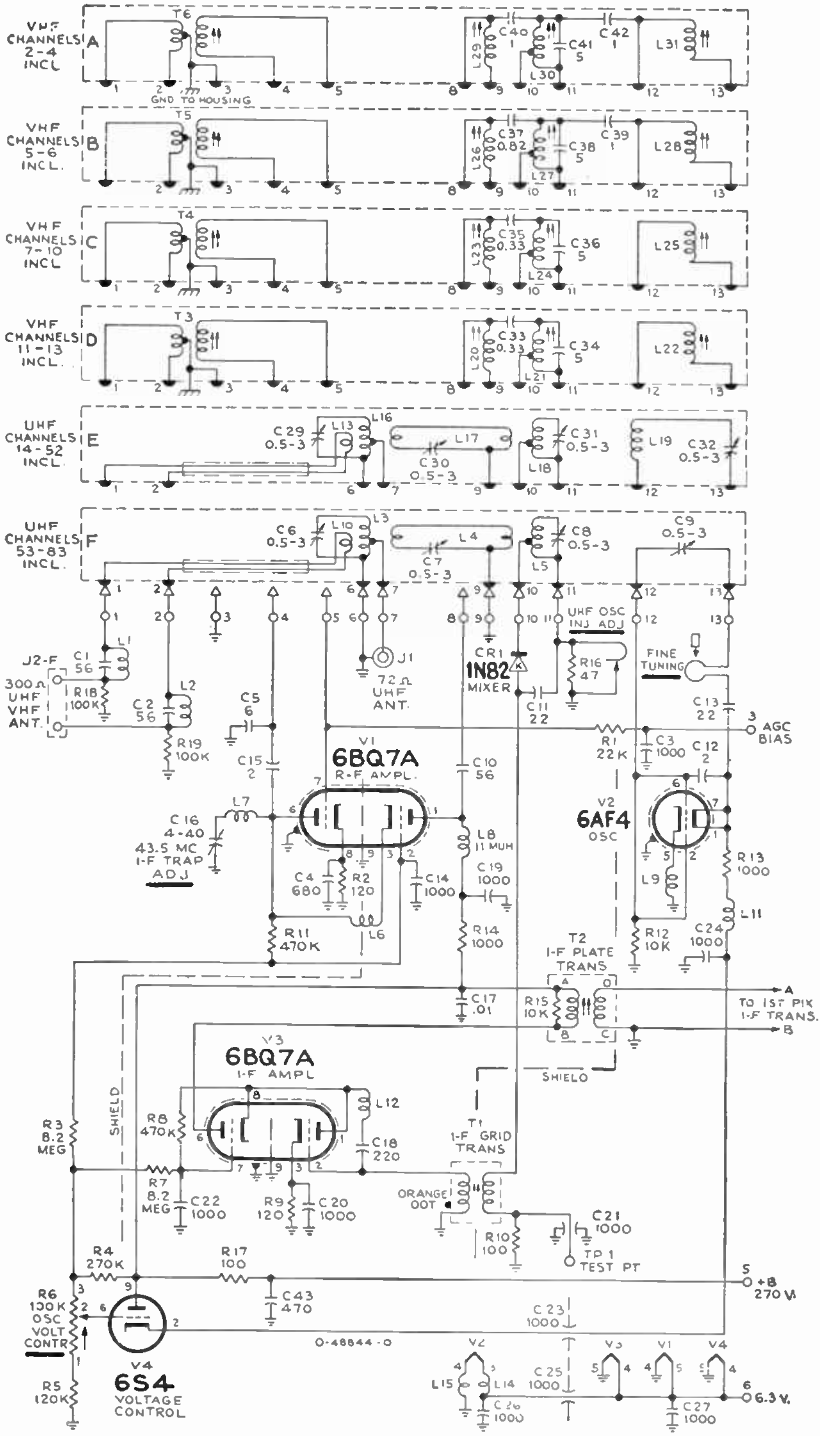


Fig. 210b RCA VHF-UHF Tuner

high-impedance side of the winding and permits a broad flat bandwidth with a sharp selective skirt for the interstage response curve. Local oscillator injection to the crystal is made at the low-impedance side of the resonant secondary via the inductive loop and resistor *R16*. Additional injection voltage on the low-band VHF channels is obtained with capacitors *C42* and *C39* on the insert strip. The local oscillator is an ultra-audion using a *6AF4* and turret-switched inductors. These inductors are placed across the oscillator tank circuit and have the proper value to resonate on the desired channels.

The mixer crystal output is applied through transformer *T1* to the 45-megacycle i-f amplifier tube V3—also a cascode amplifier stage that increases the level of the difference frequency and conveys it via transformer *T2* to the video i-f strip. An alignment test point is provided at the low side of transformer *T1* primary to permit application of scope or to measure crystal current. A-G-C bias is applied to the grid of the input cascode stage (pin 7) and also changes the cathode bias of the second section because of the direct-coupled voltage drop across resistor *R11*. Plate voltage is applied to the plate of the second section by way of the decoupling circuit and choke *L8*.

On a UHF position of the turret, the input r-f stage *V1* does not function (terminals 3, 4, 5, and 8 are idle). Instead (if the same antenna is used for UHF and VHF reception), the UHF signal is transformer-coupled via terminals *1* and *2* to the primary of the double-tuned UHF input transformer on each insert. If a separate UHF antenna is used, the signal is attached via the UHF antenna jack and is applied across the low-impedance section of the primary resonant circuit. The two windings of the transformer are separated physically and are coupled via a link and its bandwidth-adjustment capacitor to permit peak off-frequency rejection and maximum transfer with proper bandwidth for the UHF channel to be received.

The input transformer secondary connects to the crystal mixer. The injection voltage to the crystal mixer can be regulated to operate the crystal with peak sensitivity and best noise factor for reception of the UHF signal. On the much higher UHF frequencies, small capacitors are switched across the local oscillator tank circuit to form series-resonant tank circuits at the ultra-high frequencies to be generated for local oscillator mixing. The output frequency of the crystal mixer is again in the 45-megacycle i-f range, the remainder of the circuit functioning as per VHF reception. Thus, in this type of UHF conversion, a single mixer is used instead of the usual double-conversion methods (refer to Chapter 15).

The tuner uses a voltage-regulator tube V4 to set local oscillator plate current constant at a level that provides optimum injection voltage for the most sensitive biasing of the crystal mixer over the reception range. The crystal current can be adjusted for optimum UHF sensitivity with the loop injection control. Proper setting of the local oscillator at the specified plate current permits the most stable operation of the tube and the very least drift, a difficult problem at the very high UHF frequencies.

I-F SYSTEM

The output of the i-f amplifier on the tuner is coupled through a tunable link to the input resonant circuit of the first i-f stage, tube V106. This double-tuned transformer permits the proper bandwidth adjustment for reception of both picture and sound signals into the i-f amplifier. An adjacent-channel picture-absorption trap is a part of this first transformer. The first and second i-f amplifier stages are also coupled with a double-tuned transformer having two series-networks acting as mutual coupling elements—the primary and secondary windings being mounted in separate shields. A series-resonant trap forms a part of the secondary and tunes out adjacent-channel sound interference while a series 41¼-megacycle trap is used to set the associated sound-carrier frequency at the proper amplitude level on the over-all video i-f response. This adjustment permits precise setting of the sound carrier and minimizes intercarrier buzz and other disturbances. Such a bandpass *T*-type of transformer is helpful in making critical adjustments for optimum bandwidth for the over-all i-f response. Both grids are biased by the a-g-c system through suitable decoupling network.

Three stagger-tuned i-f transformers follow after the second i-f stage. For circuit simplicity and efficient performance, bi-filar windings are used, providing the maximum transfer of signal at the desired frequencies and good rejection of off-frequency interference. Each transformer is shunted by a small-value resistor to obtain satisfactory bandwidth, and each successive i-f stage has a higher cathode bias than that of the stage preceding to match the increased signal level as the signal is amplified stage by stage. The bi-filar windings permit the grid to operate at zero d-c potential and thus make it less subject to biasing by spurious impulse noises, improving the ability of the amplifier to follow the video information accurately.

The fourth picture i-f stage excites the crystal mixer connected in the secondary of the last bi-filar transformer. The crystal circuit is completely shielded together with the sound take-off transformer to minimize harmonic radiation and possible interference in the VHF range. This type of interference is a possibility, because even the low-order harmonics of the higher i-f frequencies fall into the VHF band. The output of the crystal through a suitable low-pass filter applies the video signal to the grid of the video amplifier. A series-resonant sound-trap is connected from the high side of the crystal output to ground and prevents the presence of sound interference in the video signal. Likewise, it acts as a convenient sound take-off, because a strong sound component can be developed across the inductor of the series-resonant circuit. Sound is taken off at tap *A* and applied to the grid of the first sound i-f stage.

SOUND CHANNEL

A two-stage 4.5-megacycle i-f amplifier is employed with the grid circuit of the second stage acting as an amplitude limiter because of grid current

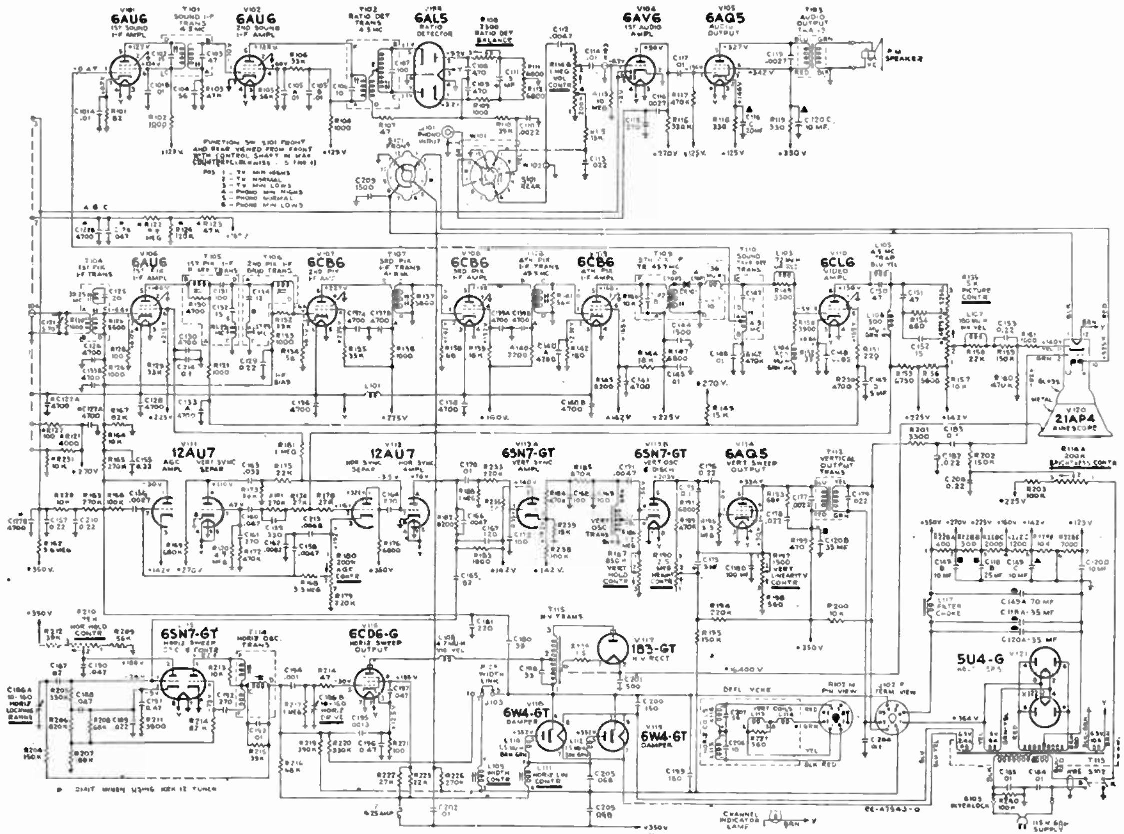


FIG. 210c RCA Television Chassis (Larger schematic reproduced on rear end papers)

biasing by network *C104* and resistor *R103*. The two-stage sound amplifier and its sharply resonant circuits minimize picture interference (intercarrier buzz) in the sound section. Limiter output is applied to a ratio detector which in itself is capable of amplitude limiting (particularly when well balanced), permitting a strong sound-output free of signal interference and intercarrier buzz. Link-coupling into the ratio detector permits best off-frequency rejection and least loading of the limiter output. The audio signal is applied through the de-emphasis network (resistor *R110* and capacitor *C110*) to the input stage of the two-stage audio amplifier which has both a tone-control system and a phono input facility.

VIDEO AMPLIFIER

The video signal, through a series-shunt peaking system, is applied to the grid of a single video amplifier using the new 6CL6 miniature tube. This single stage provides suitable amplification of the video detector output for proper drive of the large-screen 21-inch picture tube. The plate output through a suitable peaking system is direct-coupled to the cathode of the picture tube. Thus a negative-going sync signal has been applied to the grid of the video amplifier, and white compression is eliminated. The picture-contrast control is an elaborate divider system at the output of the video amplifier. It is a resistor-capacitor divider that permits signal-level adjustment without influencing frequency response. At the same time, the proper d-c bias to the picture-tube cathode is also adjusted as the contrast control is varied from one level of signal to another. In addition, the bleeder networks from the divider to plus-B and from the plate circuit of the video amplifier to plus-B permit the proper change in picture-tube cathode bias as a function of the average brightness of the picture. By suitable choice of bleeder values, the changes in d-c plate current of the video amplifier (varies with average brightness of picture because the grid circuit is direct-coupled to video detector output) cause a more natural change in the average brightness of the reproduced picture.

SYNC AND INTER-SYNC SEPARATORS

The composite video signal with sync positive is taken from plate resistor *R153* and applied to the two grids of the horizontal sync separator and vertical sync separator. The individual input stages allow time constants to be chosen in such a way as to emphasize the frequency components and repetition rates of the desired pulses and to reject the improper ones as well as impulse noises. The input to the horizontal sync separator has a short time-constant differentiating circuit that emphasizes the sharp edges of the horizontal timing pulses, while a longer time-constant integrating input circuit is used for the vertical in order to reject fast-rate leading edges and to emphasize the longer-duration vertical pulses. The separators are signal-biased and conduct only during the sync tip—biased by a charge on capacitor *C160* and by cathode bias on capacitors *C213* and *C162*. Both sync separators supply signal to the grid of the second section of tube V112. This stage amplifies both horizontal and vertical components, supplying vertical to an additional vertical sync amplifier and horizontal information to the syncro-guide control tube.

A-G-C bias is developed in the cathode circuit of the horizontal separator, and the bias level is a function of the peak amplitude of the arriving horizontal sync pulses. This d-c bias component varies with fading and other changes in sync-tip level, causing the d-c bias applied to the grid of the a-g-c amplifier to vary with the incoming signal level. A sharp positive pulse from the horizontal output circuit, after suitable attenuation, is applied to the plate of the a-g-c amplifier and permits it to conduct only during the horizontal retrace period. Consequently, the a-g-c bias developed is strictly a function of the incoming sync level and does not vary with average video information or interference. A-G-C biasing operations, consequently, occur only during the sync pulse intervals and are largely unaffected by impulse noises. This prevents sharp impulse noises from taking over in the a-g-c system and biasing down the amplifier, thus causing picture-signal level to deteriorate. A-G-C bias is supplied to the first two i-f stages and to the r-f stage of the tuner.

When biasing activities occur only during the sync pulse a much shorter time constant can be used in the bias-developing circuit of the a-g-c system. This fast action allows the a-g-c system to be made to respond much more quickly to fast changes in arriving signal level (aircraft flutter, for example). Thus the time constant in the cathode circuit of the horizontal sync separator contains rather small-value capacitors, and the a-g-c system is able to respond more rapidly to changes in sync amplitude at the grid of the sync separator.

VERTICAL DEFLECTION SYSTEM

Vertical information is applied through capacitor *C170,* the integrating network resistor *R233,* and capacitor *C172* to the grid of the cathode-follower vertical sync output tube. This stage supplies low-impedance signal to the dual-section integrating network that supplies sync signal to the vertical blocking oscillator tube. The cathode-follower output circuit also provides a return to plus B for the blocking oscillator and permits a sharp discharge curve that is less subject to erratic firing by interfering pulses or noise. The blocking tube employs an auto-transformer with a cathode circuit feedback. Sawtooth is developed in the plate circuit—capacitor *C173* charging during the cutoff period of the tube. Resistor *R191* peaks the sawtooth wave suitably for proper excitation of the standard vertical output tube. The vertical output circuit is suitably damped by capacitors across the transformer and by resistor *R227* across a portion of the vertical deflection coil. A cathode-bias type of vertical linearity system is employed. The secondary of the vertical output transformer also supplies signal to a resistor-capacitor network that integrates and develops a pulse of negative polarity which is applied to the grid of the picture tube to blank out the retrace lines during the vertical retrace period. Pulse is obtained by integrating the vertical output of the transformer with resistor *R201* and capacitor *C182.*

HORIZONTAL DEFLECTION SYSTEM

The horizontal deflection system consists of a conventional syncro-guide horizontal oscillator and of a control tube, which develops the proper drive sawtooth for the horizontal output tube across the drive capacitor *C186B*. The output of the horizontal sweep tube supplies excitation to the direct-drive auto-transformer that develops the sawtooth of current for the horizontal deflection coils. It also develops at the auto-transformer the high-voltage retrace spike that is applied to the plate of the high-voltage rectifier to develop the 16,400-volt anode potential. A dual-tube damper is used to remove retrace oscillations and to develop a boost voltage across capacitor *C203* that adds to the supply voltage and permits application of a higher voltage to the sweep output tube and sawtooth-forming section of the horizontal control tube. A width link is used where it is not possible to obtain adequate horizontal drive or where line voltage is down. When this link is connected between terminals *1* and *2*, the width control is removed from across the deflection coils and a higher supply voltage is applied to the plate of the horizontal output tube.

114a. *Large Picture Methods*

ENLARGING LENSES

An economical approach to a larger picture for owners of small-screen television receivers has been the enlarging lenses, which simply magnify the image which appears on the small picture-tube screen. These prove satisfactory because of the higher brilliance of the present-day picture tube, particularly the aluminum-backed picture tube which can be adjusted for higher illumination to compensate for light loss in enlarging lens. The disadvantage of the enlarging lens is, of course, its limitation on the field of view and, for best presentation, the picture must be viewed directly in front of the large lens. The lens, of course, must be optically good to prevent distortion or eventual discoloration. Magnification of

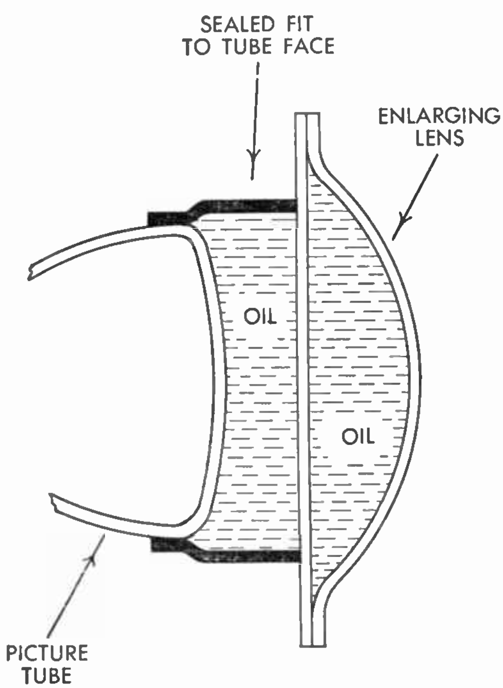


FIG. 210d Lens of Liquid Lens Corporation

the usual lens is approximately 1½ or somewhat less, doubling the useful picture area.

Most of the new lenses are made of plastic instead of glass, consisting of clear plastic sheets treated with a clear mineral oil. One unusual development has been the liquid lens which permits wider angle viewing, less weight, and, depending on type of mounting, an absence of air space between the picture tube and lens. Because of the differing transmission characteristics of air and glass there is blurring and diffusion of reproduced image by many lenses. In the liquid lens (Fig. 210d) the collar fits over the base of the tube and clear oil is present between the face of the tube and the lens. This oil has the same transmission characteristics or refractive index as the glass or plastic, the combination effectively removing two refractive surfaces to prevent distortion of the image.

DIRECT PROJECTION OR REFRACTIVE SYSTEMS

The direct approach to a larger picture by means of projection is the use of a refractive system such as employed for movie projectors. However, in contrast to the abundance of light available in a movie projector the average picture-tube screen is larger in area and lacks light intensity. If sufficient brilliance is to be obtained the system (Fig. 211) must be carefully designed to overcome the very low efficiency of the lens system.

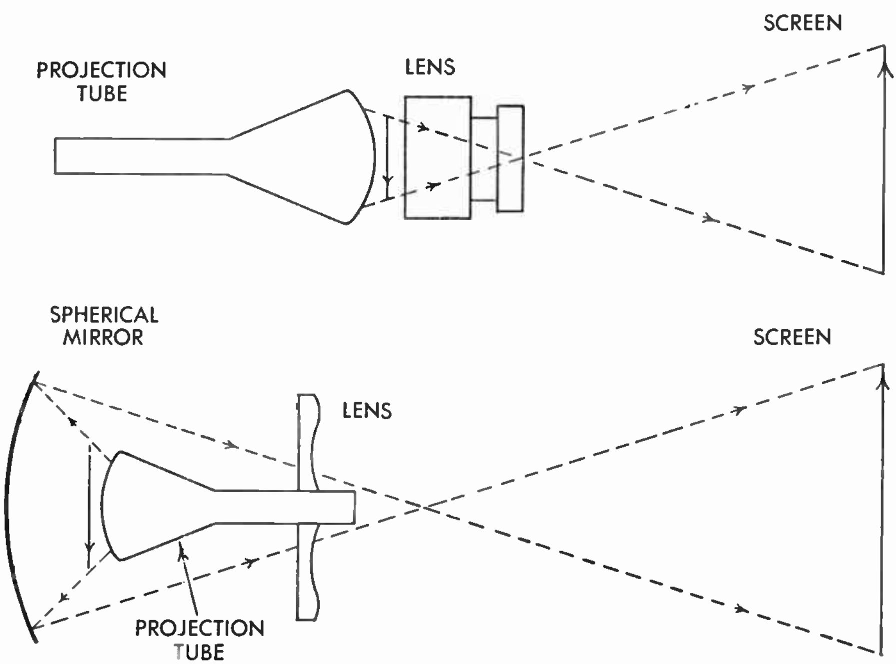


FIG. 211 Direct and Reflective Projection Systems

The advantages of the direct projection system as compared to a reflective system are simplicity of optical mounting and ability to obtain a fairly large range of screen sizes to meet specific needs.

REFLECTIVE PROJECTION SYSTEMS

Reflective projection systems (drawing B, Fig. 211) employ a spherical mirror to magnify the presentation on the screen of a small-projection picture tube. To prevent distortion a correction lens must also be used to bring the image to sharp focus on the viewing screen. The advantage of the reflective projection system is more complete utilization of the light available from the screen of the picture tube, although the system, per given design, is limited to a fixed-viewing screen dimension. Although its efficiency is better, its design and production tolerances are more severe.

115. *Light and Basic Optics*

Light is measured in candlepower or foot-lambert. Candle power is measured on the basis of a standard candle which, under a certain prescribed set of fixed conditions, delivers a definite amount of light. In application, when an incandescent lamp is said to have 60 candlepower it means its illumination is sixty times greater than that of a standard candle.

A foot-candle is the intensity of illumination which falls upon a surface at a point which is 1 foot distant from a source of one candlepower, the surface being perpendicular to light rays at that point (Fig. 212). The foot-candle, therefore, is a measure of light intensity while another term, the lumen, measures quantity of light. One lumen is the amount of light falling on a surface which has an area of 1 square foot, every point of this surface being 1 foot from the source of light of one candle. Radiation of light from a point source in space emanates in a sphere and, therefore, as the light spans outwardly the density of the light or the lumens (illumination per square foot) decrease. Inasmuch as the expansion is spherical the illumination decreases as the square of the distance from the light source. If there is a lumen per square foot of light available 1 foot from the light source, 2 feet away from the source the amount of the light will have decreased to ¼ lumen per square foot. The intensity of the light at any point 2 feet away from the light source, therefore, is ¼ foot-candle, and it can be said that intensity of illumination at a point varies inversely as the square of its distance from a point source of light.

The amount a point is illuminated, therefore, is dependent on the flux density of the beam which strikes it (number of foot-candles). An object so illuminated may reflect, transmit, or absorb light which falls upon it. If the object reflects or transmits light (a refractive lens transmits light and a spherical mirror reflects light), it in turn becomes a light source, called a secondary source. Candlepower of the secondary source is measured in candles per square foot or in lamberts (1 foot-lambert equals 1 foot-candle reflected or emitted). The illumination of the primary source of a projection system (fluorescent screen of picture tube) is given in foot-candles or foot-lamberts and illumination at the secondary source (viewing screen) is given in foot-lamberts. Thus, if a projection system is said to have a primary intensity of 1,000 foot-lamberts and the brightness of the secondary image on the viewing screen is 50 foot-lamberts with a magnification of 6 (viewing screen six times the area of the picture-tube screen) over-all efficiency of the projection system is approximately 30 per cent (6 times 50 divided by 1,000). Magnifi-

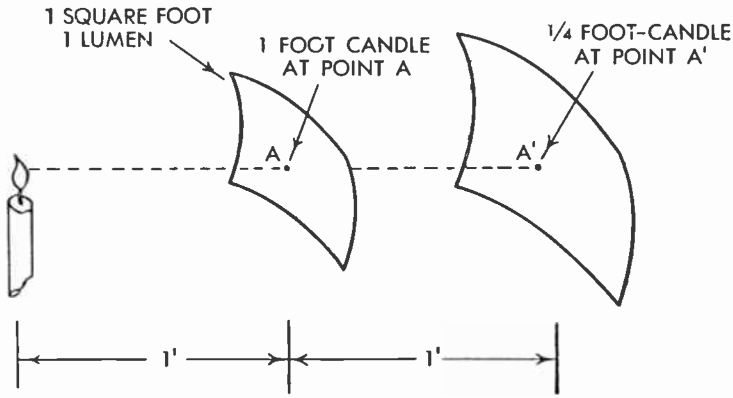


FIG. 212 Measure of Illumination

cation, of course, has to be considered in determining efficiency of the projection system and, although illumination varies as the square of the distance from the source, the reproduced image is six times larger than the original and light flux is distributed over a larger area. Efficiency of the system, however, is a measure of how much of the total flux leaving the projection tube screen reaches the viewing screen. In other words, if the viewing screen were reduced to the same size and area as the picture-tube screen the quantity of light arriving would be 300 foot-lamberts, demonstrating an efficiency of 30 per cent. Nevertheless it is apparent and, of course, to be expected, that as the viewing screen is made larger (longer light path between light source and image) apparent illumination of the screen decreases.

REFLECTION OF LIGHT

When a beam of light strikes a polished or silvered surface nearly all of the light is reflected while the same light striking clear glass would be largely transmitted through it. In a direct projection system using lenses, the light is transmitted through the lens system while with a reflective projection system the light is reflected off the silvered surface of a spherical mirror.

If light is directed toward a reflecting surface and arrives at a point on that surface at an angle, the light ray is reflected from this point, as shown in

Fig. 213. If a line, called a *normal,* is drawn perpendicular to this point source, the angle it makes with the arriving ray is the same as the angle it sets off with respect to the reflected ray. Therefore, these two rules apply to reflected waves: (1) incident ray and reflected ray form equal angles with respect to a

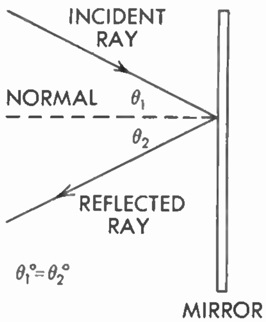


FIG. 213 Reflection from Mirror

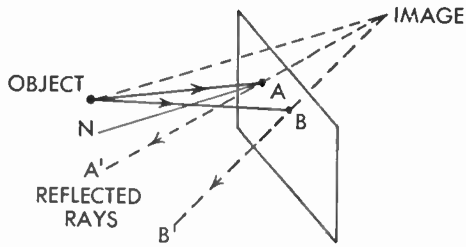


FIG. 214 Image in Mirror

perpendicular drawn to the surface at that point where the ray strikes; (2) angle of incidence is equal to angle of reflection.

A source of light, such as the object point in Fig. 214, when positioned in front of a mirror radiates a bundle of light rays which leave the light source and strike the mirror at many points. Rays strike representative points A and B, and from these points, a reflected ray will leave toward points A' and B'. The angle of departure of these rays is the same as the angle of the incident rays. Now the ray paths, if extended supposedly beyond the mirror, will converge at an image point seemingly within the mirror. Thus, when looking into the mirror an image of the light source is apparent. Likewise, an image of any other light source or object is also imaged in like manner in the mirror.

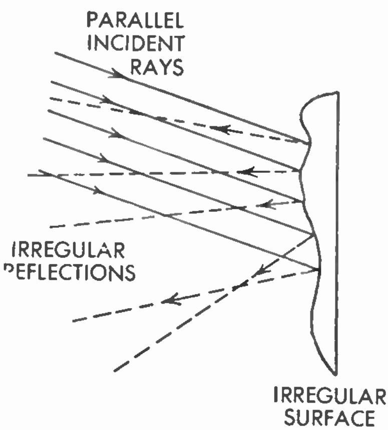


FIG. 215 Reflection from Irregular Surfaces

If a surface is irregular instead of flat, the reflected rays will not leave in a regular pattern but will scatter off the surface in accordance with the angle of arrival of the rays and the angle of the surface at which individual rays are arriving (Fig. 215). When a group of parallel rays strike an irregular surface, rays are reflected from that surface with the angle of each reflected ray always equaling the angle of its associated incident ray as compared to the angle of the surface at the point light ray arrives. It is possible, therefore, with a controlled surface to concentrate reflected light at prescribed vertical and horizontal angles. Thus, it is possible with a directional screen to distribute light emanating from it to a confined area from which seated or standing persons view it comfortably.

REFLECTION OFF A SPHERICAL MIRROR

It is possible by means of a spherical mirror to obtain a magnified image, light rays all coming to focus in bundles at a prescribed distance from the mirror. For example, in drawing *A*, Fig. 216, the light rays leaving point *O* are reflected off the spherical surface and converge again at a point *I*, some distance behind object. Or if a number of point sources are taken, second drawing, along the object (*1*, *2*, and *3*) the ray bundles from each source all converge along the image line at points *1*, *2*, and *3*. If the object is in form of

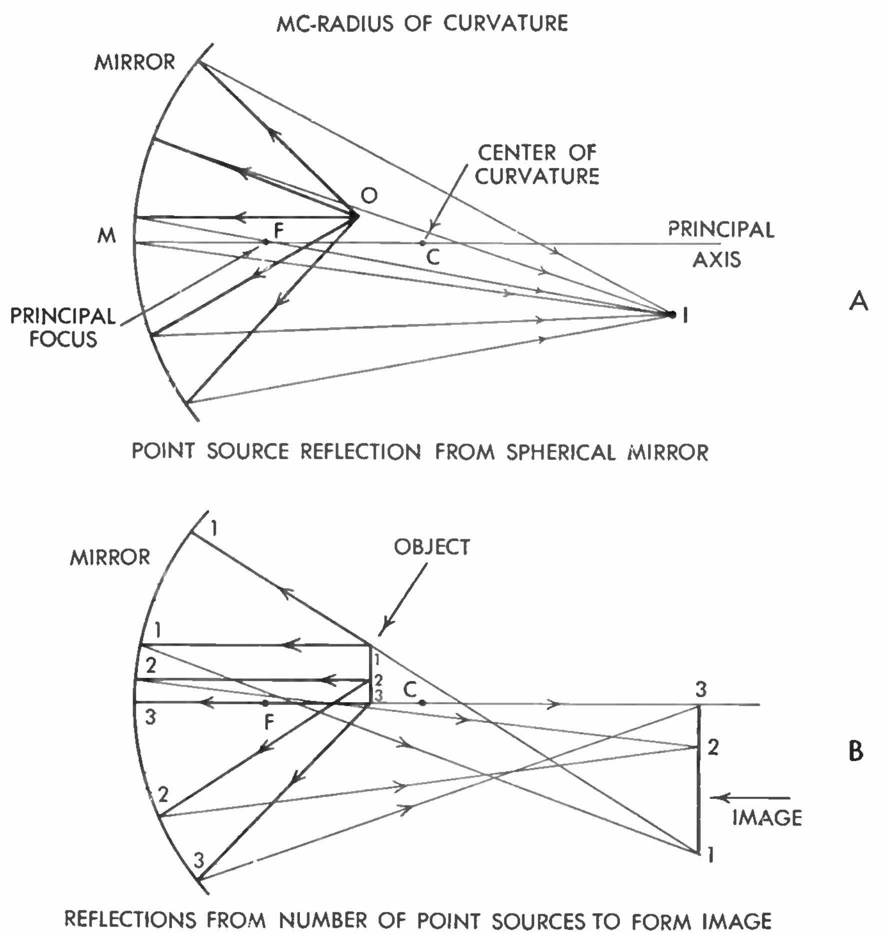


FIG. 216 Reflections from Spherical Mirror

a near plane as for television projection systems, the light sources will come to focus in a plane. Also, notice there is magnification of the object, and image is also inverted, as might be expected, in the same manner as image is inverted in a conventional lens system.

Point C is known as *the center of curvature of the mirror* and is the center point of an imaginary complete sphere, actual spherical mirror forming a section of that sphere. The distance from this center of curvature is the same to all points of the mirror and that distance is known as the *radius of curvature.* The mid-point M of the mirror is known as the *vertex* and a line drawn through the center of curvature from the vertex is the *principal axis* of the mirror. Point F, midway between the center of curvature and the mirror along the principal axis, is known as the *principal focus,* and it is a point through which all reflected rays would pass if a series of parallel rays parallel to principal axis strike the mirror.

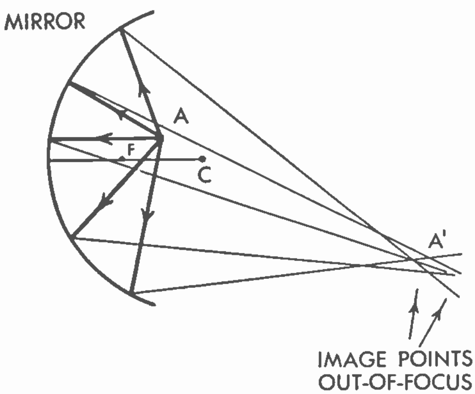


FIG. 217 Spherical Aberration

If spherical mirror represents an appreciable section of the imaginary sphere with its center of curvature at C (Fig. 217), rays emanating from the point source do not come to focus after reflection at exactly the same point. Reflected rays from outer section of spherical mirror do not pass through same image point. This defocusing action is called *spherical aberration.* If all reflected waves are to come to a sharp focus at the same point, a special correction arrangement is necessary. In a television projection system spherical mirror is made as large as possible with respect to the object (screen of projection picture tube) in order to reflect as much light as possible and, therefore, increase the efficiency of the system. However, as the size of the spherical mirror is increased, spherical aberration also increases, and it is necessary to use a correction lens if the object is to be brought to sharp focus. Likewise, the radius of curvature is held down because this means more complete utilization of the light from the object and a greater magnification for a given distance between image and spherical mirror. This system in which the spherical mirror represents an appreciable section of the sphere represented by a specific radius of curvature is known as a *wide-aperture reflection system* because it utilizes an appreciable amount of the light from the object source. In like manner a lens which transmits substantial light from the object is known as a *wide-aperture lens.* It is important to understand that in a wide-aperture reflection system it is necessary to correct for spherical aberration to bring the light rays converging from wide angles into focus at one point.

The distance between the principal focus point F and the mirror, along the principal axis through the mirror, is the focal length of the spherical mirror. Focal length depends on the radius of curvature because the principal focus exists midway between center of curvature and mirror. The distance between mirror and object point A is known as the *object distance,* and the distance between the mirror and the image point is known as the *image distance*. The quotient of image distance divided by the object distance represents the magnification of the reflection system. To obtain magnification it is important to observe that the object must be placed at some point between the center of curvature and the principal focus, these two positions representing two extremes; namely, if the object is positioned at the center of curvature the image will also appear at the same point and the magnification is unity. However, if the object is located at the principal focus it will be reflected to infinity. The object must be positioned properly between center of curvature and principal focus to obtain the desired magnification at a practical object distance. At the same time the radius of curvature of the spherical mirror is properly chosen to obtain satisfactory brightness because, as the object distance increases, magnification and image size increase while brightness decreases.

In summation, a spherical-mirror system used as a reflective projection system is dependent on the following requirements:

1. It must be a wide-aperture system to fully utilize the limits set by the fixed amount of light available from the fluorescent screen and the required viewing-screen brightness for the television projection system.
2. A wide-aperture system requires correction to bring the object into clean focus at the image point.
3. A substantial amount of magnification is desired for a practical image distance and without undue sacrifice of brightness.

REFRACTION OF LIGHT

When a light ray arrives at a boundary between two different mediums (such as air and glass), a light ray does not continue on its straight path but is bent in accordance with the differing velocity of light travel in the new medium as compared to the old. For example, in drawing A, Fig. 218, rays are traveling at a certain velocity into a second medium which slows down the velocity of light travel. Inasmuch as the light rays are crossing the boundary between the two mediums at an angle, side A of the light rays will enter the new medium first and, therefore, will be slowed down while the B section, which has not as yet reached the boundary, is still traveling at its initial velocity. Consequently, the front of the light ray is bent and actually travels through the new medium at a new angle. In drawing B the light ray is traveling with a certain velocity to a new medium through which it can travel at a higher velocity; therefore, as section A enters the new medium it is speeded up, while side B still travels at the old velocity until it reaches the boundary between the two mediums. Consequently, the entire front of the light ray is bent away from the so-called "normal" (line XY). Thus, in traveling from a higher to a lower velocity the beam is bent toward the normal while it is bent away from the normal going from a low to a higher velocity medium.

It is apparent, therefore, that if a light ray goes through a medium which has its two sides parallel (Fig. 219) the light will enter one side and be reflected away from the normal, lower to higher velocity. Upon leaving the

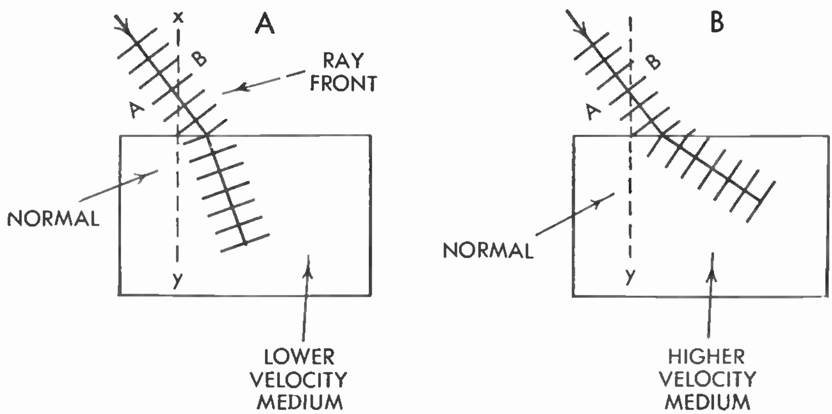


FIG. 218 Refraction of Rays Traveling between Two Media

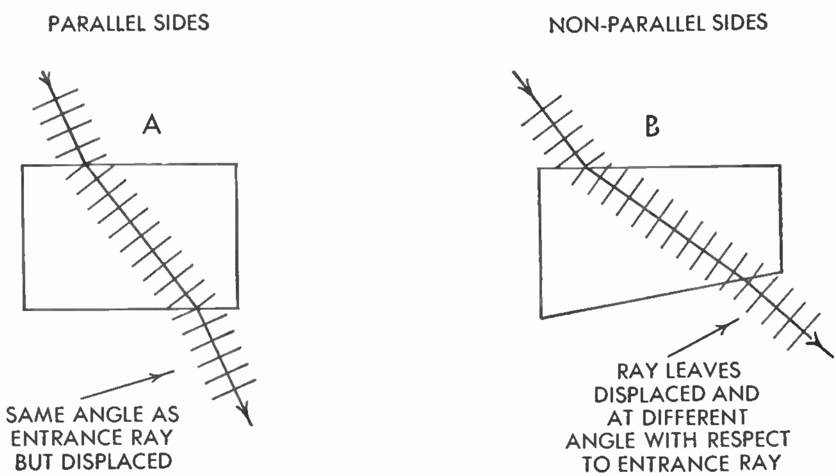


FIG. 219 Rays Leaving Parallel and Nonparallel Exit Sides of Different Media

other side it will, of course, be moving from a higher to a lower velocity medium and it will once again be reflected back to the same angle at which it originally arrived at the input side. However, the output rays will be displaced to the right with respect to the input rays.

It is also possible to have the light rays leave the medium at a different angle by changing the shape of the exit side. For example, in drawing B the ray leaves the new medium at approximately the same angle it is traveling through and, therefore, different with respect to the angle at which it arrived at the input side. This refraction of light rays, as they pass between two mediums, is the basis of operation of the conventional lens. The lens serves as a method of transmitting light and becomes, therefore, a secondary light source, transmitting a substantial portion of the light arriving at the input side of the lens. In order to control the angle at which the light rays leave the lens, the shape of the lens can have a number of different figurations according to application.

LENSES AND LENS SYSTEMS

A refractive optical system using lenses, although not as efficient as a reflective system, still has advantages when abundant light is available because of its ability to focus at various distances over a given range by simple adjustments. In addition, its alignment and physical construction are not nearly so critical as a reflective system and its closer tolerances.

Two basic lenses, biconcave and biconvex, are shown in Fig. 220. The concave lens is called a *diverging lens* because light which enters it is scattered and fans out when leaving the exit side, while light entering the biconvex lens converges into a focus point. This point at which all the emanating rays converge from a parallel beam of light entering the entrance side is called the *focal point*, and the distance between this point and the center of the lens is known as the *focal length*. Speed of a lens or its aperture represents the amount of light conveyed from an object through a lens to form an image. The greater the diameter of the lens the shorter the focal length of the lens, the less magnification, and therefore the greater the image illumination. The focal length of a lens is shortened by increasing the thickness of that lens and, therefore, a wide-aperture system is a thick lens of substantial diameter and consequent short focal length.

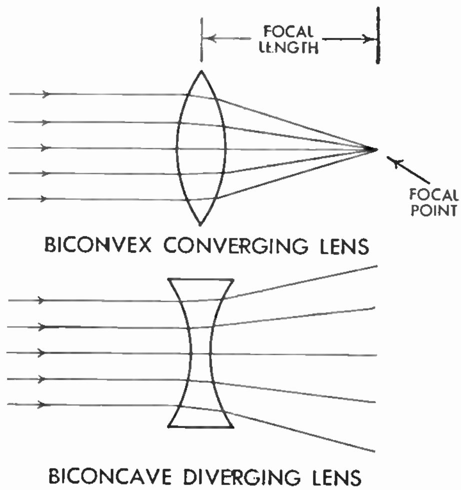


FIG. 220 Basic Lenses

The lens speed or aperture equals focal length divided by diameter of lens: $f = F/d$.

A convex lens can be used as a magnifying lens producing an inverted image, as shown in Fig. 221. If the object is placed between center of curvature and focal point of the lens, magnified image is formed on other side of lens. Convex lenses suffer from two defects, namely, spherical aberration and chromatic aberration (drawings A and B, Fig. 222). Spherical aberration, particularly in a thick lens (short focal length and high speed), causes the marginal rays to converge at a different point than other light rays which enter the lens. Likewise each color which makes up light has a different wavelength and velocity and, therefore, will be refracted at a slightly different angle when

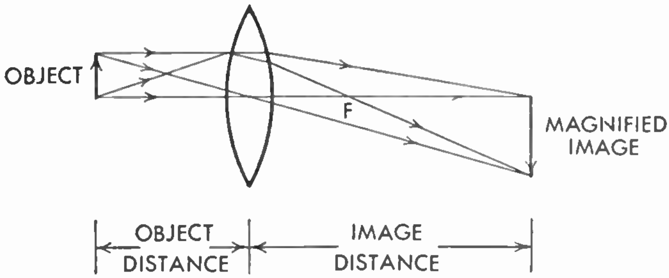


FIG. 221 Object Inversion with a Lens

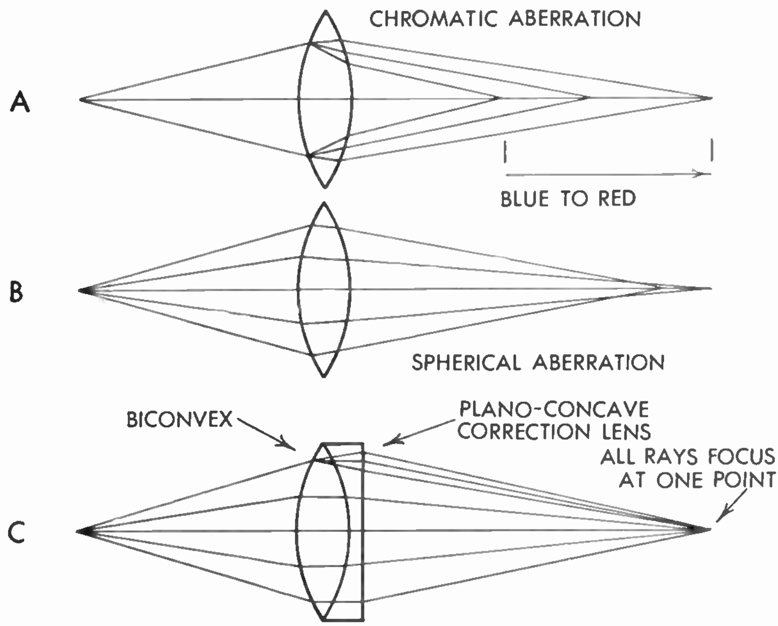


FIG. 222 Correction of Spherical and Chromatic Aberrations

passing through the lens, so the various colors will not come to focus at the same point. Both chromatic and spherical aberrations can be corrected to a limited extent with a so-called "doublet lens," drawing C, which consists of a biconvex lens and a plano-concave surface which is being used as a correction lens. This spherical surface along with its different refraction angle causes the light rays to converge on a single point. Thus an aspherical correction lens as used in a projection system consists of a very thin lens positioned a short distance away from the spherical mirror, correcting for its spherical aberration.

116. *Commercial Projection Systems*

A refractive system (Fig. 223) is used in a direct projection system. A special multiple lens with an f 1.9 aperture and a focal length of 5 inches is used. This lens is corrected for spherical and chromatic aberrations. The magnification of system is between 5 and 7½, variable over fairly wide limits by simple movement of the lens position.

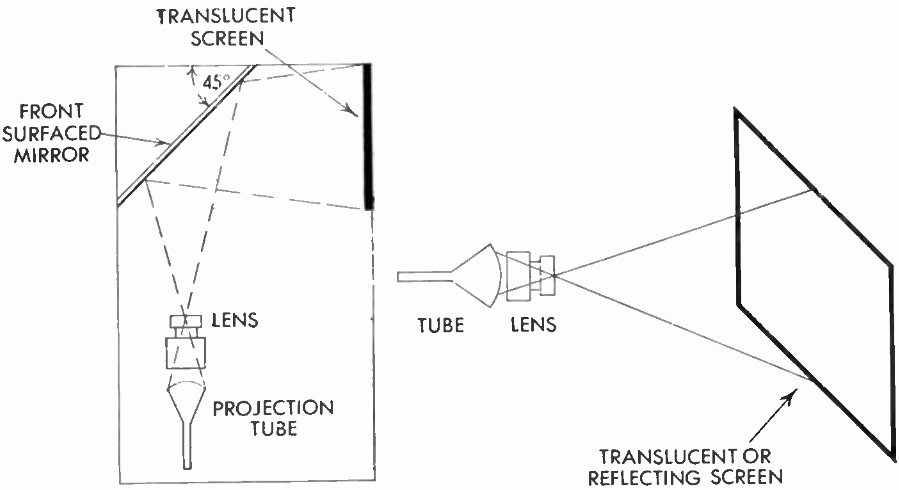


FIG. 223 Direct Projection

The lens is focused on a front-surfaced mirror which reflects the light through a transmissive screen. A front-surfaced mirror is used instead of a silver-backed glass mirror to reduce the double-reflection distortion usually produced by the glass of the mirror. If light rays are reflected immediately by a front-silvered surface, no diffusion and reflection is introduced. A 5-inch projection tube with a 27-kilovolt anode supply produces a high-light brightness on the viewing screen of about 17 foot-lamberts. There are two screen sizes—15 by 20 inches and 22½ by 30 inches.

REFLECTIVE SYSTEMS

There are three basic reflective systems, all of which use a so-called "Schmidt optical system." In such a system (Fig. 224) a picture tube is mounted at the center of curvature of a spherical mirror and the light emanating from it is reflected on the spherical mirror. The reflected rays are brought to focus, after passing through an aspherical correcting lens on a viewing screen. A viewing screen can be a reflecting directional screen, as Philco uses, or light can first be reflected off a mirror and then through a translucent viewing screen as used by RCA. Still another type used by North American Philips (drawing C) uses an extremely small tube (only 2½ inches diameter) and a so-called "folded Schmidt system." In this system, light first reflects off spherical mirror to a 45-degree plain mirror, which reflects the light through a spherical corrective lens to a viewing screen. A folded Schmidt system of this type can be mounted in a small area as a packaged unit.

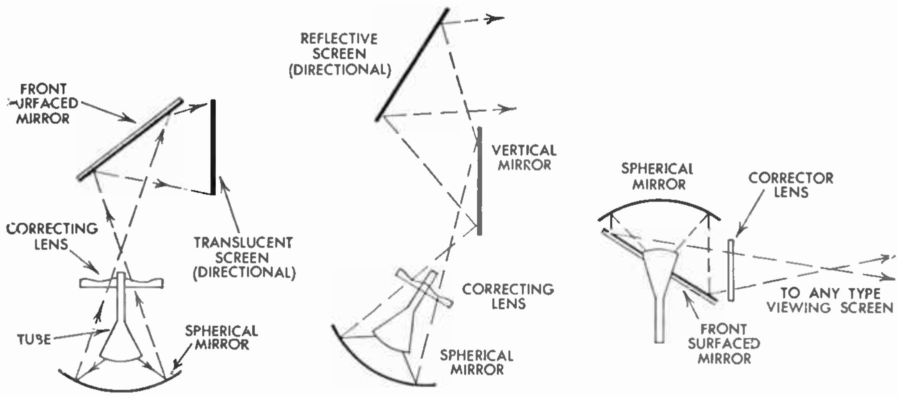


FIG. 224 Modified Schmidt Projection Systems for Television

117. *Philco Projection System*

Basic components and pertinent dimensions of the Philco projection system are shown in Fig. 225. Philco projection tubes have a 4-inch diameter and are mounted 6 inches in front of the mirror. Light from projection-tube screen is thrown on spherical mirror and then to the aspherical correction lens which is mounted at the center of the curvature of the spherical mirror. This center of curvature is approximately 11 inches from the spherical mirror, or the radius of curvature of the spherical reflection system is approximately 11 inches, with principal focus of mirror located 5½ inches down the principal axis. The projection-tube screen is about ½ inch from the principal focus and, therefore, its magnification will be substantially great but not infinity. In the case of the Philco system the magnification is approximately 6½ and therefore the image on the viewing screen is that many times larger than the scanning raster on the projection-tube screen.

To prevent multiple reflections between fluorescent screen of the projection tube and spherical mirror, a 4-inch-diameter section at the center of the spherical mirror is blacked out, and no light is reflected from its back to the object. Therefore, the contrast range of the object is not affected. Inasmuch as a spherical mirror would focus an object in the form of a rectangular plane as a curved field (edges of field nearer mirror than center), it is necessary that the fluorescent screen of the projection tube have proper curvature to bring entire object to focus at plane of viewing screen.

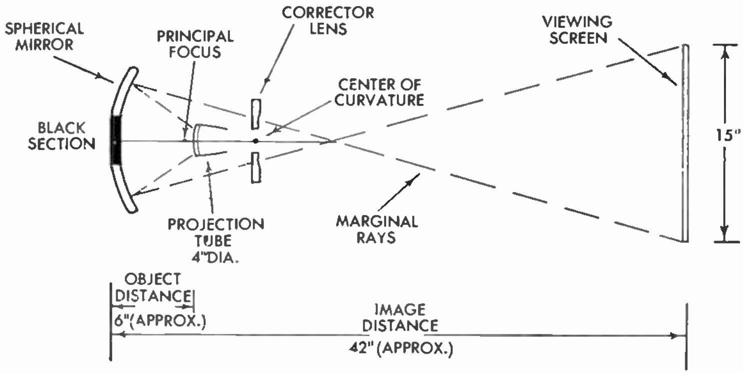


FIG. 225 Basic Plan of Philco Projection System

Proper shaping of spherical mirror and correction lens overcomes spherical aberration of the wide-aperture system (Fig. 226). The correction lens consists of a concave and convex section; concave surface bends light rays away from the normal while convex surface bends rays toward normal. As shown in Fig. 226, with ray *2* as a reference, observe that ray *3* in passing through the correction lens is bent away from the normal while ray *1* passes through the concave section and is bent toward the normal, all three rays intersecting at the image point. If the correction lens were not present the various rays would come to focus at different points along the principal axis. Of course, this same correction applies to all object points of the fluorescent screen of the projection tube and the proper ratio of curvature of the tube and curvature of the spherical mirror would cause all the object points to come to focus at the plane of viewing screen. Radius of curvature of the picture-tube screen has specific relation with radius of curvature of the spherical mirror for optimum performance at the finite distance between correction lens and viewing screen, or so-called "throw" of the projection system.

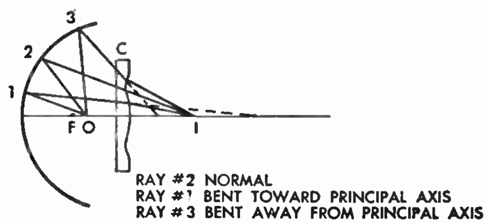


FIG. 226 Correction of Spherical Aberration by Corrector Lens at Center of Curvature

MECHANICAL LAYOUT OF PHILCO SYSTEM

The physical plan of the projection unit (Fig. 227) is arranged for ease of mounting, conventional cabinet size, and placement of screen at proper eye level for comfortable viewing. Thus, the main components of the projection unit, consisting of tube, spherical mirror, and correction lens, are mounted in a dustproof metallic case to prevent dust and injury to the lenses and consequent distortion and loss of brightness. The projection barrel is mounted at an angle of 45 degrees with respect to a front-surfaced mirror, which is mounted vertically in the front of the cabinet and there reflects the light from the projection barrel to a viewing screen. When the receiver is in operation, image is reflected from viewing screen at eye level, viewing screen being on underside of lid.

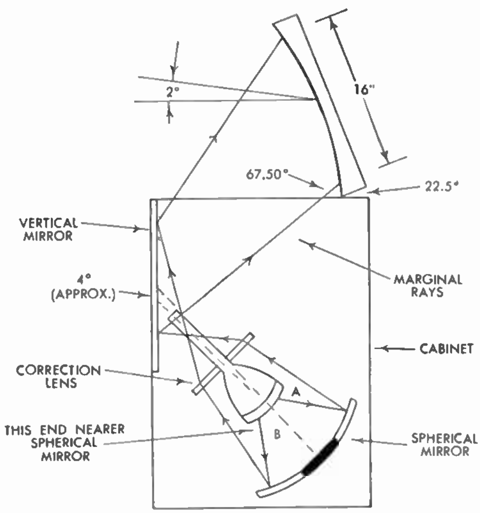


FIG. 227 Mechanical Layout Philco Projection System

Screen is tilted forward approximately 22½ degrees and light is emitted from it horizontally at eye level. There is a slight angle of approximately 1½ to 2 degrees with respect to eye level to permit comfortable viewing by persons standing in the rear of a large room or auditorium. Angle is also satisfactory for correct viewing by persons seated in front of the receiver.

One problem encountered with this type of folded Schmidt system is that the top portion of viewing screen is at a different distance from the top portion of image on the projection-tube screen than the bottom of the screen is from the bottom of the picture on the projection tube. Inasmuch as there is a difference in ray path length it would at first appear that either the bottom or top of image on viewing screen would be out of focus. However, this difficulty is circumvented by mounting the projection tube in the barrel slightly off center with respect to the optical center of the projection system (Fig. 227). This displacement of almost 4 degrees positions the bottoms of the projection-tube screen (top of picture) nearer the focal point of the spherical mirror than the top (bottom of picture), side *A*. As a result the portion of the image coming off side *B* (actual top of picture because of inversion by spherical mirror) reaches focal point over a greater distance to top of viewing screen than for the rays emitted at side *A*, which has the shorter path to the bottom of viewing screen. It is apparent, therefore, that both top and bottom of image on the screen are in focus; however, magnification at top is greater because of increased image distance. Consequently, if the scanning raster is rectangular, its reproduction on the viewing screen will be trapezoidal, sides tapering inward top to bottom.

To compensate for this keystone pattern, it is necessary to form a trapezoidal or keystoned raster (Fig. 228) on the fluorescent screen. Despite additional magnification of the image at top, its reduced size on the projection-tube screen produces a rectangular image on the viewing screen. One method of decreasing sweep width at top of image would be to generate a horizontal sweep which has an increasing amplitude as beam reproduces the picture on the scanning raster of the projection tube. As the beam scans from top to bottom of picture it would have an increasing sweep width. Another method in which to accomplish this effect as far as a television receiver is concerned is to put the scanning beam under the influence of fixed magnets mounted near to face of the projection tube.

Magnetic-pole pieces generate a magnetic field which passes from side to side of tube and, therefore, causes scanning beam to be deflected upward (motion at right angles to both forces). Thus the path length the beam travels from the electron gun to top of the tube is greater than the travel from the gun to bottom of tube. This increase in path length, of course, increases the sweep width at the top in comparison to the bottom in the same manner as increasing the distance between screen and electron gun also increases sweep width (deflection sensitivity). One disadvantage of this method is that the entire scanning raster is pulled unevenly and displaced upward; it is necessary to generate a compensating but weaker magnetic field which first causes the electrons to dip down, then up, increasing the path length and at the same time not causing displacement of the image vertically. The weaker magnetic field is generated by small magnets which for optimum performance must be mounted at a 45-degree angle with respect to axis of the projection tube (mounting angle determines strength of weaker magnetic field).

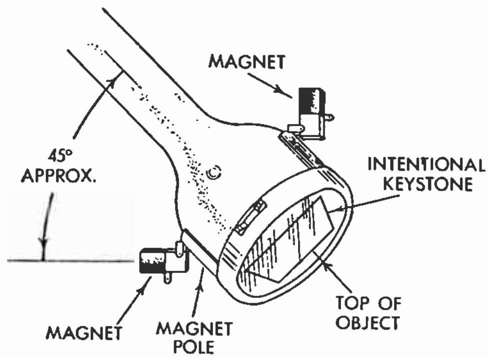


Fig. 228 Position of Keystone Magnets and Keystone Pattern

DIRECTIONAL VIEWING SCREEN

Directional viewing screen of the Philco projection receiver produces substantial illumination and contrast at viewing level and a reduced illumination at other angles. Concentration of the emanated light through a 60-degree horizontal angle and a 15-degree vertical angle results in amplification of approximately 15 with respect to the illumination (normal viewing angle) from a nondirectional screen. Another advantage of such a directional screen is the fact that light which originates in a brightly illuminated room, while it strikes the screen, is reflected at angles dominantly outside of the normal viewing angle. Thus reflections from other light sources do not affect contrast of the picture.

Viewing screen is a reflecting sheet which is a cylindrically concave (Fig. 227) partial cylinder with its axis horizontal. Many thousands of vertical grooves extend from left to right across the screen. A slightly concave screen, along with the tilt of the viewing screen and the position of the mirror (angle of reflection equal to angle of incidence), concentrates the illumination vertically. The vertical grooves spread the illumination over an angle approximately 60 degrees horizontally. Actually, the vertical grooves represent many small vertical cylindrical sections and, therefore, depending on where the light rays strike the screen, reflect a ray at the same angle as the arriving ones with respect to a normal to the surface. The normal is, of course, at various positions depending on what section on the vertical groove light strikes. The presence of the vertical groove (Fig. 229) causes the reflected rays to fan out horizontally.

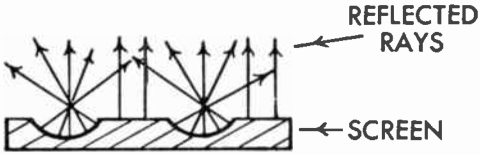


FIG. 229 Vertical Grooves Disperse Light Horizontally

We might conclude at first that if the center of the spherical mirror were blacked out to prevent multiple reflection between picture tube and mirror, portions of the image on the fluorescent screen of the picture tube would not reach the viewing screen. This is not the case, however, because each object point on the screen radiates light in a partial sphere to all sections of the spherical mirror; the ray energy from each object point strikes the center and ends of the spherical mirror. The presence of the dark spot at the center of the spherical mirror, therefore, does not reduce illumination for any one point of the screen, but does so for all points, lowering the transmission efficiency of the reflection system approximately 10 per cent.

Likewise, it would appear that the presence of the picture tube and associated deflection coils and mounting would also interfere and completely black out sections of the image. It must be remembered that light rays emanating from any one object point on the screen of the picture tube deliver light to all sections of the spherical mirror, and although some of the reflections coming off the mirror will not reach the viewing screen because of the presence of the tube and its associated components, most of the rays will pass through correction lens and onto screen. One defect, however, is apparent on the viewing screen because of the angle at which the rays are arriving at the central portion of the screen. The presence of the picture tube and associated components prevents arrival of light rays at the central portion of the viewing screen which have an angle nearly perpendicular to the viewing screen. However, rays which arrive at the central portion of the viewing screen are reflected at the same angle at which they arrive, and if they arrive at an angle substantially off the perpendicular they will also be reflected at the same angle. Consequently, if the viewer observes the picture when his eyes are physically level with the center of the screen, the central portion of the image will appear dimmer than the image on all other sides. If the viewer stands at any other position within the normal viewing angle, screen illumination will be absolutely uniform because there will be an abundance of light collected from both the central and outer sections of the screen. While you stand at the center level, however, there are no perpendicular light rays coming toward you from the central section of the screen. To circumvent this difficulty on the Philco screen, a thin, light special lacquer is sprayed over its surface to cause some diffusion of the light which permits some sharp light reflection from the central portion of the viewing screen. In addition, the defect is not nearly so noticeable from a directional screen because of the horizontal fanning action of the vertical ribs.

The special Philco projection tube, TP400A, requires a peak signal swing of some 80 volts and has a high-light brightness of 2,000 foot-lamberts. With an optical efficiency of 30 per cent, gain factor of screen an a magnification of approximately 6½, the projection system products a viewing-screen brightness of 100 foot-lamberts or better.

118. *RCA Projection System*

In the RCA projection system (Fig. 230) the projection tube, spherical mirror, and aspherical correcting lens are again used. The three components previous mentioned focus image through a 45-degree inclined mirror to a translucent viewing screen. In the RCA system, the center line of the projection tube is centered on the vertical axis of the projection system. It is not necessary to use an eccentric mounting for this type of projection system, because distance from any object point of the image on projection-tube screen to its corresponding point on the viewing screen is the same. The aspherical correction lens is made of molded plastic to lower the cost of production as compared to optical grinding methods used in manufacture of a glass correction lens.

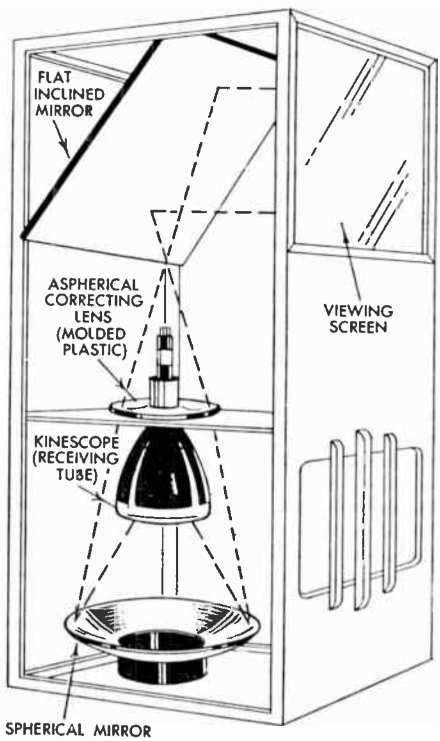


FIG. 230 Arrangement of Components of RCA Projection System

Specific features of a well-planned optical mounting assembly are: Optical alignment adjustments must be accessible and simple. The assembly must be dustproof to prevent loss of contrast and brightness by dust collection. It must be electrically shockproof to provide protection from the high voltage of the projection tube, permitting safe manipulation of the optical adjustments while viewing screen is observed. The optical barrel must be metallic to prevent radiation of the X-rays generated by the very high voltage cathode-ray projection tube. It must lend itself to an inexpensive manufacturing process.

COMPONENTS OF THE RCA PROJECTION SYSTEM

The components of the RCA projection unit are shown in the photograph of Fig. 231. At the left is the optical barrel with second-anode cable to be fed through protrusion on the left-hand side of the barrel. Optical adjustments can

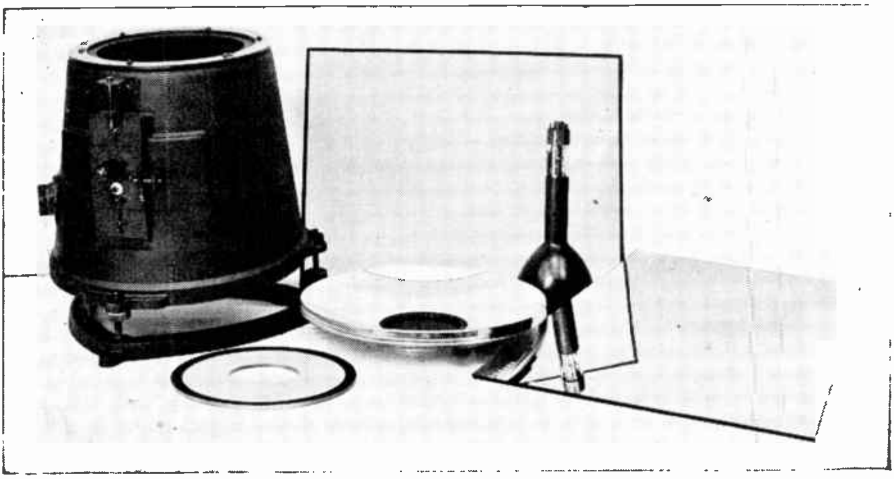


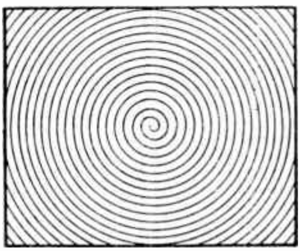
FIG. 231 Components of RCA Projection System

be seen on the front of the barrel. Left to right on the photograph, the next component is the corrector lens, followed by the spherical mirror. In the rear is the translucent viewing screen, and next in order is the picture tube and front-surfaced 45-degree inclined mirror.

One very unusual component of the RCA projection system is the translucent screen upon which the reflected image is thrown. This screen is directional and causes the light to emanate from the other side in a 55-degree cone. If a ground glass or similar screen were used, light would emanate in all directions and the light efficiency of the screen would be very poor. The actual translucent screen is constructed in three sections, a 1/16-inch section of lucite, a 1/1000-inch midsection of vinylite, and another outer 1/16-inch section of lucite.

In this composite viewing screen, the section of lucite near the mirror is a

Fresnel zone lens, which diverges the image into a cone of light with a 55-degree angle. A Fresnel lens, although only 1/16 inch thick, produces an equivalent lens having a thickness of 3 inches and extremely short in focal length. A zone lens consists of small ridges, 50 to the inch, as shown in Fig. 232, which become progressively deeper toward the periphery. To produce an equivalent of a 3-inch lens, the lucite is broken up into zonal ridges, each with a radius of curvature the same as the curvature of a 3-inch lens at its periphery, as demonstrated in the second drawing of Fig. 232. The ridges are very deep at the edges and almost smooth at the center of the screen, at which point a 3-inch lens would also be flat.



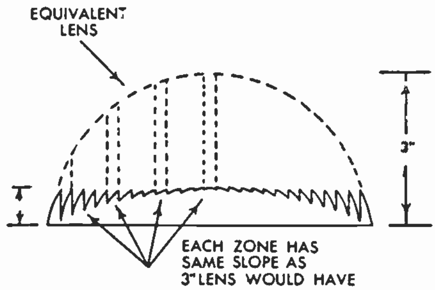


FIG. 232 Fresnel Lens System

The light is first diffused, after leaving the back lucite section, before it reaches the front viewing section. The front lucite section consists of vertical ridges, 100 per inch, which diverge the light rays into an ellipsoid, taking the 55-degree cone and producing effective radiation over a 55-degree angle, left and right, and only approximately 20 degrees above and below. The thin midsection is necessary to diffuse slightly the light passing between the two lucite sections, to prevent an interference pattern between the spirals of the back section and the ribs of the front section, and to diffuse the light to remove a dark spot at the center of the screen when image is viewed with eye level at the center normal axis of the viewing screen.

QUESTIONS

1. What is spherical aberration?
2. What causes halation?
3. Differentiate between reflective and refractive projection systems.
4. Compare various types of large-screen television systems.
5. In what manner does illumination vary with distance from light source?
6. How does a spherical mirror bring an object point to focus?

7. Explain magnification by a spherical mirror.
8. What is relation between placement of object with respect to principal focal point and magnification in a reflective optical system?
9. What angle rule applies to the reflection of light?
10. Explain formation of an image by a spherical mirror.
11. What are meant by radius of curvature, center of curvature, focal point, and principal axis?
12. What is meant by refraction of light?
13. Explain magnification by a lens.
14. How is spherical aberration in a reflective system corrected?
15. Describe in detail a commercial projection system.
16. Compare the three Schmidt optical systems discussed.
17. What is meant by a directional screen?
18. How is it possible to bring an optical surface into focus although one section of screen is farther from spherical mirror than other?
19. Detail alignment procedure for a commercial projection unit.
20. What is a Fresnel lens?
21. What is keystone correction?

Chapter 11

TELEVISION RECEIVER ANTENNAS

119. *General Antenna Types*

The antenna type for television reception can be simple or complex depending on height, signal strength, signal-to-noise ratio, and receiver sensitivity. For a good location in the primary area of the transmitter a simple dipole and good low-loss transmission line is satisfactory. In fact, if the receiver has a resistive input which matches the transmission-line impedance the antenna need not match the transmission line. This means there is some attenuation, and a lower signal-to-noise ratio, depending on the extent of the mismatch and line attenuation, but the antenna system will have approximately uniform sensitivity over a wider band of frequencies. To be free of transmission line reflections, however, the receiver must present the proper resistive termination to the transmission line.

In areas in which the signal is weak or there is difficulty keeping the signal above the noise level, a properly matched antenna is essential. A correctly designed and matched antenna not only results in peak signal strength so far as the match is concerned but it can be further improved, with the use of larger and/or additional elements, to have better horizontal and vertical directivity, resulting in better signal sensitivity and higher signal-to-noise ratio.

Antennas as used for television are either resonant or nonresonant. The dimensions of the resonant antenna are critically dependent on frequency of signal to be received and have, in most cases, a physical length less than a wavelength, generally a half wave for most common types. An example of the resonant or tuned antenna is the simple dipole (Fig. 233), the physical length of which is approximately a half-wavelength at the frequency to be received. Other tuned antennas, most of them dipole elaborations, are the folded dipole, V-antenna, fanned, and conical antennas.

The shape and dimensions of tuned antennas affect their bandwidth and, to some extent, their horizontal directivity. The more elaborate variations have greater bandwidths. The horizontal directivity of these resonant antennas can be sharpened with use of reflectors behind the antenna and directors ahead of the antenna. The directors and reflectors are generally parasitic elements and have no electrical connections to the main antenna element. In TV station relay and remote services higher gain reflectors, such as corner reflectors and parabolic reflectors, are used.

The vertical directivity of the antenna system can be narrowed to low angles by stacking antenna elements a half-wavelength apart. A typical stacked system consists of two dipoles and reflectors stacked one above the other. Any type of resonant antenna can be stacked with or without parasitic elements with a substantial concentration of the antenna sensitivity vertically.

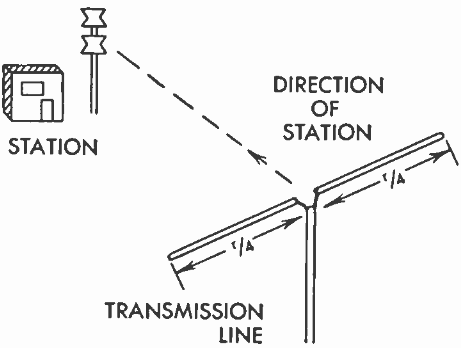


FIG. 233 Dipole Antenna Properly Oriented for Reception

The nonresonant or aperiodic antenna is sensitive to an extremely broad band of frequencies provided certain minimum dimensions are maintained. The length of such an antenna is at least three wavelengths. Both bandwidth and sensitivity rise as this dimension is increased. Typical aperiodic antennas are the rhombic and long wire. These long antennas, only occasionally used for home receivers, do find application in television relay and remote systems.

120. *Antenna Directivity and Space Patterns*

The television receiving antenna is wideband because it must have equal sensitivity to all frequencies in the channel. There are a number of good wideband antennas which have reasonably uniform sensitivity over a single channel. There are also some few which, although more elaborate, have uniform sensitivity over a number of channels.

The television receiving antenna is generally directional and is oriented to have maximum sensitivity in the direction of the transmitting station. Generally, the more directional the antenna is, the better its sensitivity becomes, in the sensitive direction, as compared to a nondirectional antenna. Another advantage of the antenna with pronounced horizontal sensitivity in the direction of the station is its insensitivity to noise which strikes the antenna from other directions. Thus, the directional antenna has improved sensitivity and reduced noise pickup.

Still another point to be considered is vertical directivity (angle at which antenna is sensitive with respect to the surface of the earth). Inasmuch as the frequencies used in commercial telecasting are extended line of sight, the vertical angle of radiation at transmission antenna is held down (20 degrees and under). Energy radiated at higher angles penetrates the upper atmosphere and does not return to earth. It represents lost power which might well be used at lower angles to improve signal strength. Thus, a feature which a good receiving antenna should have is a pronounced vertical directivity (angle at which antenna is most sensitive with respect to the surface of the earth) at low angles (20 degrees and under, same as for transmitting antenna). Here again, vertical directivity concentrated at low angles produces a more sensitive antenna than one which has no appreciable vertical directivity and further reduces noise pickup from above and beneath the antenna.

To improve vertical sensitivity at low angles, the antenna can be mounted one-half wave above good ground surface (this may be actual ground or a metallic surface, such as the top of a building which is efficiently grounded).

In the radiation of the signal from the TV-station antenna the energy is concentrated at low vertical angles, radiation spraying more or less horizontally over the surface of the earth. There is only limited radiation at high angles, and therefore little energy is lost to ionosphere penetration. The low-angle radiation skims over the surface of the earth to the receiver points. This same energy also strikes the earth at various points in accordance with its angles, terrain, curvature of earth, obstacles, and distance from station—striking the earth or obstacle at a low angle and glancing off at a similar angle to proceed on to various receiver points.

It is important to realize that any reflected signal at a given receiving point can add to or subtract from another signal which arrives at the antenna over a direct path (radiated at a slightly higher vertical angle initially than the reflected wave). This condition produces very definite areas of signal loops and nodes (space pattern). Of course, when an antenna is positioned in a so-called "space loop" more signal is picked up than if it were in a "space node." Actually, therefore, it is every bit as important to favor a weak station by placing antenna at a space loop as it is to have antenna oriented properly. Space loops in the direction of the station from receiving antenna occur approximately a full wave apart and in an open area loops are spaced uniformly. Presence of other reflecting surfaces (actual back or side reflection) or any other refractive or ionospheric conditions changes spacing at times and varies intensity ratio between loops and nodes. Except for occasional sporadic conditions loops remain in fixed positions. Loops and nodes of decided intensity separation exist even at altitudes far above the height of the higher receiver antenna installations.

121. *Dipole Antenna*

The simplest and one of the most common antennas is the simple dipole (Fig. 233), which is a half-wave antenna opened at the center for feeding and, consequently, divided into quarter-wave sections. It is important to realize that a quarter-wave in free space is longer than the physical length of dipole, which represents an electrical quarter wavelength. Thus, to make

Channel		*Dipole*	*Element*	*Reflector*	*Director*	*Folded Dipole*	*Conical Element*	*Wavelengths in Space—Element Spacing—Transmission Line Spacing, Air Dielectric—Multiply by Velocity Constant for Other-Than-Air Dielectric*							
		$\lambda/4$*	$\lambda/2$	$\lambda/2 + 5\%$	$\lambda/2 - 4\%$										
No.	Center Freq.	$\frac{2770}{MC}$	$\frac{5540}{MC}$	$\frac{5817}{MC}$	$\frac{5318}{MC}$	$\frac{11450}{MC}$	$\frac{4205}{MC}$	λ	$\lambda/2$	$\lambda/4$	0.1λ	0.15λ	3λ feet	10λ feet	Channel Limits
2	57	48.6	97.2	102	93.3	200	73.7	202	101	50.5	20.2	30.3	51.9	173	54– 60
3	63	44	88	92.3	84.4	181	66.7	182.8	91.4	45.7	18.3	27.4	46.8	156	60– 66
4	69	40	80	84.3	77	166	60.9	166.8	83.4	41.7	16.7	25	42.9	143	66– 72
5	79	35	70	73.6	67.3	145	53.2	145.6	72.8	36.4	14.6	21.8	37.2	124	76– 82
6	85	32.6	65.2	68.2	62.5	134	49.4	135.6	67.8	33.9	13.6	20.3	34.8	116	82– 88
7	177	15.6	31.2	32.8	30	64.7	23.7	64.8	32.4	16.2	6.48	9.72	16.7	55.6	174–180
8	183	15.1	30.2	31.8	29	62.6	22.9	62.8	31.4	15.7	6.28	9.42	16.1	53.7	180–186
9	189	14.6	29.2	30.8	28.1	60.6	22.2	60.8	30.4	15.2	6.08	9.12	15.6	52	186–192
10	195	14.2	28.4	29.8	27.2	58.7	21.5	58.8	29.4	14.7	5.88	8.82	15.1	50.4	192–198
11	201	13.8	27.6	29	26.4	57	20.9	57.2	28.6	14.3	5.72	8.58	14.7	49	198–204
12	207	13.4	26.8	28.2	25.7	55.3	20.3	55.6	27.8	13.9	5.56	8.34	14.3	47.5	204–210
13	213	13	26	27.3	24.9	53.8	19.7	54	27	13.5	5.4	8.1	13.9	46.2	210–216
Center LO Group	71	39	78	81.9	74.9	161	59.2	162	81	40.5	16.2	24.3	40.5	135	
Center HI Group	195	14.2	28.4	29.8	27.2	58.7	21.5	58.8	29.4	14.7	5.88	8.82	15.1	50.4	

* Dimension in inches except 3λ and 10λ.

FIG. 234 Antenna Dimension and Spacing Chart a dipole element resonant at a given frequency, it is necessary to shorten its length (because of capacitive end effect) with respect to any free-space calculation using a standard wavelength formula. For this reason, most tuned antenna elements are cut somewhat shorter. Likewise, it is necessary to cut tuned sections of transmission line (with other-than-air dielectric) shorter than free-space dimensions in accordance with dielectric constant of line. Spacing between various elements of the antenna is based on free-space wavelength formula.

The following extremely simple formulas can be used to calculate the physical and free-space lengths.

Length of a quarter-wave for antenna dimensions after considering end effect is:

$$\lambda/4 \text{ in inches} = \frac{2770}{\text{frequency in megacycles}}$$

Length of a free-space quarter-wave to be used in calculating spacing and matching section lengths is:

$$\lambda/4 \text{ in inches} = \frac{2875}{\text{frequency in megacycles}} \times \underset{(1 \text{ for air})}{\text{velocity constant}}$$

In Fig. 234, a chart has been compiled which gives the actual element lengths as computed for the center frequency of each of the 12 channels, and for the mid-frequency of the low-frequency channels and high-frequency channels.

ANTENNA EQUIVALENT

The dipole and other resonant half-wave antennas, when current fed at the center, have characteristics very much like a series-resonant circuit. The antenna has a certain Q, a reactive component, and a resistive component. At the resonant frequency of the antenna, its impedance is largely resistive and constant over a certain band of frequencies, depending on the Q of the antenna. For a wider band television antenna, a low Q is preferable because,

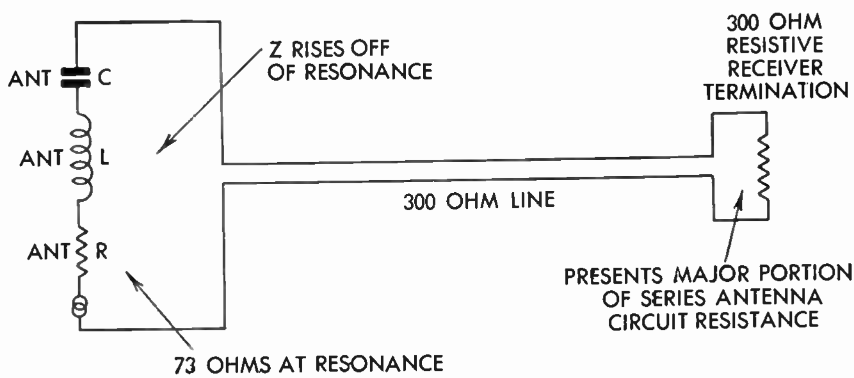


FIG. 235 Equivalent Antenna Circuit—Current Take-Off

as in a conventional lumped-constant tuned circuit, the lower Q tank circuit has a broader bandwidth. In a high-Q antenna, the reactance is appreciable (Q equals reactance over resistance) and rises rapidly off the resonant frequency of the antenna. Consequently, the impedance of the antenna, which is now *resistive* and *reactive,* increases and a mismatch occurs.

However, if a low-resistance antenna is made to feed a higher impedance transmission line, the equivalent antenna circuit (Fig. 235) shows us that the mismatch does not affect the relative sensitivity of the receiver over a wide band of frequencies. A common antenna used by television-set manufacturers is a simple dipole connected to 300-ohm line (Fig. 236). The characteristic resistance of a dipole at resonance is 73 ohms and, consequently, there is a 4:1 mismatch at the frequency at which maximum signal is induced into the antenna. Off the resonant frequency, the signal pickup is less but the antenna impedance increases (impedance is now the vector sum of reactive and resistive components) and the mismatch is not as severe. Thus, at a sacrifice in gain, this antenna presents a reasonably flat response over a number of channels and permits a simple antenna to be used on a number of channels.

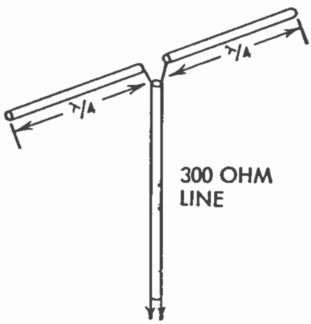


FIG. 236 Dipole Antenna—Mismatched

It is important that we realize the satisfactory operation of this system depends on three factors:

1. Strong signal from transmitter.
2. High signal-to-noise and interference ratios.
3. Receiver presents a 300-ohm resistive component on all channels. As shown, a mismatch of 2:1 causes only a slight attenuation.

Impedance ratio of mismatch to	**1**	**1.1**	**1.3**	**1.5**	**1.75**	**2**	**2.5**	**3**	**5**
1% additional power loss	**0**	**1**	**4.5**	**10**	**18**	**27**	**45**	**66**	**160**

FIG. 237 Power Transfer Loss Caused by Mismatch

Actual attenuation caused by mismatch can be observed in Fig. 237. Mismatch is based on standing-wave ratio or antenna resistance/line resistance (place higher value as numerator). It is important to realize that the effects of mismatch are more pronounced as line length increases—the additional loss is a prescribed percentage of an initially higher attenuation (more decibel loss because of line length).

122. *Transmission Lines*

The transmission line conveys the signal from antenna to receiver input. To reduce loss, attentive consideration must be given to the choice of a transmission line—type of line, physical dimensions, over-all length, and impedance match.

The television transmission line is an untuned feeder system because in any tuned system there is a definite impedance-frequency characteristic which would limit the wide-band response. Consequently, an untuned line which approximately matches transmission line to antenna and matches transmission line to receiver is necessary. This match is determined by the physical dimensions of the line, for every line, according to its composition, spacing, and size, has a characteristic or surge impedance.

There are two common types of lines—the parallel open-wire line and the coaxial line. In a parallel line wires are spaced a prescribed distance from each other; spacing is held constant by insulated spacers placed conveniently along the length of the line or by continuous low-loss flexible insulation. The parallel line has low loss and matches relatively high impedance (75 to 700 ohms); however, it is more susceptible to stray pickup, and for that reason the receiver input must be perfectly balanced and matched to reduce noise pickup. The surge impedance of a parallel line can be conveniently calculated from the formula

$$Z = \frac{276}{\sqrt{K}} \log \frac{2s}{d} \text{ ohms}$$

where s is the wire spacing and d the diameter of the conductor. Any unit of dimension can be used so long as the same unit is used for both s and d; k is the dielectric constant (1 for air).

A common transmission line for the television receiver is the coaxial line. Its advantages are the effective shielding afforded by the outer conductor, which prevents stray pickup, the fact that it is little affected by its proximity to other surfaces and matches a low impedance (50 to 150 ohms) which is characteristic of a dipole antenna. The ideal coaxial line has maximum Q and minimum resistance; however, these characteristics are in opposition to each other and an optimum value must be chosen.

A coaxial line has four parameters—resistance, capacity, inductance, and Q. The larger the inner conductor, the lower the resistance. A large conductor, however, means a low-value inductance and a high capacity, reducing the Q. A condition of maximum Q and minimum attenuation occurs for an air-dielectric coaxial line when the ratio between the diameter of the outer conductor to the diameter of the inner conductor is 3.6:1. With this 3.6-to-1 ratio the characteristic impedance is 77 ohms, which conveniently enough matches the resonant impedance at the center of a dipole antenna. Consequently, the dipole antenna (half-wave antenna opened and fed at the center) and coaxial transmission line form a much-used combination for television.

The above-mentioned ratio may be varied from 2.5 to 7 without increasing attenuation more than 10 per cent. Thus, at times, to secure a greater transfer of signal and a minimum noise pickup, the coaxial line is of some other impedance higher or lower than 77 ohms (varies between 50 and 150 ohms); the small increase in attenuation being counterbalanced by the increased signal amplitude delivered to the receiver. For the most common flexible coaxial lines, which use a polyethylene dielectric, the impedance is 50 to 90 ohms.

The surge impedance of the coaxial line using air dielectric is given by the formula:

$$Z_k = \frac{138}{\sqrt{K}} \log \frac{D}{d} \text{ ohms}$$

D is the inner diameter of the outer conductor and d is the outer diameter of the inner conductor.

Inasmuch as the surge impedance of a coaxial line with other-than-air dielectric is dependent on the type of dielectric, the surge impedance of this type of line must be obtained from the manufacturer.

A special parallel-line flexible cable with polyethylene insulation (dielectric constant 2.29) has been developed as a lead-in for television and FM receivers. The parallel line has dual stranded conductors and a characteristic impedance of 300 ohms. The dielectric is continuous, maintaining a constant conductor separation, plus flexibility. A 300-ohm parallel line with outer shield is also available. Many television receivers are designed with a standard 300-ohm input circuit to match this type of line.

Choice of a 300-ohm standard transmission line was dependent on a number of factors: impedance match, bandwidth, attenuation, noise pickup, flexibility, and cost. The simple dipole has a resistance of 72 ohms on the channel for which it is cut, and impedance of this antenna rises to a few thousand ohms on television channel frequencies removed from this channel. A 75-ohm coaxial line is an ideal match on this one channel but presents a serious mismatch on adjacent and remote channels. Use of a higher impedance line, 300–600 ohms, provides a better match over a number of channels. This improvement in bandwidth is obtained at a sacrifice of peak gain on a single channel. A 300-ohm line was decided on in preference to a 600-ohm line, despite the lower internal losses of the high-impedance line (loss inversely proportional to impedance), because of cost, adaptability to standard hardware, and reduced width. On the higher channels the wide separation between conductors of the higher impedance line becomes an appreciable percentage of a wavelength and additional losses arise due to radiation from the line.

The coaxial line, although it has effective shielding, has greater loss and a higher cost. The flexible 300-ohm polyethylene line costs less than inferior grades of coaxial line. Coaxial lines have a loss of 2 to 5 decibels per hundred feet at 50 megacycles as compared to 0.75 decibel per hundred feet for the 300-ohm line. At 200 megacycles losses are approximately the same.

For special applications where some of the characteristics of a parallel line plus the shielding of a coaxial line are desired, dual conductor coaxial line is used or two individual coaxial lines are run parallel to each other with their outer conductors tied together and grounded. For dual coax

$$Z = \frac{276}{\sqrt{K}} \log\left(\frac{2h}{d}\right)\left(\frac{1-\frac{h^2}{D}}{1+\frac{h^2}{D}}\right)$$

where K equals dielectric constant (1 for air); d, diameter of inner conductor; h, spacing between centers of inner conductor; D, inside diameter of outer coaxial conductor.

Although most transmission lines are approximately matched to receiver input, and perhaps to the antenna too, some standing waves still exist on the line. Thus, for most satisfactory reception of the weaker signals it is advisable to consider the line as a tuned line and to control the over-all length of line in such a manner that peak signal is delivered to receiver input for those stations to be favored. The presence of standing waves on twin lead can be readily detected by grasping the line at various intervals and noting the change in signal strength which results.

If the line is considered as a tuned line, which it must be to deliver peak signal when standing waves are present, the antenna functions as the generator and the receiver input as the termination for the line. Thus, if any reflections exist they start from the receiver end and set up standing waves with loops and nodes spaced a quarter wave back to the antenna. If maximum signal is to be delivered to the line from the antenna the impedance of the line ($Z = E/I$) at point it attaches to antenna must match antenna resistance. Inasmuch as this impedance is a function of the phase of the standing wave, it is possible by changing the over-all effective length of the line to reflect to the generator (antenna) the proper matching impedance. Since this impedance repeats every quarter wave the addition of a short length of line at the receiver of some critical segment of a quarter wave can produce a noticeable increase in signal strength. The correct length can be ascertained by observing signal strength on the picture-tube screen as various additional lengths are tried.

The section can be added between incoming line and receiver, or effective length can be changed by simply adding stub across antenna terminals of receiver and progressively clipping and shorting the stub until peak signal is delivered. Length of stub will vary with each installation (function of line length) and with channel frequency. Thus a stub is necessary for each weak station, or perhaps with considerable cut-and-try one stub can be found (quarter wavelength or less on lowest frequency channel) which improves results on two or more channels.

123. *Weak-Signal Dipole*

If the signal is not strong, noise components strong, or receiver not too sensitive, a properly matched antenna is essential. The dipole can be matched by either of the two systems shown in Fig. 238. In the first drawing a quarter-wave matching section of 150 ohms matches the standard 300-ohm line to the 75-ohm antenna resistance. Impedance of quarter-wave matching sections can be calculated with following simple formula:

$$\text{Matching } Z_M \text{ Section} = \sqrt{\text{Transmission Line } Z_L \times \text{Antenna } Z_A}$$

Be certain to consider the velocity constant of the dielectric when calculating the physical length of the matching section. Velocity constant is the reciprocal of square root of dielectric constant or $1/\sqrt{k}$.

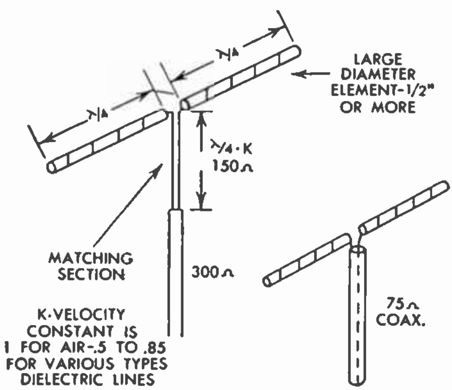


FIG. 238 Matched Dipoles

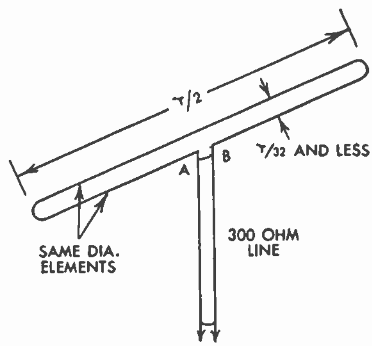


FIG. 239 Folded Dipole Antenna

In the second drawing the antenna is properly matched by attaching a 75-ohm coaxial line. This matched system is excellent as far as noise rejection is concerned because of the shielding of the outer conductor. However, for a very long span the 300-ohm line has less attenuation. When a 75-ohm line is used, be certain the receiver has a 75-ohm input. Receivers are now designed with 300-ohm inputs although some few have a 75-ohm input in addition, to be used in noisy locations. Some models have 90-ohm inputs to match 90-ohm coaxial line.

When the dipole is properly matched its bandwidth is narrowed considerably, and it performs best on only one channel because now the sharply rising reactance off resonance influences matching seriously (sharp, higher Q system). Consequently, it is important that the antenna Q is held down. In effect we wish to reduce the reactive component of the antenna while the resistive component is held essentially constant. This reduction in the reactive component is accomplished by increasing the diameter of the antenna elements (larger surface and less inductance per given length of dipole) causing a slower rate of reactance increase off resonance.

For a mismatched dipole it is permissible to use a small-diameter dipole element because the *Q* of this system is inherently low. When a higher gain matched dipole is used the element diameter should be more than one-half inch to obtain a good bandwidth.

Antenna *Q* can also be reduced by causing an increase in the resistive component of the antenna. This method is used to increase bandwidth in a number of the antennas to be discussed. At the same time the sensitivity of the antenna is not reduced.

A dipole antenna type cut for a given frequency will also function well on the third harmonic of this frequency (odd harmonic will be at a current maximum at point of feeder attachment). For example, a dipole or folded dipole sensitive to channel *3* band of frequencies will also be sensitive on channels *8*, *9*, and *10*. Third harmonic of channel *4* frequencies falls in channels *11*, *12*, and *13*; third harmonic of channel *2* falls in channel *7*.

124. *Folded Dipole and Modified Types*

The folded dipole (Fig. 239) is ideal for television because of its higher resistance and broader bandwidth. The folded dipole has a more constant impedance and uniform sensitivity over a wide band of frequencies because its electrical equivalent is represented by not only a series resonant circuit (as discussed for simple dipole) but a parallel resonant circuit in shunt with it. The parallel resonant circuit is contributed by the two quarter-wave elements shorted at the ends. Combination of series and parallel resonant circuits means the rate of impedance shift off resonance changes less rapidly because the reactance changes in both circuits balance each other. Its resistance is approximately four times greater than the resistance of a dipole element and, therefore, conveniently matches a 300-ohm line. If a shielded line is required, two 150-ohm coaxial lines can be used with their outer conductors common.

The folded dipole is a full wavelength of line bent around to form an antenna approximately one-half wavelength long. The spacing between the elements for best performance should be 1/32 of a wavelength or less.

The most common folded dipole is the antenna of Fig. 239. However, a number of other versions find wide application. One of these is the folded dipole constructed of three elements which has an impedance nine times that of a single dipole or 675 ohms (Fig. 240). In the second drawing a folded dipole with unequal element diameters is shown. When the element diameter of the fed element is decreased with respect to the other element the resistance increases; when the diameter is increased the resistance of the antenna decreases. Inasmuch as the side of a folded dipole directly opposite point at which transmission line is attached is a maximum current or ground point, it can be directly attached to the mast, putting the antenna at ground potential so far as lightning is concerned (if mast is grounded) but does not disturb antenna characteristics.

Another folded dipole often used for temporary or test installations is a folded dipole constructed of 300-ohm twin-wire line (Fig. 241). This type of covered antenna again should be approximately 0.95 $\lambda/2$ so far as signal pickup is concerned. However, so far as the currents in the antenna are concerned the dielectric constant is less than air and, consequently, it is necessary to shorten the antenna at 86 per cent.

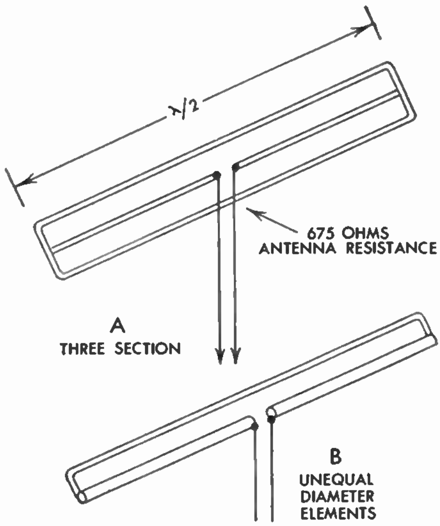


FIG. 240 Special Folded Dipoles

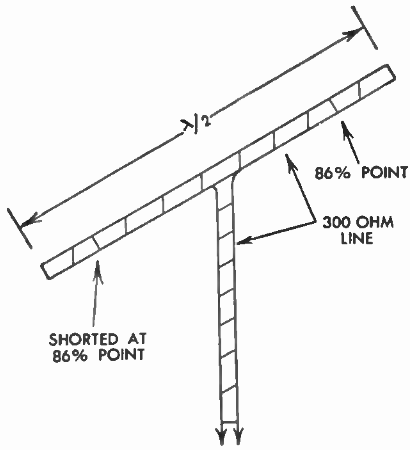


FIG. 241 Twin-Wire Folded Dipole

Therefore, as shown, the antenna is made the correct length, but it is shorted at a point 86 per cent from the ends. The ends may or may not be shorted together.

Some basic dipole types, Fig. 242, illustrate the various shapes that assist in obtaining specific characteristics and, in the case of television, a broader bandwidth. A circular-type construction improves the vertical directivity characteristics of the antenna and permits a reasonable bandwidth. A very common commercial antenna construction is found in the two fanned types or bi-conical construction of the next types. Higher impedance and greater bandwidth (as compared to a folded dipole) can be obtained with this type of construction. A still greater bandwidth can be obtained with the true conical or bat-wing construction used by some transmitting stations. However, this type of construction represents a much more expensive physical structure and is not widely employed for television-receiving systems.

It is significant that all these basic dipole types have approximately the same gain at the channel for which they are cut. For example, if we cut a straight dipole, a folded dipole, or a biconical for channel *3*, each will have the same sensitivity (assuming proper match) on channel 3. However, at channels removed from the cut frequency the relative sensitivities of the three basic dipoles will not be the same. In our example the dipole would have the least pickup in channel 6, with the fanned biconical types having the highest sensitivity and the folded dipole's sensitivity falling between dipole and conicals.

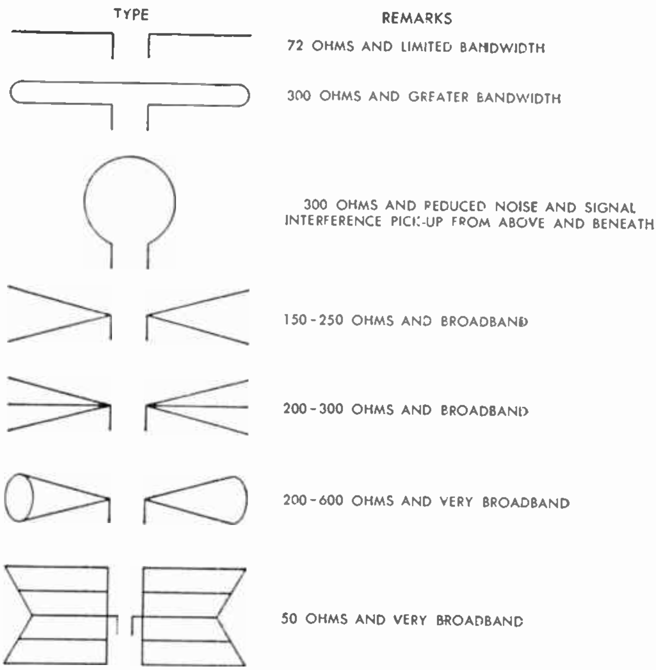


FIG. 242 Modified Dipole Types

125. *Antenna Patterns and Forward Tilt of Elements*

The polar pattern of an antenna is not difficult to interpret and tells much about the performance of the antenna. It indicates the direction or directions toward which the antenna is most sensitive with respect to the physical position of the antenna, Fig. 243. At resonant frequency a dipole has a figure-eight pattern—a line through the lobes of the figure eight indicates the two directions of peak sensitivity. These directions are broadside to the antenna element. The dipole has minimum sensitivity in the direction of the elements.

At frequencies higher than the cut frequency of the dipole, the antenna picks up additional lobes and becomes sensitive in more directions than the initial two. For example, at the third harmonic frequency of the antenna it becomes sensitive in six directions. This means an antenna system is sensitive in many directions and if the antenna is positioned physically to have either one of its lobes face the station, it is able to pick up appreciable energy. However, the presence of the additional lobes in the pattern means the antenna is sensitive to reflections and interference pickup at other angles, and consequently, co-channel and adjacent-channel interference problems can be magnified with this type of antenna and its multi-lobe pattern.

The directional pattern of an antenna that is to operate on higher frequencies than those for which it has been cut can be improved by proper tilting of the antenna elements. For example, if the dipole elements mentioned are tilted, the lobes are aligned to forward direction as indicated by the

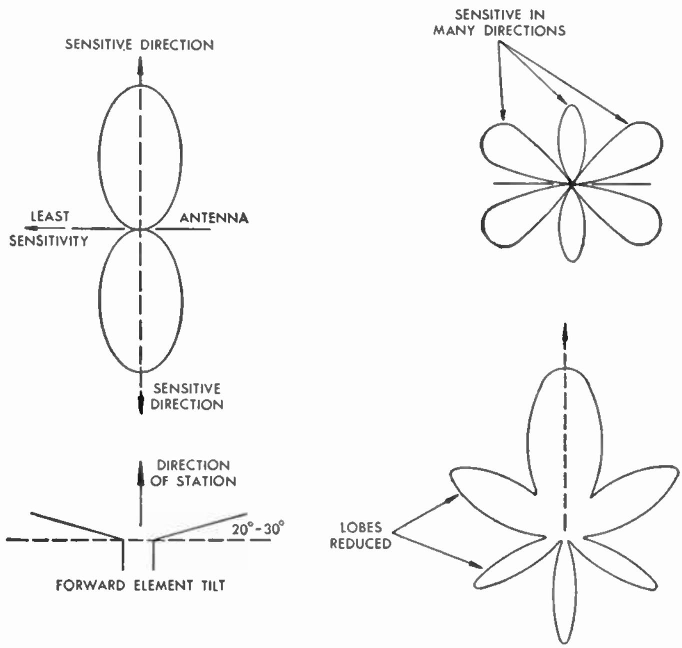


FIG. 243 Antenna Patterns

polar pattern of the tilted dipole. It can be seen that the lobes have been maximized in the direction of the tilt and that the back and side lobes have been reduced by a limited amount. Although this tilting does not completely remove secondary lobes, it does result in improved gain in the direction of the tilt.

An antenna pattern can also be used to depict the sensitivity of the antenna in the vertical plane. The patterns just mentioned have to do with the direction of signal arrival in terms of compass point or in relation to the bearing of the station with respect to the receiving location and the physical positioning of the antenna elements. However, an antenna system can also be judged in terms of its sensitivity to a signal arriving from above, in line with, or beneath the antenna itself. Inasmuch as the high frequency signal arrives at a low vertical angle (almost at zero in most locations), it is preferable to have the antenna sensitive at these vertical angles and not sensitive to interference signals that could possibly arrive from above or beneath the antenna system. It is possible to improve the vertical sensitivity pattern of a basic antenna by properly stacking a number of antennas one above the other.

126. *Reflectors and Directors*

To improve the horizontal directivity of the television antenna and to reduce noise pickup from the back, directors and/or reflectors can be used. The reflector is positioned in back (drawing A, Fig. 244) of the main antenna element approximately a quarter-wave and is 5 per cent longer than the dipole element. A director is generally positioned a quarter-wave ahead of the dipole

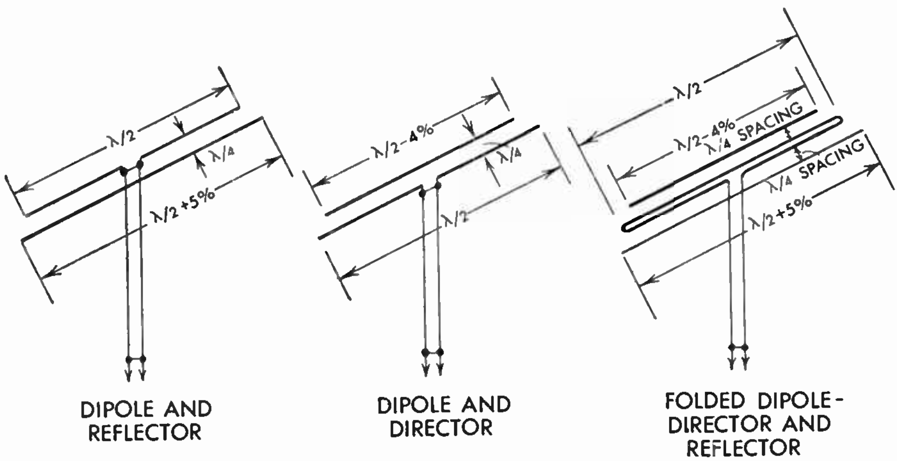


FIG. 244 Reflectors and Directors

element (with respect to the location of the transmitting station) and is 4 per cent shorter. Use of both reflector and director with a simple dipole is not recommended because of bandwidth reduction. If a director and reflector are used to obtain still better sensitivity a folded dipole should be used as the main element which is inherently broad and will prevent a serious reduction in bandwidth (drawing C). A dimension chart for directors and reflectors is tabulated in Fig. 234.

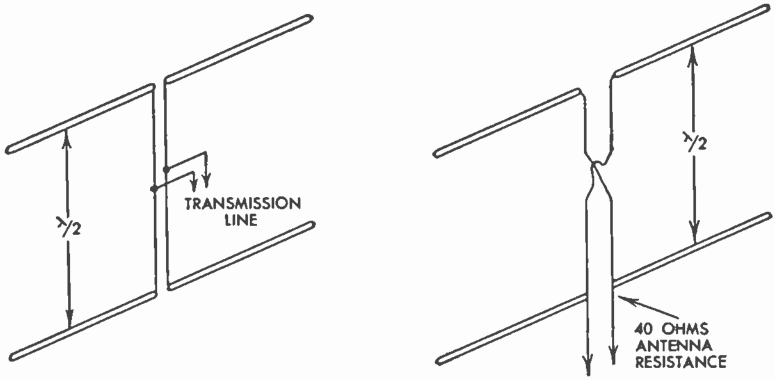
When a director or reflector is spaced a quarter-wave ahead of or behind a dipole or folded dipole element the antenna resistance decreases only 10 per cent to 20 per cent and, consequently, has little effect on the antenna-transmission line match. It is true, however, if the parasitic element is moved closer than a quarter-wave spacing, the antenna gain increases somewhat but the antenna resistance and bandwidth decrease considerably. Thus, if a simple dipole is used with close-spaced parasitic elements the bandwidth is occasionally too narrow. It is important to remember that the antenna resistance also decreases and a lower impedance transmission line must be used for a close-spaced parasitic element. In the case in which a parasitic element is spaced 0.15λ instead of 0.25λ, the antenna resistance is approximately halved.

When a folded dipole and a parasitic element spaced 0.1λ is used antenna matches a low-impedance coaxial line. It is apparent, therefore, that antenna resistance and transmission-line impedance required can be varied by changing the spacing of the parasitic element. Three considerations are given to choice of parasitic element spacing: gain, impedance, and front-to-back sensitivity ratio. If antenna is to be attached to a high-impedance 300-ohm line, spacing must be λ/4 to have least effect on impedance. For peak sensitivity and gain parasitic elements are brought near to main element. A substantial reduction in impedance occurs and antenna must be matched with a low-impedance coaxial line or an open quarter-wave matching section used. In some locations antenna is designed to prevent noise pickup from the back, and spacing is somewhat different from that required for peak gain to obtain a greater ratio between forward sensitivity and back pickup.

For least impedance change, parasitic spacing should be λ/4. Best gain is obtained with reflector spacing 0.15λ and director spacing 0.1λ. This close spacing reduces antenna impedance to less than 20 ohms for a dipole. Highest front-to-back ratio is obtained only at frequency for which antenna is critically cut. Therefore, if noise is particularly severe on one station, cut antenna and adjust spacing for peak gain on this channel.

127. *Stacked Array*

The stacked array is used to concentrate sensitivity at low vertical angles. The use of parasitic elements concentrates the antenna sensitivity in one horizontal direction and reduces side and back noise pickup. A stacked system reduces noise pickup from top and bottom and, therefore, further improves reception sensitivity in the desired direction.



A-CENTER FEED B-TRANSPOSED FEED

FIG. 245 Stacked Dipoles and Feed Methods

A simple stacked system consists of essentially two dipole elements spaced a half-wave vertically and excited in phase—that is, signal arriving in the proper direction to induce in-phase voltages into the antenna which add in-phase in the transmission line. Noise signals arriving from top or bottom induce voltages which cancel at the transmission line. There is a limited increase in gain—perhaps 40 per cent with an ideal match.

A simple stacked array and two methods of feeding it are shown in Fig. 245. In the method of drawing A transmission line point of attachment is centered and at the same distance from the dipole elements. Consequently, the signals from both dipoles always appear in-phase at the input to the transmission line. A transposed feeder system can also be used (drawing B), the signal from the top element appearing in-phase, through the transposition, with signal at bottom element where the transmission line is attached. However, the spacing has to be exactly a half wave and, therefore, the method supplies in-phase voltages over only a limited range of frequencies.

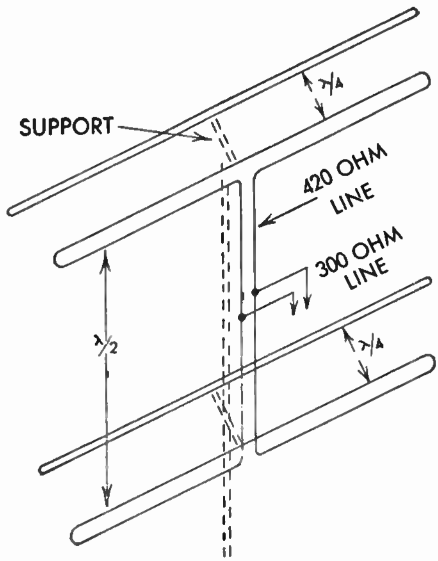


FIG. 246 Stacked Folded Dipoles and Parasitic Reflectors

When two dipoles or similar antennas are stacked and excite the transmission line in-phase the antenna resistance is halved. Thus, if two simple dipoles are stacked the antenna resistance is cut down to less than 40 ohms. These stacked dipoles, along with a pair of reflectors, are often used in a mismatched system, as discussed earlier. An ideal stacked folded dipole with reflectors, shown in Fig. 246, ideally matches a 150-ohm line. Only a slight loss occurs using 300-ohm line. A 300-ohm line connects elements. Another arrangement would be to use a 420-ohm section between dipoles, which matches 300-ohm transmission line. To use a 300-ohm line two 600-ohm points must parallel where line attaches. Thus each $\lambda/4$ section between center and dipole must match 300-ohm dipole to 600 ohms, or a 420-ohm section.

In summation, most any type of resonant antenna can be stacked with a resultant improvement in vertical directivity. A stacked antenna can also incorporate parasitic reflectors and directors. If the receiver has *only a 300-ohm resistive input,* always use a 300-ohm line and let the mismatch occur at the antenna where it does least damage.

An antenna system with additional elements alters the directivity pattern of the simple, driven, dipole-type antenna. In general, a group of horizontal elements which are either driven (connected to transmission lines) or parasitic (no connection to line) and which are mounted in the same horizontal plane as the initial driven elements, sharpen and concentrate the horizontal sensitivity pattern and make the system more unidirectional than bidirectional,

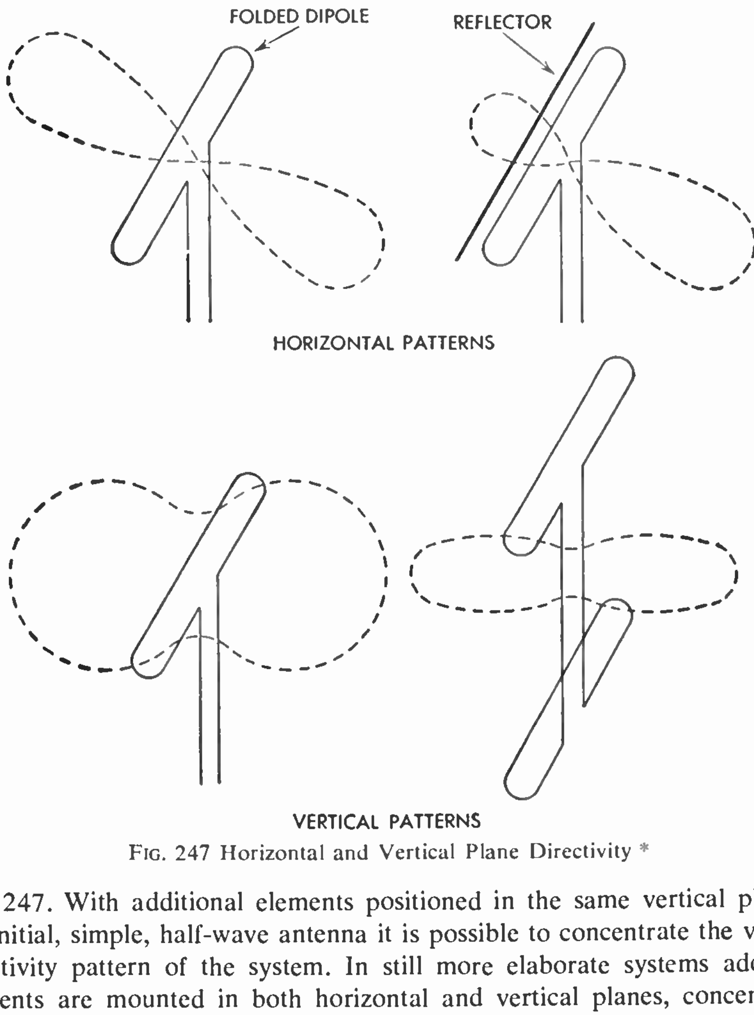


FIG. 247 Horizontal and Vertical Plane Directivity *

Fig. 247. With additional elements positioned in the same vertical plane as the initial, simple, half-wave antenna it is possible to concentrate the vertical-sensitivity pattern of the system. In still more elaborate systems additional elements are mounted in both horizontal and vertical planes, concentrating both horizontal- and vertical-sensitivity patterns.

128. *Phased Antennas*

A phased antenna system or phased array consists of a proper grouping of antenna elements to obtain a higher gain or a prescribed directional pattern.

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Parasitic elements are at times used along with phased elements. Generally speaking, a phased antenna system can be made to have a broader bandwidth for a given gain as compared to a directional system using parasitic elements. Antenna resistance need not fall to the exceptionally low value obtained with groupings of parasitic elements.

There are three basic phased antennas, Fig. 248, broadside, collinear, and end-fire. Two broadside elements can be stacked one above the other and fed in-phase, producing a broadside vertical sensitivity with a figure-eight pattern.

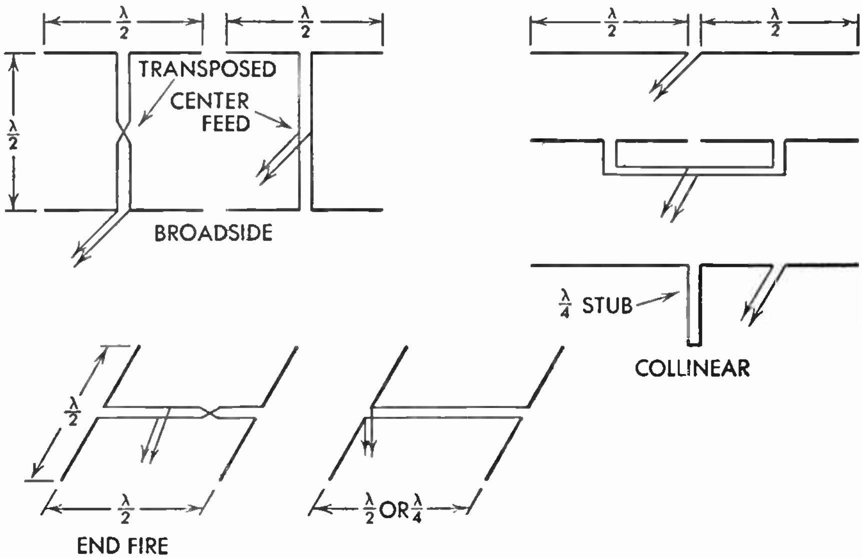

FIG. 248 Types of Two-Element Phased Antennas *

Broadside elements are most often found spaced a half wavelength apart. Collinear elements are positioned in the same vertical plane, side by side instead of being stacked one above the other. Collinear antennas are also fed in-phase and produce a sharpened figure-eight horizontal pattern, while vertical patterns remain essentially the same as for a single dipole. By comparison, the broadside elements obtain their additional gain by sharpening the vertical pattern, allowing the figure-eight horizontal pattern to remain essentially the same as for a single dipole.

End-fire elements are positioned in the same horizontal plane but are fed a hundred and eighty degrees out-of-phase. This type of antenna produces a sharpened figure-eight horizontal pattern and also sharpens the vertical directivity pattern. A number of the commercial phased systems use special combinations of phased elements to produce desired patterns. As far as comparative performance of the three phased systems is concerned, the broadside array

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ANTENNA TYPE	REMARKS	DECIBEL GAIN OVER DIPOLE		ANTENNA IMPEDANCE	
Straight dipole	Narrow bandwidth	0		72	
Folded dipole	Limited bandwidth	0		300	
Circle	Limited bandwidth	1		300	
Short V	Wide band	0		200–300	
Fanned	Wide band	0		200–300	
Forward tilt dipoles (any of above types)	Extends bandwidth	—		—	
Cone	Impedance changes with angle of revolution—wide band	1		200–600	
Driven element and director	Director 4 per cent shorter than driven element				
$\lambda/4$ spacing		3.5		$0.75R$[1]	
0.1λ spacing		5.5		$0.2R$	
Driven element and reflector	Reflector 5 per cent longer than driven element				
$\lambda/4$ spacing		4.5		$0.83R$	
0.15λ spacing		5.5		$0.35R$	
Driven element with director-reflector	Reflector 5 per cent longer—director 4 per cent shorter				
$\lambda/4$ spacing		6		$0.35R$	
close spacing		8		$0.14R$	
Four-element Yagi	Close-spacing	9		$0.1R$	
Collinear	Close end-to-end spacing of elements			Voltage feed	Current feed
Two-element		2		1500	—
Three-element		3.2		1000	300
Fanned		4.3		300	—
		Spacing		$0.4R$	
End fire	Fed 180 deg	$\lambda/2$	$\lambda/4$		
Two-element		2.2	3.8		
		Spacing			
Broadside	Elements in-phase	$\lambda/2$	$3/4\lambda$		
Two-element		3	4	$0.5R$	
Three-element		5	7	$0.35R$	
Four-element		6	8.5	$0.25R$	
Two-element broadside with parasitic reflectors	$\lambda/2$ spacing broadside—$\lambda/4$ spacing of reflectors		7	$0.4R$	
Lazy H	Two collinear elements broadside			Midpoint feed	End feed
			6	100	2000
Corner reflector	Driven elements $\lambda/2$ from apex—reflectors one wavelength at 90 deg		10	$2R$	
Long wire		5λ	10λ	High Z at voltage loop and low Z at current loop	
		4	7.5		
Rhombic		3λ	5λ		
		10	12	800	

[1] R = Resistance of driven element if no parasitic elements were present.

FIG. 249 Antenna Performance Chart *

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seems to be beneficial in those areas subject to noise pickup because of the substantial sharpening of the vertical directivity pattern. A grouping of collinear elements which sharpen the horizontal directivity could be definitely advantageous in areas subject to reflections and ghosts. In end-fire grouping there is some sharpening of both patterns. End-fire elements, however, are the most critical as to proper length and spacing and have a somewhat narrower bandwidth than have other phased systems.

129. *Antenna-Type Performance Chart*

A performance chart, Fig. 249, lists the various types of basic antennas with their gain and impedance characteristics. In most cases the impedance is given in terms of the normal resistance of the driven element. For example, with a driven element with quarter-wave spaced reflector, the impedance factor is $0.75R$. Thus if the driven element is a straight dipole, antenna resistance drops to 0.75×72 or 54 ohms; if the driven element is a folded dipole antenna resistance drops to 0.75×300 or 225 ohms. For some of the arrays and phased antenna systems it is not possible to give specific impedances because of the many variables which contribute to the antenna resistance such as element size, spacing, phasing, and number of elements. Only approximate figures are given, and the optimum matching can best be obtained experimentally by using matching stubs or Q sections. Gain and impedance figures are, of course, given for the frequency at which the antenna is cut. Antenna-gain figures are given in terms of decibel gain over a simple matched dipole cut to the same frequency.

It is quite difficult to calculate the decibel gain of the elaborate phased antennas because of the influence of mutual couplings between elements and various amounts of spacing. An approximation can be obtained by simply adding the decibel figures. For example, if a certain three-element antenna consisting of a driven element, director, and reflector has a gain of 6 decibels, the gain jumps, when two of these are stacked, to approximately 9 decibels (6 plus 3) because two antennas stacked one-half wavelength produce an additional 3-decibel gain. Stacking of four groups broadside and in phase increases the gain by 6 decibels, four such similar elements producing a total gain of 12 decibels.

130. *Commercial Antenna Types*

Perhaps the four most common television antennas are the folded dipole and reflector, Fig. 250, the high-low folded dipoles with reflectors, Fig. 251, the biconical dipole with X reflector, and the fan-type dipole with single reflector. The folded dipole and reflector which is generally cut for channels *3* or *4* has an average gain of 3 to 4 decibels, which gradually drops off at both ends of the low-band spectrum, declining to approximately one decibel at channel 6. It has a reasonably good impedance characteristic over this band of frequencies and a satisfactory response pattern. This type of antenna does not generally employ any forward tilt to its driven element, and consequently, the gain and response pattern are none too good on the high-band channels.

In areas where high-band channels are in operation a second, smaller, folded dipole and reflector are often mounted above the low-band antenna to permit reception of these higher-frequency channels. Often both antennas are

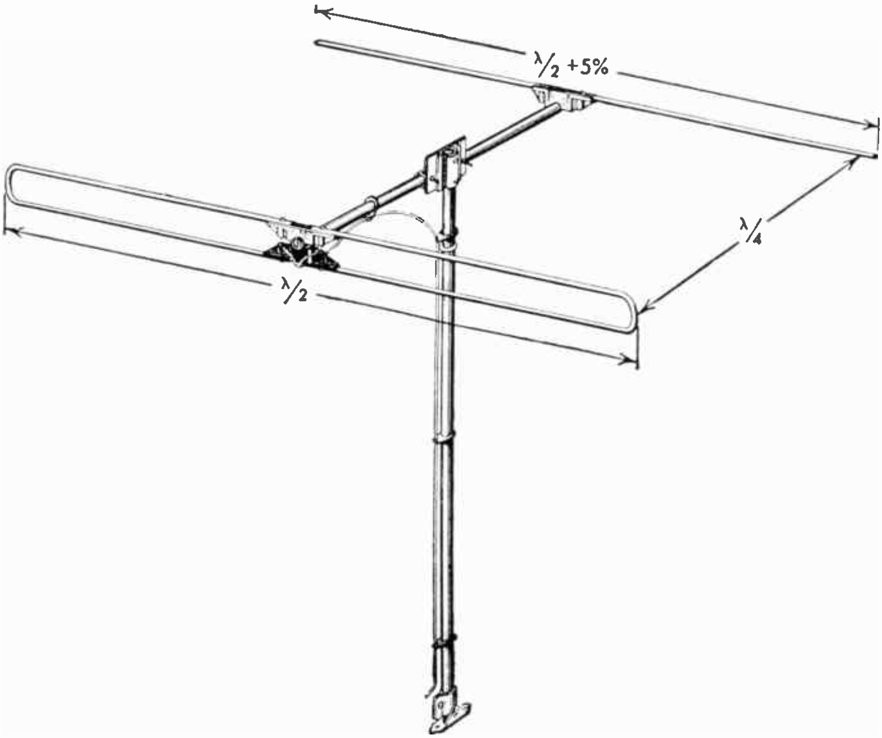

FIG. 250 Ward Folded Dipole and Reflector (Courtesy of The Ward Products Corporation) *

fed by the same transmission line through a suitable length of interconnecting line. However, if maximum gain and best pattern are to be derived from each of the antennas, a separate transmission line and a small double-pole double-throw slider switch at the receiver are advisable. This approach prevents mutual coupling between antennas and the resultant loss of gain or poor response pattern that could make the system more subject to ghosts, signal interference, and gain reduction.

A very common antenna, Fig. 252, is the biconical driven element with reflector. This type of antenna has a broader bandwidth and, when cut for

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channels *3* or *4*, has less drop-off in gain at the low-band extremities. In addition, the conical elements are tilted forward, producing a good gain and an in-line lobe pattern on the high-band channels. Thus in many areas where reflections are not particularly bothersome, a second high-band antenna is not necessary. This type of antenna has gains of 2½ to 4 decibels on the low band and from 4 to 6 decibels on the high band. The reflector element which

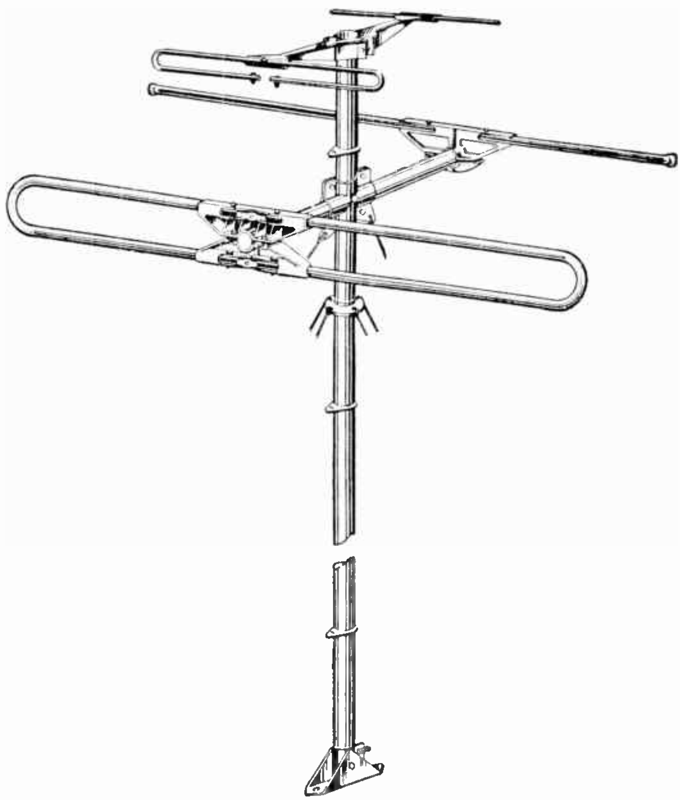

FIG. 251 Snyder HI-LO Folded Dipoles *

is spaced approximately at a quarter wavelength somewhere on the low band adds significantly to the gain on the low-band channels. However, its influence on the high-band channels is not as pronounced, and high-band gain is largely a function of the longer, driven elements (in terms of wavelength) and their forward tilt. The response pattern on the low band is good, but the response pattern on the high-band channels is only fair because of the many multiple lobes of rather high sensitivity. Consequently, it is sensitive to reflections and to possible interference of the co-channel or adjacent-channel types.

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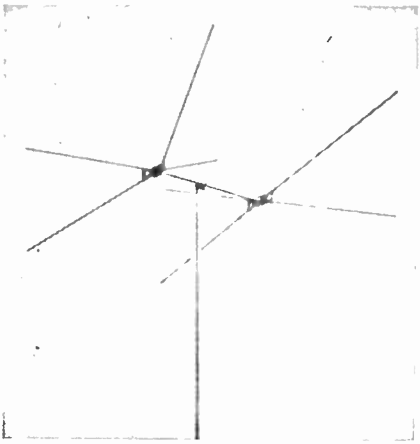

FIG. 252 Telrex Conical V-Beam Antenna (Courtesy of Telrex Incorporated) *

The six-element fan-type conical antenna with straight reflector has a somewhat more uniform impedance characteristic over the VHF band and also an improved gain at the high ends of the low- and high-band channels, particularly at the high-frequency end of the high band. This type of construction, Fig. 253, is often somewhat lower in gain at the low end of the low band.

The gain of the conical-type antennas can be raised and the vertical directivity pattern improved by using a stacked broadside combination of the basic style, Fig. 254. The normal stacking dimension of approximately a quarter

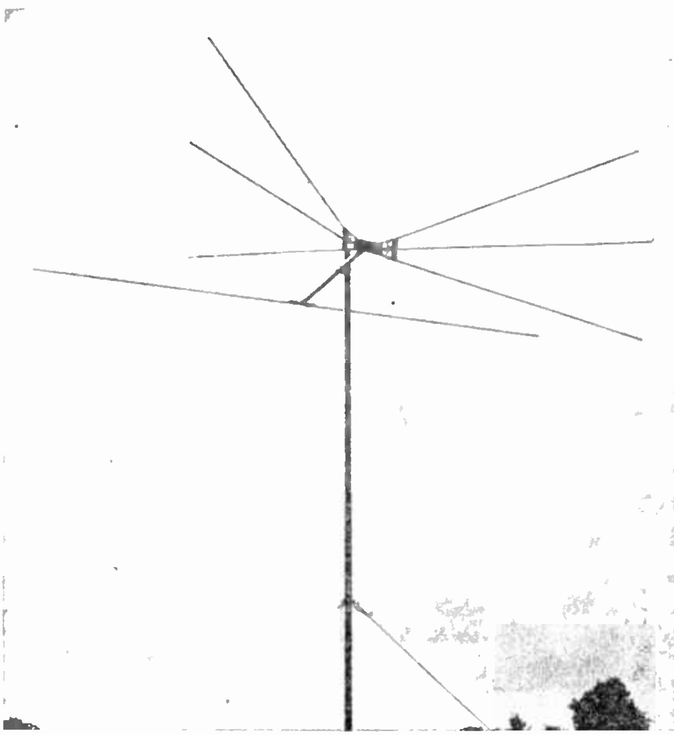

FIG. 253 Channel Master Fan Conical (Courtesy of Channel Master Corporation)

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wavelength on the low band permits a limited increase in gain on the low band (up to 2½ decibels) but a decided improvement in high-band gain (up to 5 decibels). A quarter-wave stacking dimension on the low band means a spacing of about ¾ of a wavelength on the high band and, therefore, a more optimum spacing dimension. Stacking on the low band can be improved by approximately doubling the stacking dimension (½ wavelength spacing) but only with less improvement in the high band because of a now greater than optimum separation in terms of high-band wavelength.

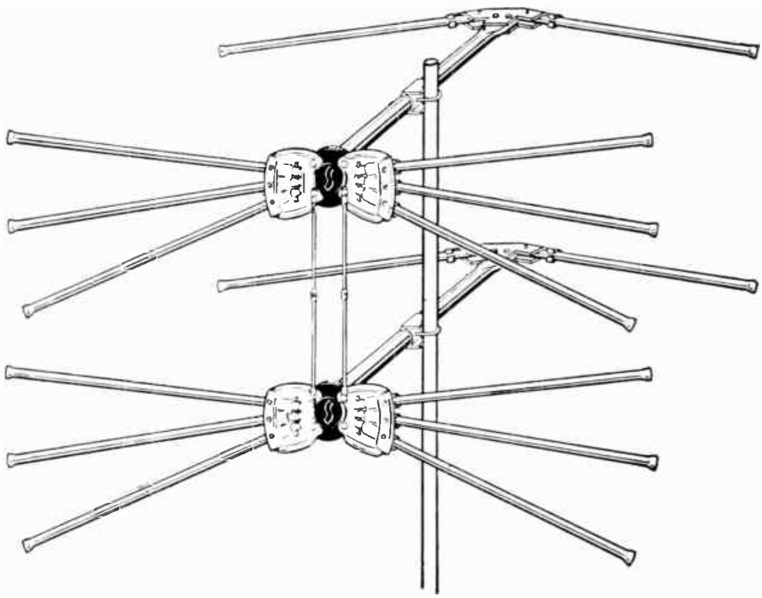

FIG. 254 Snyder Stacked Conicals

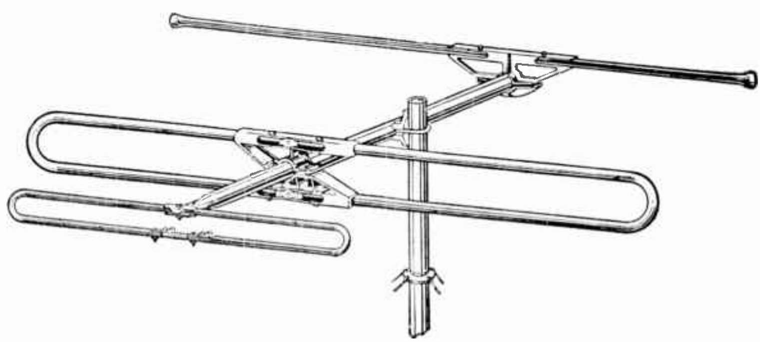

FIG. 255 In-Line Type Antenna *

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In general, the conical-type antennas must be oriented carefully on the low band to prevent picture smear at the low end of the band or echoes in the received picture at the high-end channels. High-band orientation is critical because of the sharper directional lobes and the presence of strong minor lobes that can introduce reflection or signal interference.

A very good response pattern is obtained with the Amphenol in-line antenna style. In this type, Fig. 255, two end-fire low- and high-band folded dipoles are employed, the shorter folded dipole helping to obtain a broader main lobe on the high band and to reduce minor lobes. Consequently, sensitivity at odd angles is reduced, and a maximum major lobe is obtained on all channels in a single direction. The gain of this type of antenna averages from 1½ to 3 decibels on the low-band channels and from 4 to 5 decibels on the high-band channels.

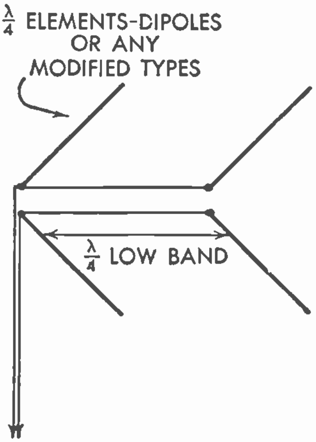

FIG. 256 End-Fire V Antenna (Courtesy of The Workshop Associates Incorporated) *

The end-fire V construction, Fig. 256, has two tilted V sections connected end-fire. Spacing between driven elements is approximately a quarter-wave-length on the low band. This type of antenna has a good response pattern on the high band and is often used where there is difficulty in receiving high-band channels because of multiple reflections. The low-band gain of this style of antenna is limited, and orientation is rather critical if picture smear is to be prevented on low-band channels in certain areas. This same style of antenna with a sharper tilt of its driven elements can be used as a UHF antenna in certain areas.

131. *Yagi Antennas*

The Yagi antenna provides the most economical and simplest means of obtaining a high gain and an excellent antenna pattern. However, its one limitation is bandwidth, which to a degree is overcome with a number of the new Yagi designs. The Yagis presently available for television reception consist of from 5 to 10 element types, Fig 257, having a gain of from 8 to almost 13 decibels as a function of the number of elements. This antenna when positioned carefully is able to make the best use of a received signal and has minimum pickup at other angles.

The wideband Yagi has many inviting features for obtaining peak performance on all channels and is of modest cost, representing a custom installation in terms of performance. With the increases in the number of stations,

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in transmitter powers, and in rising receiver sensitivities, the problems of interference from co-channels, adjacent channels, and other signals have increased. The pure pattern of the Yagi antenna minimizes these disturbances. However, in many locations the single-channel type of Yagi is outmoded because of the greater number of channels now occupied, and therefore the wideband Yagi types have been developed by a number of manufacturers.

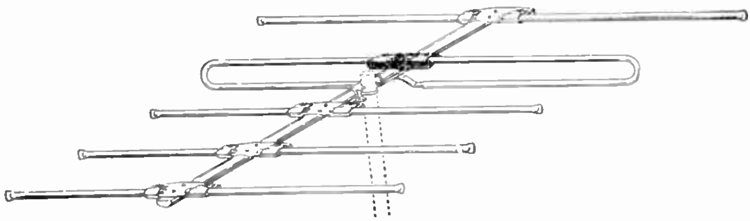


FIG. 257 Single-Channel 5-Element Yagi

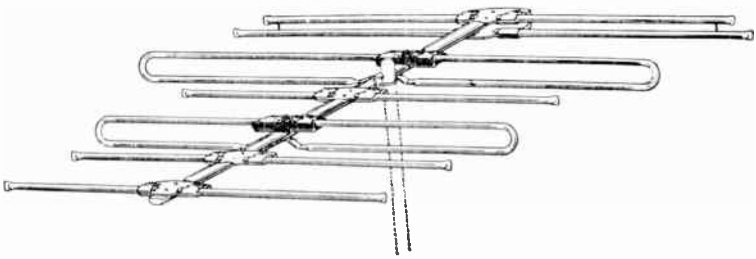


FIG. 258 Snyder Wideband Yagi

The basic plan of the wideband Yagi is to use twin-driven or end-fire folded dipoles as driven elements. The folded dipoles can be of the same or differing over-all length but are properly spaced and fed to make the system resonant and of constant impedance over a much broader band of frequencies. Length of element must be properly dimensioned to provide wideband service, with the reflector system generally favoring the lower end of the spectrum to be passed and the director system peaked at the high end of the spectrum to be received. A typical wideband Yagi, Fig. 258, employs two folded dipoles of differing lengths fed by a transmission-line phasing section between dipoles. It consists of a double reflector, designed to obtain the most suitable front-to-back ratio over the wide band of frequencies to be received. A group of from two to four directors can be placed in front to obtain peak gain and good front-to-back ratio at the high-frequency end of the spectrum to be received. An additional parasitic element is inserted between the driven dipoles and acts both as reflector and director, as a function of channel to be received—as reflector at the high end of the band and as director at the low-frequency end of the pass band.

It is possible to cover the entire television spectrum (channels *2* to *83*) with 3 or 4 wideband Yagis and to have the benefit of high gain and, what is equally important in the present state of television development, a much better response pattern, making the system less subject to interference. To derive full benefits from a Yagi installation, it is advisable to use separate transmission lines for each Yagi. Choice between antennas can be made at the receiver with the new miniature-type slider switches or, as recently developed, antenna-mounted slider switches that can be remotely controlled by a small switch-box mounted near the receiver. This seems to be a trend in many areas where both weak and strong signals are to be received with the best quality of picture and the least interference.

132. *Directronic Antennas*

In most television locations, the individual signals arrive from a number of directions. In a strong-signal area it is possible to obtain a strong signal when the antenna is not directed exactly at the station, but this same picture, strong as it might be, can contain reflections, smear, or be subject to interference. Consequently, to obtain the very best picture, it is advisable that the antenna system should be capable of orientation for peak results. In weaker-signal areas and fringe districts, with the increase of the number of stations, both weak and strong signals can be arriving at the receiving site from a substantial number of directions. Thus it is to be expected that some means of instantaneously switching the directivity of the pattern has definite advantages. A Directronic-type antenna that permits changing of the beam directivity without physical rotation of the element has a number of inviting features in terms of low cost, instantaneous switching, and small size.

A better understanding of the Directronic system can be obtained by reviewing some basic characteristics of antennas. Peak sensitivity of the standard half-wave dipole as mentioned earlier is broadside to its elements. Such an antenna has a figure-eight pickup pattern or is sensitive in two major directions (180 degrees related). The acceptance angle of such an antenna permits reception over a range of approximately 80 to 90 degrees. Thus there can be an appreciable error in orientation without too serious loss in signal level—an error of plus or minus 30 degrees maximum would result in only a 10-to-15 per cent decline in sensitivity. Such a slight loss goes unnoticed on the modern receiver with the a-g-c circuit in operation. Therefore, one possible method for receiving signals from any bearing through 360 degrees is that of mounting three separate dipoles on a mast, running three separate lines, and making a switch-mount at the receiver, thus permitting any line to be selected. The three separate dipoles, however, are mounted 120 degrees apart, producing a lobe maximum every 60 degrees (6 lobes). Desirable orientation advantages are thus secured, but the system has three separate lead-ins to contend with, and besides, is costly and inconvenient.

It is possible to obtain the benefits of such a system with a relatively simple installation—by mounting three quarter-wave conical elements 120 degrees apart in the same horizontal plane, as shown in Fig. 259. A single three-wire line connects the three elements to a switch at the receiver, and the switch selects any two of these antenna elements to form a half-wave antenna with a specific direction of orientation in exactly the same manner as can be done with the three entirely separate dipoles.

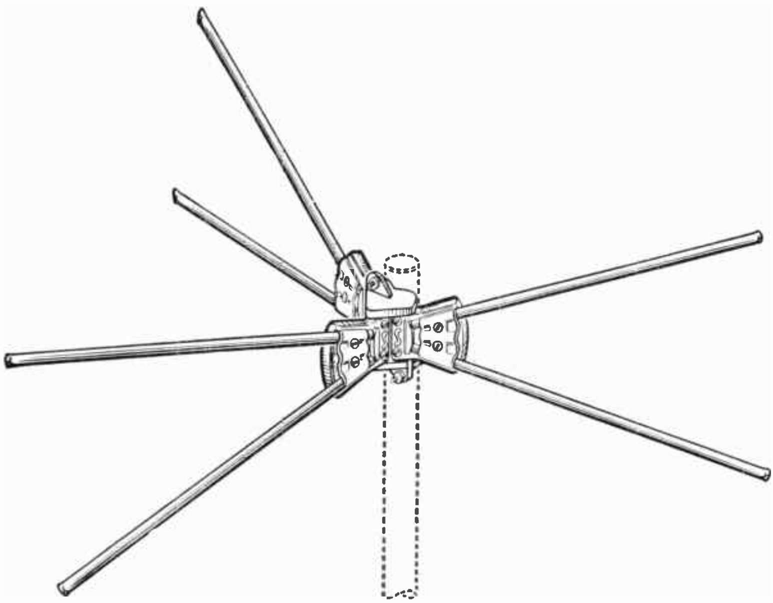


FIG. 259 Snyder Directronic Antenna

With the 120-degree mounting relation between quarter-wave elements there is a forward tilt to these elements—a tilt that does no harm to low-band performance and has the advantage of improving reception and raising gain on the higher channel by properly aligning the lobe pattern of the antenna. The Directronic antenna is generally cut for channel *3* and has the same broad acceptance angle through channel *6*. On the high-band channel, the lobes narrow, but the forward tilt of the elements increases gain. In a critical location for a particular high-band station, some slight initial orientation is of help in obtaining maximum results. Such an antenna lends itself to stacking for higher gain in fringe areas and reduces pickup from above and beneath. It has a balanced physical construction with weight centered at the mast.

There are numerous applications for such a lobe-switching antenna system. In metropolitan and near-fringe areas it solves orientation and interference problems. This is one way of minimizing the influence of noise, since it makes signals as strong as possible for each station without a compromise in antenna orientation. It should also be mentioned that in a strong-signal area good results can be obtained on at least two positions of the switch, due to overlap of lobes. The antenna can thus be helpful in minimizing not only local oscillator interference but reflections and other types of interference as well.

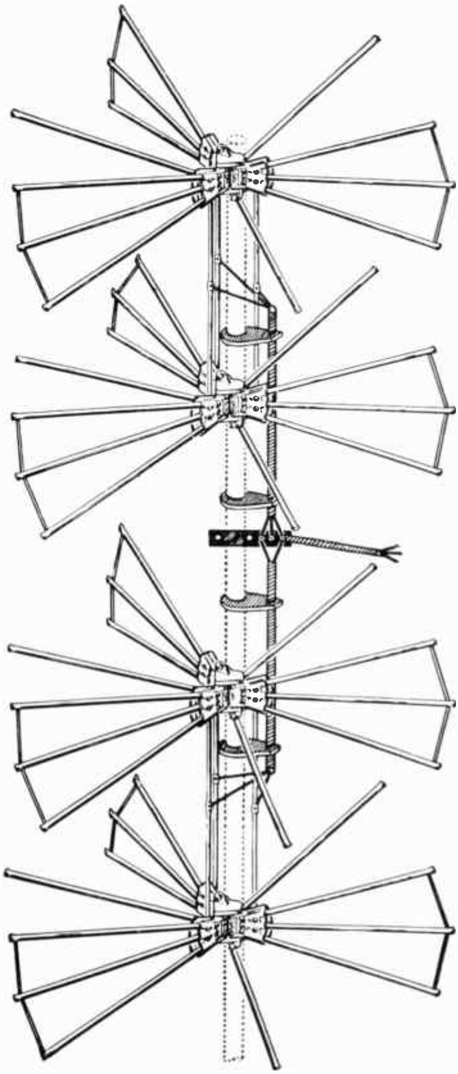


FIG. 260 High-Gain Fringe Directronic

A newer higher-gain Directronic has been designed specifically for use in fringe areas where signal level is lower. As shown, Fig. 260, it consists of three sets of fanned dipoles with reflector elements that are joined electrically and fan outward from the center of the structure. Fanned dipoles are spaced 120 degrees apart from each other while the reflector elements are interspersed with the same spacing. The ends of the fanned dipoles are joined electrically to provide higher gain at the low end of the low band and better impedance-matching to the transmission line with the close physical spacing of the reflector.

A special tubular three-wire transmission-line cable is used with a six-position switch at the receiver, permitting any pair of the driven elements to be chosen. In addition, there are three more switch positions that combine two antenna elements and use them in conjunction with the remaining element. The Directronic can be used either as a single bay or as an array with from two to four stacks, the number being a function of the gain required in a specific fringe area.

In many locations the Directronic type of antenna need not be oriented at the time of installation, although in overcoming certain difficulties and in obtaining the very best signal, orientation by the installer can be helpful. In particular, if a weak high-band station is to be received, orientation to favor this channel is obtained by directing two of the tilted driven elements in such a way that an imaginary line bisecting the angle they form is in the direction of the station to be received. On occasion, low-band smear can be introduced by the proximity of nearby obstructions. In this case, the antenna should be oriented in ten-degree steps and positions switched until the smear has been eliminated. In all antenna installations it is best to mount the antenna high and clear of nearby metallic surfaces. Adding even a few feet in order to get the antenna over the apex of a roof will materially improve the signal level. The transmission line should always be routed as direct as possible and as little lead as possible used. Line should not remain curled up in a roll. These rules apply for all types of antenna.

133. *Long-Wire, Rhombic Antennas*

A long-wire antenna, commonly called an *aperiodic antenna,* is not critical as to length so long as it is a number of wavelengths long. Long-wire antennas can be made highly directional—the directivity increasing with the number of wavelengths. The most common long-wire types used for television are the rhombic and long V antennas. The V antenna (Fig. 261) is made of two lengths of wire fed at the apex of a V. Maximum directivity is in a line passing through the center of the V. It is bidirectional.

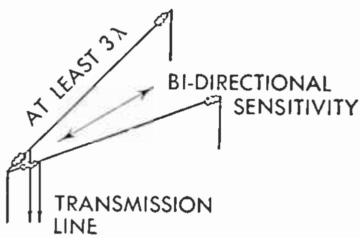


FIG. 261 Long V Antenna

A rhombic (Fig. 262) is formed by further extending the V into a diamond-shaped construction. The rhombic has improved horizontal directivity and a lower vertical angle but requires more space for erection. It is also bidirectional when it is not terminated, but can be made unidirectional by terminating the far end in an 800-ohm noninductive resistor. The antenna resistance of the rhombic is 800 ohms and can be suitably matched to a 500–800-ohm line or to a 300-ohm line through a matching section.

Inasmuch as the rhombic and V antenna are sharply directional, they must be staked with a compass before construction. The angle of the antenna legs

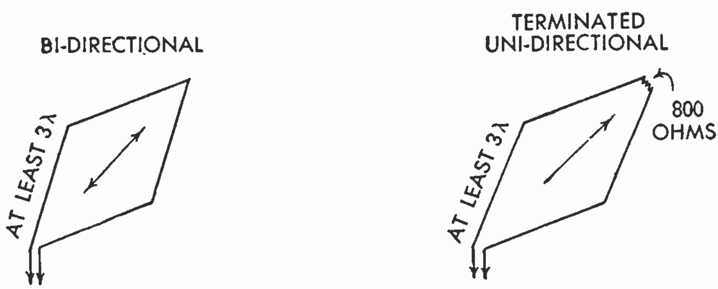


FIG. 262 Rhombic Antennas

must be considered in the layout. For satisfactory service the tilt angle (angle between antenna wire and direction of wave propagation) is considered satisfactory if the length of each antenna leg is made a half wavelength longer than a line which starts at the apex and is stepped off toward the station to a point where a right-angle line will meet the other end of the antenna. (Refer to Figs. 263 and 264.) It is apparent, therefore, that the longer the antenna is in wavelengths the smaller the tilt angle becomes. The long-wire antennas have exceptional bandwidth, and one antenna will will take in all television channels if it is made at least three wavelengths on a leg on the lowest frequency channel. Long-wire antennas require considerable space and are used only for extended range reception and television-station relay systems.

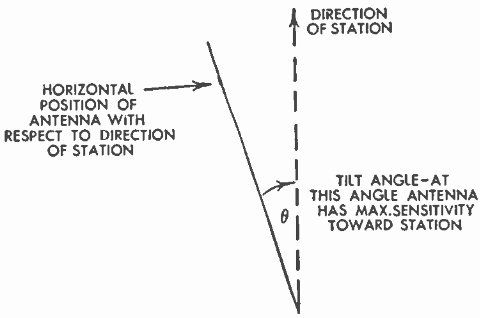


FIG. 263 Tilt-Angle Layout

CONSTRUCTION OF A RHOMBIC

In the construction of a rhombic the first step is to estimate just how many wavelengths on a leg can be erected on the space allotted. The more wavelengths that can be used the sharper and more sensitive the antenna will be. The next step is to lay off a string line from the near end to the far end in the direction of the station (Fig. 264). If it has been decided that the space will accommodate 5 wavelengths to a leg, step off 4½ wavelengths from the near end (end to which transmission line will be attached) along the station-direction line. At the 4½-wavelength point (to be center of entire rhombic), lay off a perpendicular line. The proper tilt angle is being used when

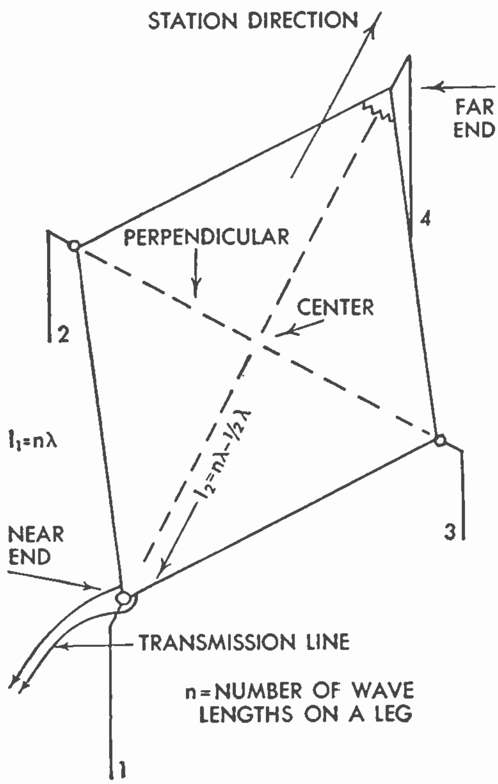


FIG. 264 Rhombic Construction

the very end of a 5-wavelength section of line attached to the near end just coincides with this perpendicular.

With the above procedure, the positions of two points of the diamond-shaped rhombic have been located and two poles can be erected. A similar procedure can be used to locate the third point (pole 3). The fourth point is now another 4½ wavelengths down the station-direction line from the point at which the perpendicular crosses the station-direction line.

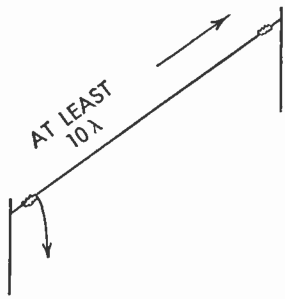


Fig. 265 Long-Wire Antenna

SINGLE LONG WIRE

A single long wire is also highly directional off its end if it is 10 wavelengths in length or better (Fig. 265). This is really understandable when we consider that the tilt angle becomes smaller and smaller as number of wavelengths increases. This, too, is a high-impedance type and must be attached to a receiver which has high-input impedance facilities.

134. *Multi-Outlet Antenna Systems*

A number of factors influence the choice of a satisfactory multi-outlet antenna system:

1. Frequency of station or stations to be received.
2. Absolute and relative strengths of signals (dependent on transmitting and receiving antenna heights, transmitting power, and distance separation).
3. Direction of various stations.
4. Number of receiver outlets.
5. Proximity of other objects and antennas.

All of the above factors must be considered in the choice of a system and, therefore, each location presents its own problems. They can, however, be grouped into a number of general classifications.

A. *Strong Signals—Less than Six Outlets—Stations in Same General Direction.*
Under these conditions a single wideband antenna for low- and high-channel groups, such as a folded dipole or conical antenna, is satisfactory. To prevent interaction between receivers and receiver local oscillations an outlet pad, such as shown in Fig. 266, is used. This pad, although it attenuates the signal, prevents interference between receivers.

B. *Weak Signals—Less than Six Outlets—Stations in Same General Direction.*
If signals are weak it is necessary to use booster amplifiers (wide band) to amplify the r-f signals from the antennas before distribution to the various outlets.

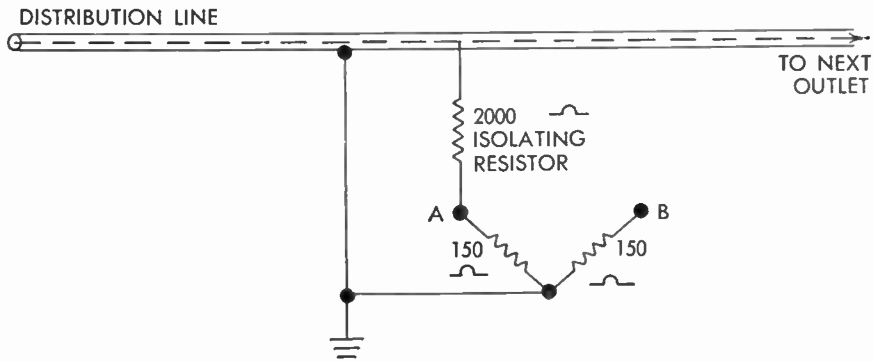


FIG. 266 Outlet Pad

C. *Less than Six Outlets—Stations Not in Same General Direction.* If the stations are not in the same general direction it is necessary to use antennas, such as circular folded dipoles, which have equal sensitivity in all directions. In areas where signals are weak or reflections particularly troublesome, separate antennas must be used for each station or specific groups of stations. Each antenna system feeds a mixing stage the output of which feeds the various outlets.

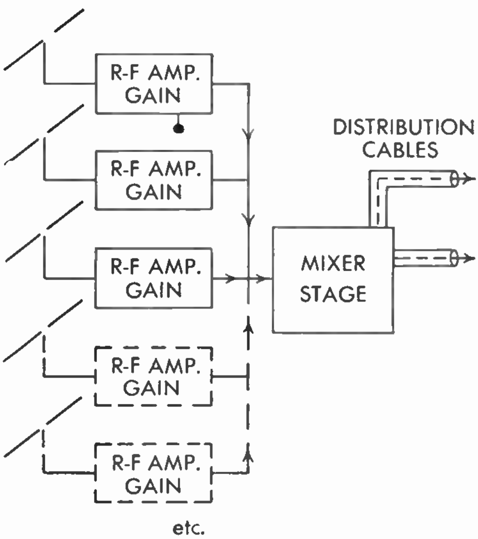


FIG. 267 Multiple-Outlet System. Separate Antennas

D. *Multiple Outlets—Various Directions and Signal Amplitudes.* The only really satisfactory multiple-outlet system which provides proper signal levels and minimum interactions is one which uses a separate antenna and, in most cases, an individual amplifier for each station. Outputs of the amplifiers feed a mixer, which in turn feeds the distribution system (Figs. 267 and 268).

OUTLET PAD

At each point where a receiver is connected an isolating outlet pad is used to isolate the receiver from the common distribution line. The pad (Fig. 266) consists of a series isolating resistor of a comparatively high value and two small resistors which serve as a match to the receiver input. A 300-ohm input receiver is attached between points *a* and *b*; a 75-ohm input receiver, between *a* and *b* (shorted together) and *c*.

It is apparent that the series resistor attenuates the signal, but at the same time it prevents local oscillator feed between receivers and prevents one receiver from loading the other receivers which are fed by the distribution line. Consequently, receivers can be attached and detached, turned on and off, or tuned from channel to channel without affecting the remainder of the receivers on the line.

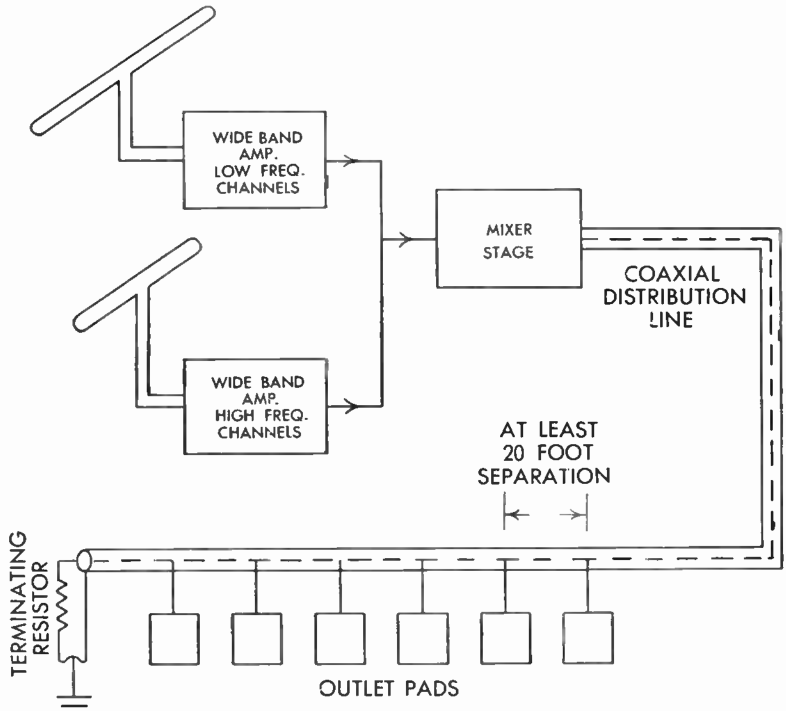


FIG. 268 Receiver Feeding With Outlet System

Outlet pads are spaced at, at least, 20-foot intervals along the distribution line to further minimize interaction. The special requirement of the system is to supply a strong signal to the line, enough signal that the pad attenuation cuts down the signal to a level for suitable excitation of the receiver.

TYPICAL MULTI-OUTLET SYSTEMS

Typical multi-outlet systems are shown in Figs. 267 and 268. In Fig. 268 a system is shown for use with only a few outlets. One folded dipole is used for the low-frequency set of channels; a second dipole for the high-frequency set. Each antenna has its own wide-band amplifier which feeds a circuit which has no gain but serves as an extremely wide band-pass stage which mixes all the television signals (same as a broadcast studio mixer that mixes all the various microphone outputs) and feeds them to a terminated coaxial line.

Along this distribution line the various outlets are attached with a minimum spacing of 20 feet.

A more elaborate distribution system is shown in Fig. 267, and can be designed to feed 50 or more outlets. In this system a separate antenna is used for each station. Each antenna feeds an individual amplifier with a gain control. Consequently, the signal amplitudes from all stations can be equalized and constant amplitude signal fed into the distribution line from all stations.

QUESTIONS

1. What factors influence choice of antenna type and complexity?
2. What factors influence vertical and horizontal directivity of an antenna?
3. Give all dimensions for a dipole and director cut for channel *6*—element lengths, spacing, and quater-wave matching section to a 300-ohm line (velocity constant of dielectric = 0.67).
4. What are the advantages gained by mismatching a dipole to a 300-ohm line?
5. Discuss the importance of a good low-loss transmission line.
6. Compare characteristics of open-wire and coaxial transmission lines.
7. What are the advantages of a folded dipole?
8. Give all dimensions for a folded-dipole director-reflector antenna to be used on the high-frequency set of television channels.
9. Design a stacked array to be used for channels *3, 6, 10,* and *12* (use folded dipoles and reflectors).
10. How does a stacked antenna cancel noise pickup from the street?
11. What are the characteristics of a fanned antenna compared to a conventional dipole?
12. Design a cone antenna to be used on low set of TV channels.
13. What are the characteristics of long-wire antennas?
14. Design an omnidirectional antenna to be used on the high set of TV channels.
15. Discuss a rotating antenna system.
16. What would be dimensions of a rhombic antenna if it is to be 3λ on a leg on lowest frequency TV channel?
17. Mr. A lives 2 miles from one TV station and 10 miles from the second, which is in just the opposite direction. Describe an economical antenna and reason for your choice.
18. Mr. B lives 10 miles from both stations in a noisy congested area. The stations are not in the same general direction. What type of antenna would you install and what tests would you make?
19. Mr. C lives 36 miles from transmitters in a rather crowded section of a small city. What antenna would you recommend?
20. Mr. D lives 80 miles from station concentration. What antenna would you recommend?

Chapter 12

INSTALLATION, ADJUSTMENT, AND OPERATION

135. *Installation Procedure*

Many television receivers are delivered and unpacked; some are competently installed and adjusted. Many TV antennas are fastened to roofs; some are erected to give peak performance and rigidity. The performance of a television receiver is a function of the installation as well as the receiver proper. A poor installation and one requiring frequent return calls reflects on dealer, manufacturer, and, what is most detrimental, on television itself. Installation crews, both independent and those of manufacturers, must realize that a painstaking installation is the most economical path to customer satisfaction.

Customer confidence gains with every inconvenience removed and every precaution taken to prevent property or defacement damage. There is no need to make the customer's house a receiving department or his living room an assembly line. His home is not a testing point for a defective receiver. All television receivers should be assembled at the shop and given a lengthy test before they are delivered. In transit between the shop and customer they can be wrapped in moving blankets and should ride so they are not subject to shock.

INITIAL SURVEY

Initial survey of location is of paramount importance and involves a number of decisions—what type of antenna, antenna position and tentative orientation, type of support and guying, location of receiver, and path of transmission line. The choice of antenna itself is often a difficult task because it is dependent on relative signal strengths, the number and frequency of stations, direction of stations, and signal-to-noise ratios. In general the antenna is either a wideband type or cut to favor the weak channels in the area. In areas in which all signals are weak separate antennas are often necessary. Likewise, antennas must be directed away from noise sources, whenever possible. The type and height of the mast depend on location and building structure. Support and guying of mast to withstand high winds is very important. Likewise, the antenna mast should lock in tightly after orientation to prevent a shift in antenna-element position and antenna directivity over a period of time (a frequent cause of gradually degenerating performance).

The general positioning of the receiver in the house should be ascertained and a plan evolved for running the transmission line unobtrusively and rigidly. Transmission line should be tied down and isolated from metallic and other surfaces against which it may rub or transfer signal.

On many jobs the survey and antenna installation proceed concurrently in the hands of an experienced two-man crew. The most important consideration in the erection of the antenna is rigid support and effective guying so it will withstand wind and will not shift position. The antenna should be high and in the clear. It is worthy of note that in city areas the signals often seem to be in layers and an increase in signal strength is often encountered when the antenna is lowered. This condition is attributed to a wave pattern in space, reflections, and a shift in the vertical angle directivity of the antenna as its wavelength above a ground is changed.

In city areas and noisy locations the best antenna position can be found only by searching for it. This is particularly the case when optimum performance is desired on a number of stations. To assist in this task a phone link is established between the roof and the test receiver at the tentative receiver location in house. The antenna is shifted to various positions and performance checked as to signal strength, noise, and reflections. When the optimum point is found, the antenna is securely clamped in that position.

The antenna installation should be neat and as little an eyesore as possible. Erect the antenna without damage to the customer's property.

RECEIVER INSTALLATION

The actual antenna installation and orientation can be made with the assistance of a test receiver or field intensity meter, which in the hands of the crew permits a thorough check of antenna performance. In addition the receiver is tuned and adjusted for peak performance before it reaches the customer's house.

The customer's cooperation must be obtained in choosing a position for the receiver. Place it where it does not require shifting of furniture whenever it is to be used. It is preferable to have it away from intense lighting at night and away from bright sunlight for daytime viewing.

A responsible member or members of the family must be taught, not told, complete operating procedure. This is extremely important if they are to receive optimum results and you are to be freed of unnecessary call-backs for minor adjustment. The installation man should now observe and guide the customer as he goes through tuning procedures. Concise written instructions can be prepared and left with the customer to augment verbal instructions.

The above procedures, if followed, present a direct approach to TV installation. They do not require any increase in man-hours and most definitely cut down return calls and installation uncertainties. The customer's house is not used for receiver assembly and there is no collection of debris to be carted away. At the instant the customer's receiver is turned on he will see a clean picture, which is likely to remain so for a considerable period of time.

136. *Test Chart*

The station test chart transmitted by various TV stations contains a wealth of information useful in judging the performance of television receivers. At present the charts have not been standardized and vary from station to station with respect to the type of data presented and the accuracy and extent with which the performance of the receiver can be judged from the chart.

A standard chart has been designed by the Radio-Electronics-Television Manufacturers Association (Fig. 269), which, if adopted by all television stations, will permit uniformity in station charts and presentation of more information than is transmitted on the present test charts. With the RETMA standardized chart it is possible to check resolution of the pattern at all sections of the scanning raster. Other checks which can be made with the received test chart are contrast, linearity, aspect ratio, interlacing, low- and high-frequency response, brightness, and shape of the scanning beam spot. Thus, the data on the standardized RETMA chart permit a thorough check of receiver performance, presenting substantially more data than the charts which are on the air today. It is unlikely that any station will use a chart which contains any more information. Therefore, a thorough study of this RETMA chart is significant because each station chart contains some of the data present on this standard but, at present, only a few special charts can compare with it.

RESOLUTION CHECK

Resolution is measured horizontally and vertically. The horizontal resolution is dependent on the frequency response of the system and the amount of detail which can be conveyed as the beam scans one line left to right. The faster the television system responds to an abrupt change from black to white or white to black the better the resolution becomes and the more distinctly this system can present a very thin line or sharp detail. It is evident, if a series of thin, vertical, white-and-black lines are transmitted, that the rate at which the signal changes, as these lines are scanned, is dependent on the width and nearness of the lines to each other—the narrower the lines the faster the system must respond to a sharp change in brightness. Thus, a series of vertical lines are used to judge the horizontal resolution of the television system and, more specifically, the performance of the receiver under check. The narrower the lines and the nearer they are to each other the faster the signal voltage must rise and fall in a given amount of time, because the scanning beam moves at a constant velocity, left to right, across the screen.

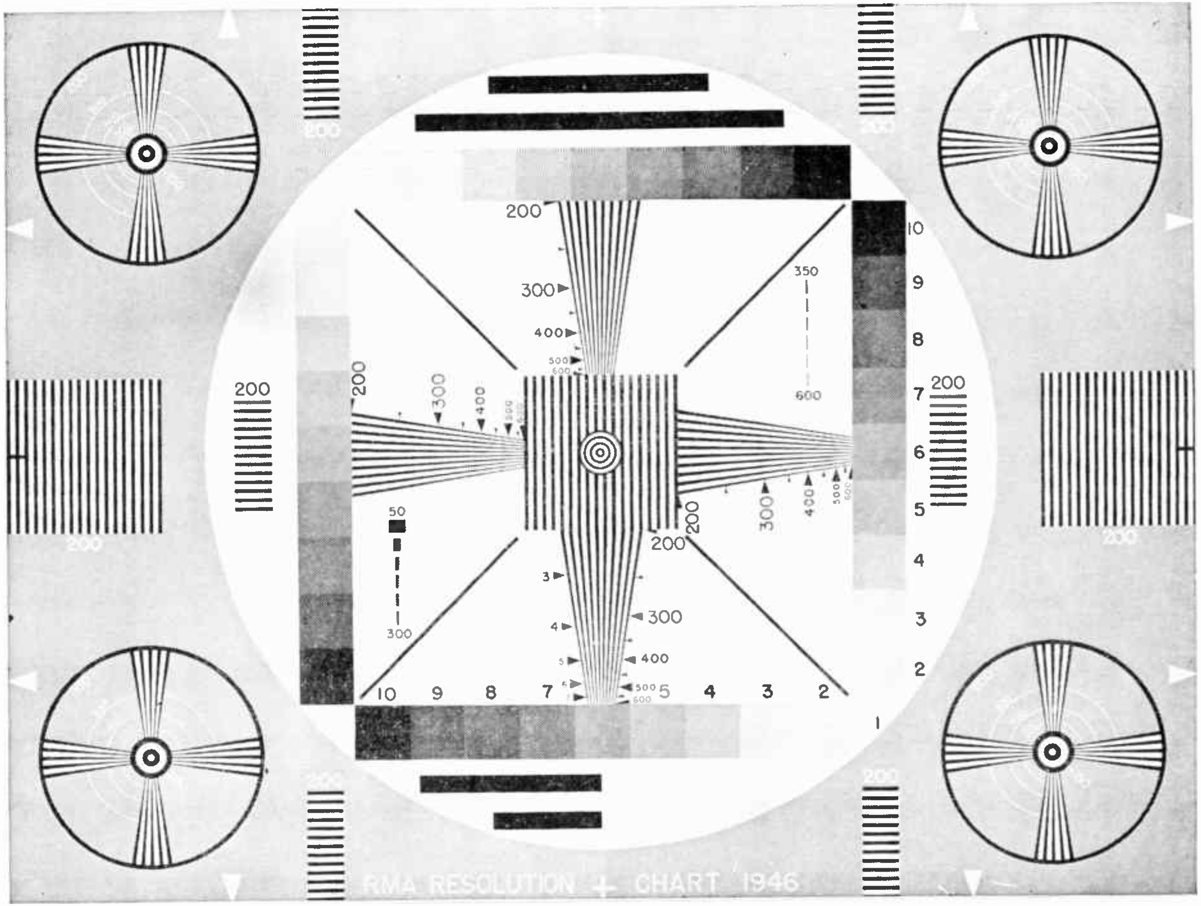


FIG. 269 RETMA Standard Test Chart

Thus, as shown at the center of the RETMA chart, a series of vertical lines converge toward a point, and it is to be noted that the numerical calibration increases as the lines become thinner and closer spaced. These lines represent resolution lines, and the system is said to have a certain numerical definition in accordance with the highest number (lines definition) at which point the separation between lines is discernible. For example, if it were possible to discern the line separation up to the point calibrated as 400, and after that the lines blended together, the system or receiver would be said to have *400-line resolution*. The horizontal resolution can be checked at the top center and bottom of the screen, as well as at the four corners of the screen, by means of the smaller circles and their associated resolution wedges.

The horizontal resolution wedge at the bottom—vertical converging lines—is calibrated in terms of lines of resolution and in terms of frequency response in megacycles. The left-hand side is calibrated in terms of megacycles and the right-hand side in terms of lines of resolution. For example, to transmit a 350-line resolution requires a bandwidth of approximately 4 1/3 megacycles. Likewise, if the receiver has a 4-megacycle response it will produce a picture with 325 lines of resolution.

The vertical resolution is judged by means of horizontal wedges at the right and left and at the four corners. The vertical definition is determined mainly by the active lines, interlace, Kell factor, and size of the picture-tube spot, determining the sharpness of the line separation as the beam gradually scans down.

In tuning the receiver, with help of the resolution chart, focusing and other adjustments are made for clean-line separation as far down the vertical wedges as possible, thereby tuning the receiver for best horizontal resolution. This procedure is used at the receiver because the vertical resolution is generally good, and determined largely by the number of active lines and the size of picture tube itself. The horizontal resolution, however, is determined by the frequency bandwidth and, therefore, the response of the r-f, i-f, and video section of the television receiver. At the transmitter, however, a compromise adjustment between best horizontal and vertical resolution is made. Best horizontal resolution is obtained with a very small-diameter camera-tube spot, increasing the number of elements along each horizontal line. So far as the receiver is concerned, however, it is more difficult to obtain good horizontal resolution compared to vertical resolution, which is an inherent result of the fixed characteristics of the television system.

CENTERING

The small arrowheads on all sides of the standard chart are used to properly center the pattern and to indicate boundaries of the chart on the scanning raster at the receiver. Thus, width and height adjustments can be made by causing the arrowhead to be located at the very ends of the scanning raster. Likewise, as is done in many small receivers, the sweep can be somewhat extended in order to obtain a larger picture of the central portions of the screen which carry the most interesting information. This is done by increasing the width and height an amount which will make the circumference of the circles in the four corners tangent with the end of the scanning raster.

In some receivers considerable width and height expansion is used to blow up center action. This can be a fixed adjustment or a switching choice. Still other receivers attempt to fill the entire surface of a circular tube by width and height expansion or intentional nonlinearity and just height expansion. The disadvantages of such systems are the limitations they impose on studio activity (already limited) and creation of atmosphere.

White crosses at top and bottom and black crosses on the sides permit ease in adjustment of optical systems of projection television receivers. They indicate exact center of all four sides of scanning raster.

ASPECT RATIO

The aspect ratio is properly set when the contrast squares within the center circle set off a perfect square. Thus, if a ruler is used to measure the height and width of the contrast square a like dimension for both horizontal and vertical measurements indicates proper 4:3 aspect ratio.

CONTRAST

The contrast squares or so-called "light scale" permit a check of the contrast performance of TV receivers. The light or half-tone scale is composed of 10 steps which vary in an approximate logarithmic manner from maximum white to approximately 1/30 this value. Actually, the number of half-tone squares chosen is greater than the performance of any of the television systems of today, and the television receiver is adjusted until as many individual squares as possible (tonal range) are visible. This is done by proper setting of brightness and contrast controls of television receiver.

LINEARITY

A series of small horizontal lines at top, center, and bottom of the chart are used to indicate vertical linearity and, if there is any nonlinearity at any point (top, center, or bottom), the lines will tend to crowd together compared to the same type lines found in other areas. If linearity is very poor the lines will blend together. Actually, the lines which make up the sections that indicate linearity represent a definition of 200 lines.

A series of vertical lines at left, center, and right are used to check horizontal linearity. A dissimilarity in vertical-line groups at left, center, and right indicates crowding or stretching of picture horizontally and therefore presents an indication of horizontal linearity. Again, the width of the lines represents 200 lines of horizontal resolution.

FREQUENCY RESPONSE

The heavy black lines at the top and bottom of the center circle are used to check low-frequency response. Poor low-frequency response causes streaking to follow after a long black interval as the beam scans across the screen. Thus, if the low-frequency response of the receiver is poor, streaking occurs to the right of the black bar. What is actually occurring can be seen in Fig. 270, which shows that actually each horizontal line that makes up the heavy black bar represents a relatively long duration squared wave (white to a long black interval and back to white again), and if the low-frequency response is poor

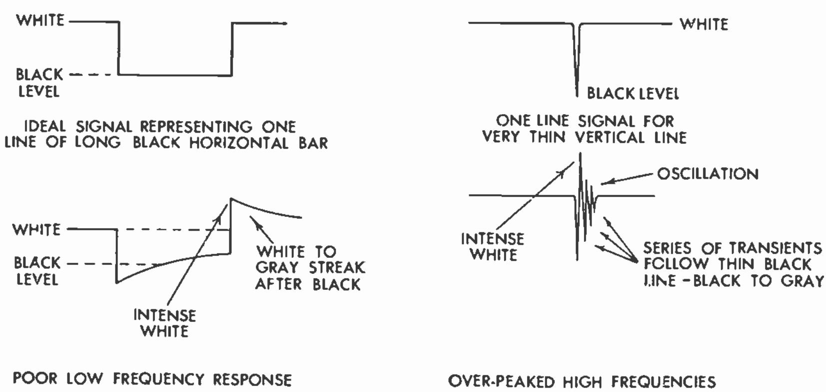


FIG. 270 Frequency Response and Its Effect on Portions of Picture Signal

the squared-wave signal which represents the line will be differentiated. Thus, the top or flat top of the squared wave which represents the black bar slides down or tilts, and its light range shifts from black to a dark gray. However, the television receiver is generally adjusted at a brightness and contrast setting which brings the black level at a lower amplitude. Therefore, the entire black bar reproduces as a black line. The streaking is caused by the negative sweep which follows and first drives a good deal beyond the actual white level, causing a saturated white to be developed which gradually declines into a light gray as the beam progresses farther to the right. Thus, a white tail follows after the black bar which gradually fades from white to gray. The various lengths of the horizontal bars determine the severity of the low-frequency deficiency, it being more noticeable on the longest black bar (longer duration of the top of the pulse which represents the transmission of the black line).

OVERSHOOT AND TRANSIENTS

It is to be noted that at the top right and lower left portion of the center circle short isolated vertical lines are present, calibrated at the bottom from 50 to 300 and at the top from 350 to 600. These isolated lines, although they can be used to check resolution, produce a clear indication of any resonant rise or oscillation in the television system. As far as resolution is concerned, the limit of the response is indicated at the point where the succeeding and supposedly thinner lines blend together to form lines of approximately the same width. Thus, if the limit of the resolution of the system is 350 lines, all the isolated lines from 350 to 600 would appear with approximately the same width (not necessarily same blackness).

The most important function of these isolated lines is to show presence of transients in the receiver. For example, if in the video amplifier section there is a decided hump (resonant rise and high Q circuit), the very sharp change in voltage represented by such an abrupt change from black to white and back to white again would, if applied to a high Q resonant circuit, cause it to generate a singe transient or a series of damped oscillations (Fig. 270, second drawing). These oscillations would generate a series of shadows or echoes behind the isolated line which represents a frequency of transmission nearest to the resonant frequency of the high Q circuit. Therefore, a series of black, vertical lines would follow immediately behind the first line, all with a gradually decreasing brightness as indicated by the decreasing amplitude of the transient oscillation in the drawing. If resonant rise is not severe and just causes a single transient alternation to be generated, a sharp white line would follow immediately in back of the black line—this is commonly called a *white-following-a-black* and is an indication that a transient is present but not severe enough to generate a series of damped waves.

INTERLACE CHECK

The diagonal lines in the center square are used to check interlace. If the interlace is not correct a jagged line appears because of the displacement of the even-numbered lines with respect to the odd-numbered lines. The cause of the jagged lines is indicated in Fig. 271, which demonstrates in a few lines

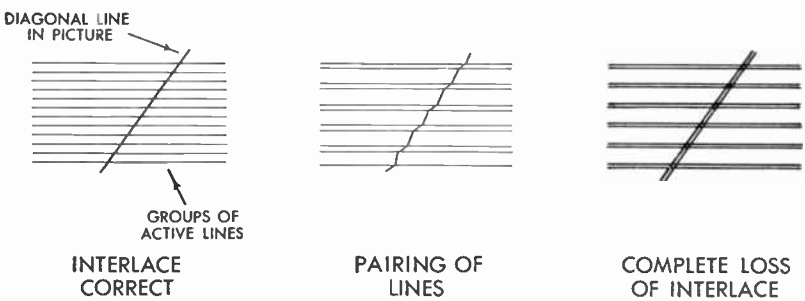


Fig. 271 Effects of Interlace on Diagonal Line

what would happen if even-numbered lines in the first drawing were raised up to a point where they did not fit midway between the odd-numbered lines. As shown in the third drawing, if the odd- and even-numbered lines coincide the diagonal lines again interlace. Thus a complete loss of interlace is indicated by a serious loss of resolution (widening of lines) and a less serious loss of interlace produces a jagged diagonal line.

VARIOUS OTHER CHECKS

Resolution and proper focus cannot only be determined at the center of screen but also at the four corners of the chart. The central resolution wedges at the apex indicate a maximum resolution of 600. With present standards only a 300 to 350 maximum is obtainable. Resolution is considerably poorer at the ends than at the center of the screen, and by means of the outer circles the extent of the degradation can be shown.

The resolution circles at the center of the screen and in the center of the corner wedges can be used for testing spot ellipticity. If the picture-tube scanning spot is circular the concentric circles which make up the circular resolution chart will be uniform; if the spot is not, the width of the lines which make up the circle will be nonuniform.

The gray background of the chart produces a signal with an average brightness very similar to the usual television studio set. Therefore, if the receiver is adjusted for peak performance with the chart it should perform satisfactorily and should be pleasing to the eye for the average studio show without receiver readjustment. The charts used today are dominantly white. It is often necessary to readjust brightness and contrast when the studio show begins, to obtain the most satisfactory picture.

At present the commercial television stations transmit charts which do not contain the abundance of test markings present on the RETMA chart. The transmitted charts are not so critical of the performance of the television system. A good receiver will completely resolve the segments of the horizontal resolution wedge since it is calibrated to cut off at a point corresponding to about 3.5 megacycles. Gradation scales are limited to about five tones.

LINE RESOLUTION AND MEGACYCLE BANDWIDTH

It is customary to speak of picture resolution with respect to the number of lines of resolution instead of frequency response of the system necessary to transmit a picture with the resolution of that number of lines. The association between the response and number of lines is not always clear. Likewise, the association between the number of active lines which make up a picture and the lines of resolution is not immediately apparent. The association between these various aspects can be more clearly presented with a series of step-by-step considerations.

The standard RETMA chart is 18 inches high, and therefore, assuming an aspect ratio of 4:3, the chart also has to be 24 inches wide. The basis of our calculations will be in relation to this 18-by-24 chart, but it is to be stressed that, regardless of the size of the original chart, if aspect ratio of 4:3 is retained throughout the system, the same frequency-response requirements and number of lines of resolution will be imposed. For example, the lines of resolution on the large chart will impose the same restriction on the reproduction of the chart by any size of picture tube. For example, the same frequency response is required to reproduce the picture on a 10-inch tube as on a 20-inch tube, because all elements of the picture, including the size of the beam and its velocity, have been reduced a corresponding amount.

As an example, let us assume a line which is 0.075 inch in width is positioned on our standard chart, which is 24 inches wide. Let us assume this is a series of black vertical lines interspersed with white vertical lines of the same

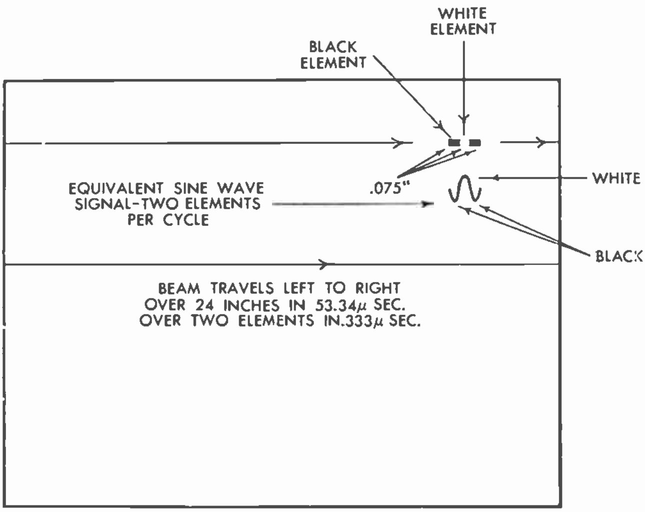


FIG. 272 Scanning of Three Equivalent Elements of a Line of Picture

width. This means that one high-frequency cycle is generated as the beam scans from a white line to a black one and back to a white one again. This can be compared with a sine wave that is going from minimum to maximum and back to minimum again.

Thus, the time required to scan over these three lines representing two complete picture elements is actually the period of the sine-wave signal which represents change in brightness when transmitting these elements. In reproducing this change from white to black and back to white again the beam scans over a width of two elements, and if the time required to cover these two elements is calculated the frequency of the representing signal can be found.

If the chart is 24 inches wide and it takes the beam 53.34 microseconds (active horizontal-trace interval) to scan 24 inches (Fig. 272), the time required to cover two elements along this line (two times the line width of 0.075) can be easily calculated as shown:

$$\frac{24}{53.34} = \frac{0.15}{\chi} \qquad X = 1/3 \text{ microsecond}$$

The frequency of the comparing sine wave, therefore, is 1 over the period, or, 1 over 1/3 microsecond equals a sine-wave frequency of 3 megacycles. The frequency response necessary to convey a series of lines 0.075 inch wide on a 24-inch chart is therefore 3 megacycles. Now, if all parts of the chart were reduced a corresponding amount, let us say to 12 inches, the number of lines of resolution and frequency response necessary would remain unchanged, because both the width of the lines and velocity of the scanning would now be reduced by a factor of ½. It would therefore take the beam the same time to scan the very same two elements in the smaller picture. It is important to understand that the same frequency response is necessary to reproduce a given number of lines of resolution on a small picture tube of 10 inches as compared to a large-screen picture of 20 inches or larger.

The number of lines used as a gauge of system resolution is always with respect to the number of lines of equivalent width which could be placed vertically from top to bottom of the screen. For example, if the line width of 0.075 inch on the RMA chart is divided into the height of the chart of 18 inches, the total number of lines of resolution is 240, representing the total number of these lines that could be placed on the chart in the space from top to bottom. Again, it is interesting to point out that if the size of the chart is reduced by a factor of ½ the width of each line is cut in half and therefore the same number of lines fits from top to bottom of the screen.

If a line width of 0.0562 inch is used on the RETMA chart it requires a frequency response of 4 megacycles to convey this line. This represents a total number of lines of resolution of 320, or 18 inches divided by 0.0562 inch.

In summation, it is the width of the line, whether it is a vertical, diagonal, or horizontal, which determines the frequency response necessary to convey that line in the case of horizontal resolution and, in the case of vertical resolution, it is determined by the number of active lines and the picture-tube spot size. Nevertheless, the number of lines of resolution is represented by the number of these lines which can be traced vertically down the screen, whether they are horizontal or vertical lines. The maximum lines of resolution horizontally is determined by the frequency response of the system and is limited to not more than 350 lines, which requires a bandpass of 4½ megacycles. It might be expected that the maximum lines of resolution vertically would be the number of active scanning lines. However, this is not the case because of finite spacing between individual lines and position of image detail with respect to scanning line. If the spot size is large without having one line overlap its adjacent lines it might be expected that the maximum lines of resolution can approach the number of active lines. However, the spot is not of uniform brightness, and sharp gradation from black to white or white to black vertically depends on the position of image change with respect to center of the scanning line. The vertical resolution is often in excess of 300 lines but does not reach the total number of active lines because of spacing between those lines. A near enough approximation is 64 per cent of the number of active lines.

137. *Pre-Installation Checks*

It is a good practice to assemble the components of the television set and check its performance before it is delivered to the customer's home. Some manufacturers who deliver their receivers from a central distribution point for all dealers pre-check the receiver at the center before it is delivered; all should. Others deliver equipment as it is received from the production plant and assemble and check the receiver in the customer's home. Most dealers, however, who deliver sets right from their shops, check the complete set before delivery to reduce the number of call-backs and speed the installation.

The manufacturer's procedure should be followed to the letter with respect to the mounting of the picture tube and the placement of the deflection coils, as well as the focusing and ion-trap assembly on the neck of the tube. It is important that the service shop have an ideal antenna installation which can receive satisfactory noise-free signal from all stations in the vicinity. Thus, any defect which shows up in the receiver is not caused by the antenna system and can be isolated to the receiver under check. It is particularly important that the microvolt signal strengths are known for signals in the vicinity and also that the performance of representative models of the receivers are known on each of the signals. Thus, if an insensitive receiver is placed in position the defect will immediately show up. At present there seems to be a considerable variation in receiver sensitivity of like models. Sensitivity is a very important factor in the fringe areas of the station and in locations where antennas are crowded or noise level is high.

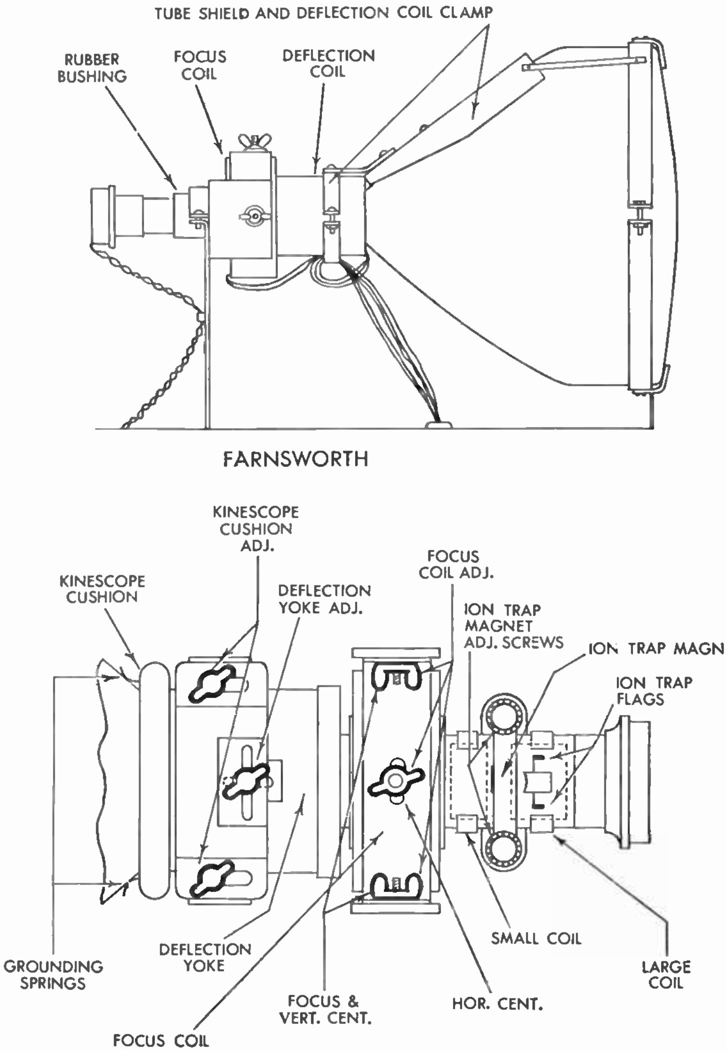
In checking the performance of the television receiver, the transmitted station chart will be of great importance because it indicates many characteristics which the receiver should have.

The many installation adjustments and performance checks that can be made during the pre-installation procedure are: Signal strength, receiver sensitivity on all channels, aspect ratio, resolution, linearity, picture width and height, brightness, and contrast. The receiver should be left in operation for a number of hours and its components should be checked for overheating or potential breakdown. In addition, the stability of the receiver should be checked during this period, and if possible the performance of receiver under noise reception should be tested. Many receivers of like models have varying susceptibilities to noise and the dealer should know how various models should react upon reception of noise.

138. *Picture-Tube Installation*

The modern picture tube, although it is many times more rugged than the early models, should still be handled with care. It must never be forced into its mounting nor should undue strain be placed on the neck of the tube. Danger from implosion when the front of the tube is jarred sharply is still present although the severity of the implosion and the likelihood of its occurrence has been very much reduced by rugged design. Nevertheless, it is advisable to wear shatterproof goggles and heavy gloves when handling picture tubes. Whenever the tube is not in its mounting or is being prepared for mounting, place the tube face down on a soft pad to prevent rolling or scratching of the front surface.

In some receivers the deflection and focus coils are a part of the picture-tube mount, and it is necessary only to slide the picture tube into position;



R.C.A.

FIG. 273 Typical Picture-Tube Mountings

while in other models the tube is placed face down and the coils are slipped on, after which the tube and coils are fastened to the mounting assembly. Two typical mounting arrangements are shown in Fig. 273. The deflection yoke is generally placed over a rubber cushion and both are slipped on the tube, or the tube is slid through them, until the cushion and yoke are as far up the neck of the tube as they will go. Focus coil is next in order, and it is mounted on a smaller cushion or corrugated paper and is slid into position, resting against the deflection-yoke cushion. Ion-trap assembly, if used, is the final coil mounted on the neck of the tube. It must be placed on the tube in proper relation to the screen and gun; therefore, the manufacturer provides some sort of guide or flag to indicate correct positioning. No ion-trap assembly is necessary (second drawing, Fig. 273) when an aluminum-backed screen is used.

Inasmuch as the outer coating on the picture tube forms the ground side of a high-voltage filter capacitor, it must be grounded by the mount assembly either by means of grounding springs or clamps. The shields surrounding any of the other coils prevent stray pickup and interference and must form a part of the same grounding system (Fig. 273).

The ion-trap assembly used must not only be placed over the gun at the proper point but must be positioned on the tube with the weaker magnetic field nearer the screen end of the tube. In the RCA picture tube, ion-trap flags mounted internally on the second section of the electron gun (Fig. 274) have been provided to indicate the correct position for ion-trap coil. The coil is slipped in position over the flags with the smaller coil positioned nearer the screen. When permanent magnets are used to generate the ion-trap field (second drawing, Fig. 274) the larger magnet is mounted nearer the tube base.

A typical mounting procedure would be as follows:

1. Insert the tube into mounting after placing coils in the proper position as recommended by the manufacturer. Second-anode contact for most picture tubes is a recessed metal well in the bulb, and the tube must be positioned properly to permit application of second-anode lead. For example, in many RCA receivers the tube is inserted with the contact approximately on top. The final position of the tube is determined by location of ion-trap flags. These metal flags are placed to appear as shown in Fig. 274 when looking down on the chassis.

2. Insert the tube through deflection and focus coils until the tube base protrudes an indicated distance beyond focus coil.

3. Slip the ion-trap assembly on neck of tube and place it over flags with stronger magnetic field (larger coil or magnet) near base of the tube. When a permanent magnet is used the gaps in the two magnets can be moved readily if the gap assembly is pulled apart slightly with the fingers.

4. Connect the picture-tube socket to the tube base. Center and fasten the entire assembly properly with respect to the viewing mask.

5. Slide the cushion firmly against flare of the tube and slide deflection yoke as far forward as possible. Connect high-voltage leads.

6. Turn on the receiver. Set contrast to minimum and brightness full on. Move the ion trap forward and backward and at the same time rotate it slightly until raster is brightest. Fasten in position firmly.

7. Reduce brightness to average level and focus for clearest line structure.

8. Turn centering controls to mid-position. If a corner of the raster is shadowed, rotate the focus coil about its axis until entire raster is properly centered and free of shadows. When permanent magnets are used as ion traps it is at times necessary to shift the position of the front magnet slightly with respect to the rear one to obtain a shadow-free raster.

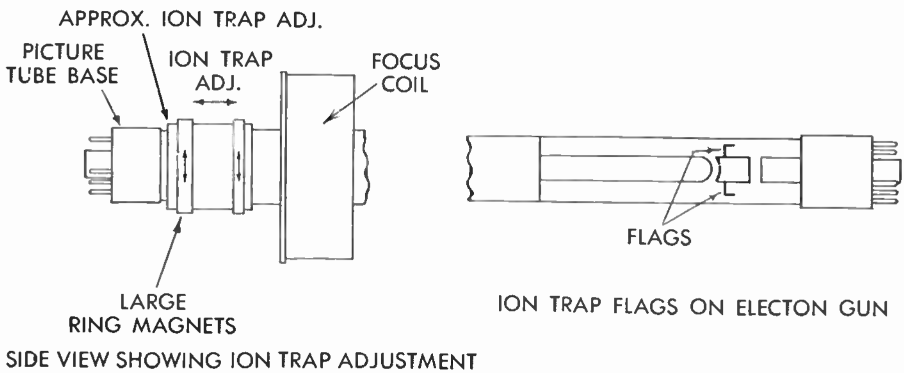


FIG. 274 Ion-Trap Positioning

9. If the lines of the raster are not squared with the picture mask, rotate the deflection coils until this condition is obtained. In some receiver models no centering controls are incorporated and positioning of focus and deflection coils must be made carefully to ensure proper centering of the raster.

The mount positions and adjustments for a General Electric assembly are shown in Fig. 275. The shielded deflection yoke is mounted at the screen end of the picture-tube neck. There are two yoke adjustments—one is a centering tilt lever that permits positioning of the raster correctly with respect to the mask, and a second adjustment is a small trimmer capacitor that permits proper balancing of the yoke to eliminate wiggle and motion of the horizontal sweep lines. The focus coil is in the second position on the neck of the tube, and it also has two adjustments. A set of three wing nuts can be adjusted to move the focus coil up or down and right or left to center the picture properly, while a second wobble-plate lever is adjusted until the tube-neck shadow is removed from the raster.

The ion trap is mounted near the base of the picture tube and is, of course, adjusted for maximum brightness in accordance with standard procedure. The illustration also shows the electrode connector, picture-tube sockets, and deflection-yoke plug, as well as the various mountings and strap assemblies

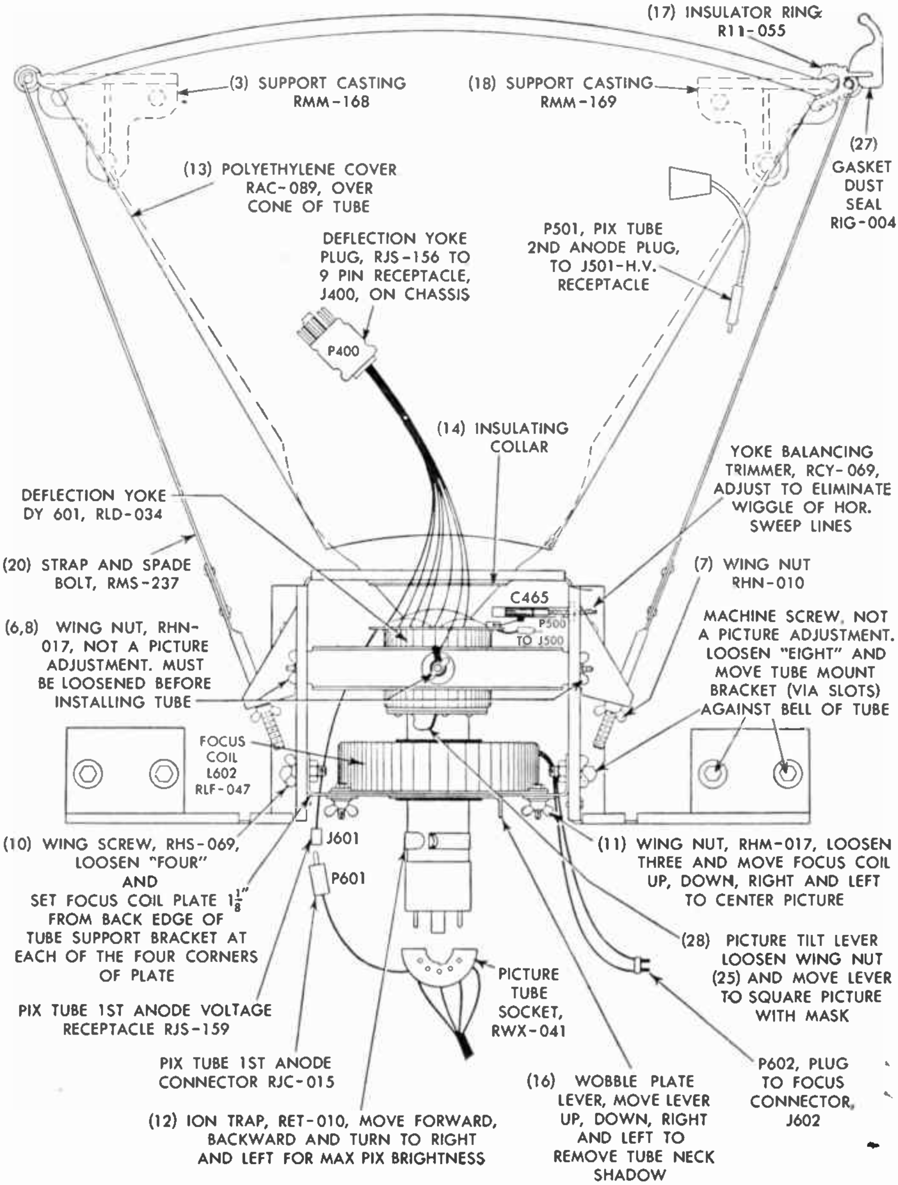


FIG. 275 G.E. Picture-Tube Mounting Assembly

that hold the picture tube firmly in place. A typical Sylvania mount for a large-screen picture tube is shown in Fig. 276. Notice that the picture tube has a slight forward tilt. This arrangement minimizes the effect of room glare, and reflections from bright objects in the room are directed downward instead of toward the viewers' eyes.

When a large picture tube is used, it is necessary to correct the raster for pincushion defect at top, bottom, and sides. Consequently, corrector magnets are used near the deflection yoke to remedy this disturbance. Large-screen picture tubes with wide-angle deflection are most subject to pincushion disturbance, the sides of the raster bowing-in, because of the much greater distance over which the beam travels in order to reach the extremities of the raster and its nearness to the edges of the deflection field in its passage. The fixed magnets pull out the beam at these extremities and, when properly adjusted, correct exactly for the bow in the top, bottom, or sides of the raster.

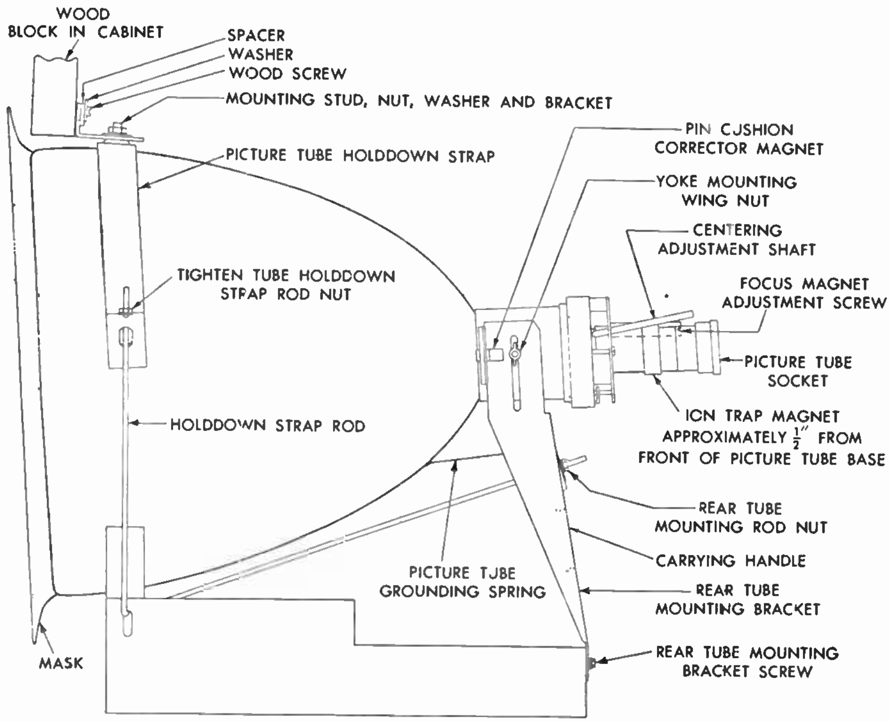


FIG. 276 Sylvania Picture-Tube Mount

The complete positioning and adjustment procedure for a typical Sylvania picture tube is as follows:

The ion-trap magnet, focus-magnet, and centering-shutter adjustments are interdependent and must be checked at the same time.

Before any adjustments are made, the function of each magnet should be noted as follows:

1. The ion-trap magnet is used to obtain maximum brilliance of the raster or picture and should be adjusted as described below.

2. The focus magnet is used to obtain correct focus of the picture.

3. The centering shutter is an integral part of the focus-magnet assembly. Its function is to position the picture both horizontally and vertically.

Before making any adjustments, check to assure that the deflection yoke is positioned in such a way as to press against the flare of the picture tube. To ensure this, loosen the wing-fasteners located at each side of the yoke, and push the yoke as far forward as it will go. If the picture is not square with the screen mask, rotate the yoke and then tighten the wing-fasteners.

Next, check to see that the focus magnet is firmly held in position.

In adjustment of the focus of the receiver, it is to be noted that optimum focus of the picture is not necessarily attained when either the vertical or horizontal definition is adjusted to maximum. Optimum focus is frequently a compromise between these two settings. It is highly desirable, therefore, that a transmitted picture, containing both vertical and horizontal lines, be available for correct focusing of the receiver.

Make sure the ion-trap magnet is correctly adjusted before proceeding with the receiver-focus adjustment.

Adjust the picture control on the chassis to approximately three-fourths of maximum position. Position the ion-trap magnet on the picture-tube neck approximately ½ inch forward of the tube base. Set the brightness control on the chassis to maximum.

Do not operate the receiver longer than necessary with brightness at maximum. Rotate and move the ion-trap magnet backward and forward on the picture-tube neck until the picture or raster is visible on the screen. Continue adjustment of the trap until the greatest possible brilliance is obtained. Adjust brightness to less than normal, and readjust the ion-trap magnet for maximum brilliance.

The adjustment screw on the focus magnet should now be turned to obtain a picture which is focused—this preliminary adjustment will not be necessary if the picture is already in focus.

If the picture is not centered on the screen, properly position it by adjustment of the centering shutter, and with brightness at a low level, check to see that no corner-cutting exists.

Adjust contrast and brightness controls to obtain a normal picture, and then adjust the ion-trap magnet to obtain the highest possible brilliance level. The focus should now be adjusted to obtain the best horizontal and vertical focus, as previously mentioned.

a. In some cases optimum adjustment of the ion-trap magnet may be obtained with the magnet located on either side of the diagonal slot in the picture-tube electrode assembly; it is permissible for the magnet to be located either between the slot and the tube base or over the slot. Do not locate the magnet between the slot and the focus magnet.

b. Optimum adjustment of the ion-trap magnet may be obtained irrespective of which way it is placed around the picture-tube neck. It should be noted, however, that in some cases one way will result in a better focus characteristic than will the other.

c. Some receivers may have the facility to allow the focus magnet to be rotated. On such receivers a better focus characteristic may be obtained by rotating the focus magnet to a different angular position and again adjusting the focus screw. This will require a check of the centering shutter to make sure that there is no corner-cutting when the picture is properly centered on the screen. Recheck the ion-trap magnet, as previously mentioned, with brightness and contrast adjustments set for a normal picture. Carefully adjust the focus for the best possible compromise between horizontal and vertical focus. Since these adjustments are interdependent, recheck all three until the best possible picture is obtained.

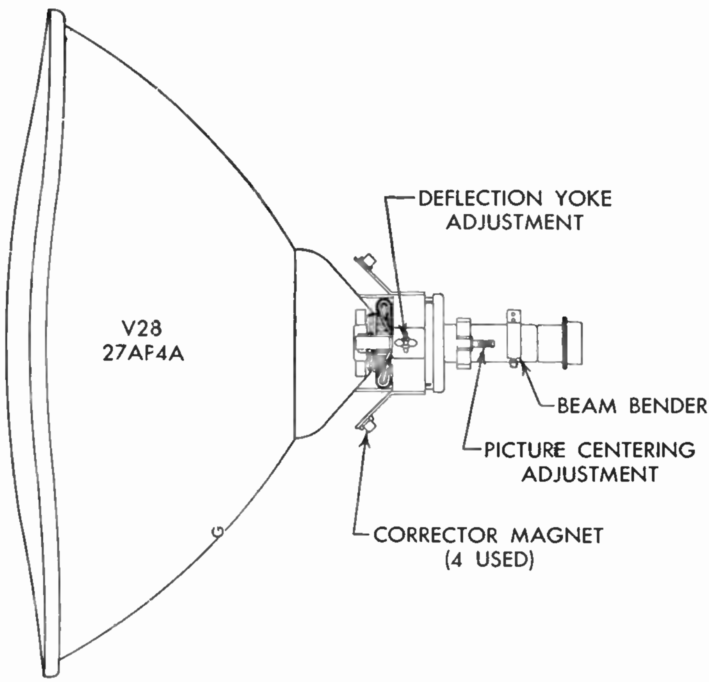


FIG. 277 Zenith Picture-Tube Mount

The pincushion-corrector magnets eliminate curvature of the edges of the raster. Move the centering shaft on the focus magnet to a position where one edge of the raster is approximately ½ inch from the edge of the picture-tube screen. Adjust the pincushion-corrector magnet on the corresponding side of the picture tube until the edge of the raster is a straight vertical line. Repeat this procedure for the other edge, using the magnet on the opposite side of the picture tube. Move the picture up and then down the screen to check the top and bottom edges. The pincushion-corrector magnets should be adjusted for the best over-all compromise.

A typical arrangement for picture-tube components on the neck of an electrostatic-focus picture tube is illustrated in Fig. 277. Inasmuch as it is a large-screen picture tube, corrector magnets are again used and are fastened to the deflection yoke assembly. In this arrangement a magnetic centering adjustment is used; it consists of a bar magnet, the position of which is adjustable with a knurled control. Centering is accomplished by adjusting this control and by the relative position of the entire assembly on the neck of the tube. It is placed approximately ⅛ inch in back of the deflection yoke. A single magnet beam-bender is used.

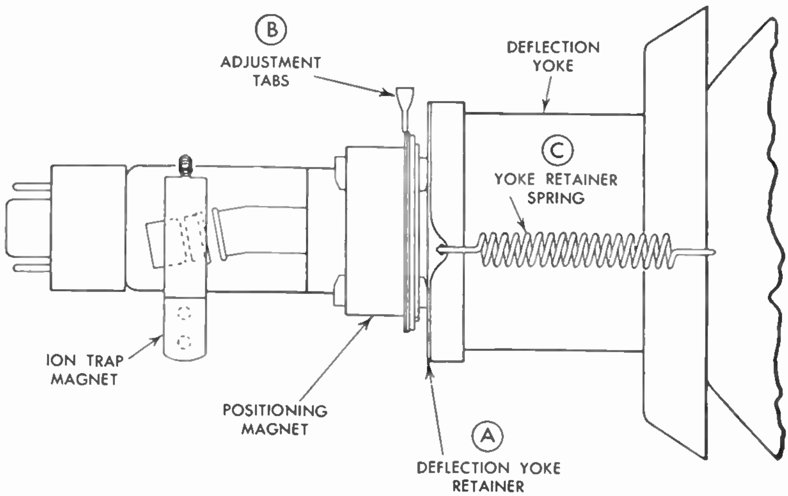


FIG. 278 DuMont Picture-Tube Mount

A simple arrangement of components is possible with the self-focus type of picture tube, Fig. 278, in which the ion trap must be adjusted carefully, because proper focus and maximum brilliance occur with the correct setting of this magnet. Consequently, brightness and line sharpness must be observed carefully as the magnet is set. An adjustable magnet is used for proper centering of the picture. Associated with the centering magnet are two adjustment tabs that can be moved in relation with each other; the entire centering assembly can also be positioned to center the raster properly at the mask.

139. *Pre-Set Control Adjustments*

The next adjustments in order are those pre-set controls which the customer does not use but which must be carefully set by the dealer if good performance on all stations is to be obtained. Typical procedures for checking pre-set controls are as follows:

1. Adjustments can be best made when a test chart is received. Turn on receiver and set "station selector" to desired channel.

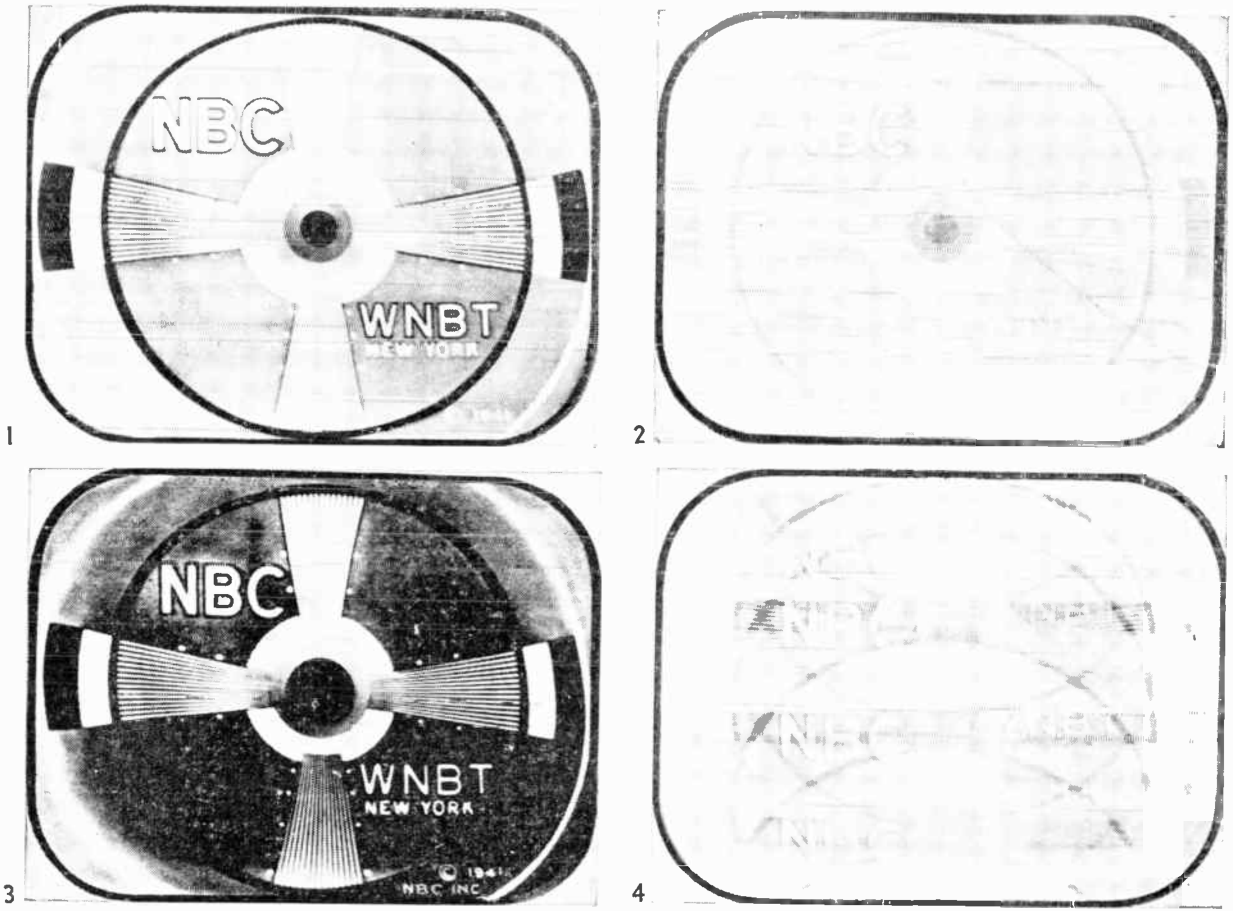


FIG. 279 Installation Test Patterns

2. After a warm-up period, begin adjustments by advancing "sound volume" to mid-position and "contrast" to minimum.

3. Turn "brightness" until a glow appears on the screen and then back off slightly just until glow disappears.

4. Advance "contrast" control until glow or pattern appears on screen.

5. Adjust "fine tuning" for best sound clarity and "sound volume" for comfortable hearing.

6. Adjust "vertical frequency" or "hold" control until pattern stops vertical movement ("flopping over").

7. Adjust "horizontal frequency" or "hold" control until pattern stops horizontal movement ("tear-out").

8. In many receivers the three controls mentioned in steps 5, 6, and 7 are not controls which the customer tunes; in other receivers, they are front panel controls. When automatic frequency control is used a "fine tuning" control is not necessary. In other receivers the two frequency controls are pre-set adjustments the technician should set very carefully to permit the best sync stability on a *weak signal.*

9. Touch up "brightness" and "contrast" settings for most comfortable viewing and best possible contrast range, as indicated by greatest number of discernible contrast squares.

Typical patterns received when the set is not properly adjusted are given in Fig. 279, the NBC station chart being used as an example. Photo 1 shows the receiver properly adjusted. Improper setting of the brightness control is indicated by photo 2, which shows an absence of illumination. "Contrast" set too high produces the overexposed picture of photo 3. Notice the center two of the contrast circles and the outer two blend together. Compare with photo 1.

Vertical displacement of pattern (photo 4) indicates improper adjustment of "vertical hold" control. Tear out of the picture (photo 5) Fig. 280 shows improper setting of "horizontal" frequency.

A received-station chart is also helpful in positioning the focus, deflection, and ion-trap coils. For example, photo 6 shows improper positioning of focus coil and consequent displacement of picture. This photo also shows improper ion-trap positioning as indicated by the shadow at lower left. The ion trap should be set for peak brightness always. When "focus" control is not set properly the entire image becomes blurred (photo 7).

When the deflection coil is improperly rotated the image does not line up with the mask (photo 8).

10. After the receiver has been on for some time, it is often necessary to readjust the "fine tuning" control for best sound fidelity. The above-mentioned controls are operating controls, except as stated in number 8, and are used by the customer in adjusting his receiver properly. The following controls are strictly pre-set adjustments and must be carefully set by the technician.

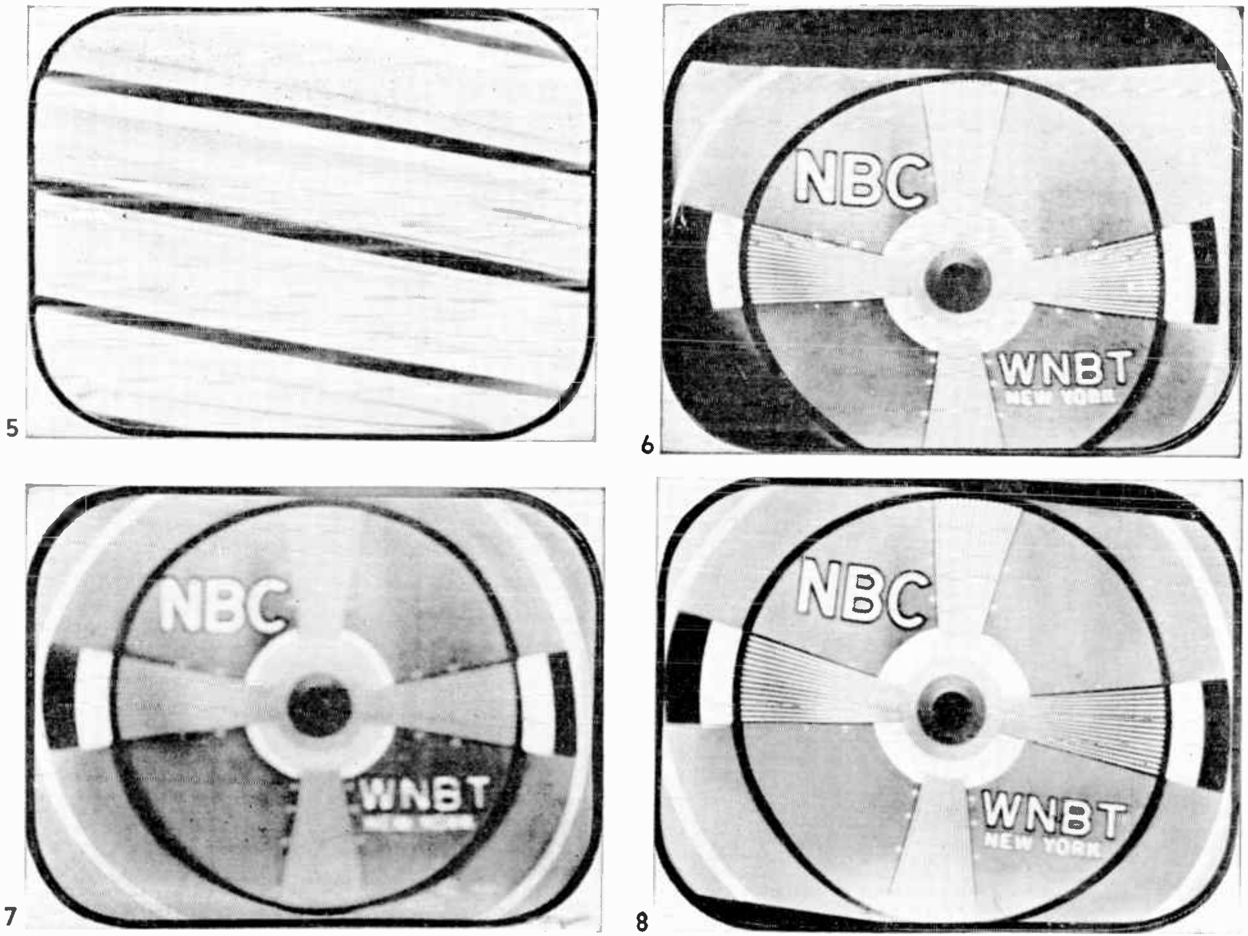


FIG. 280 Installation Test Patterns

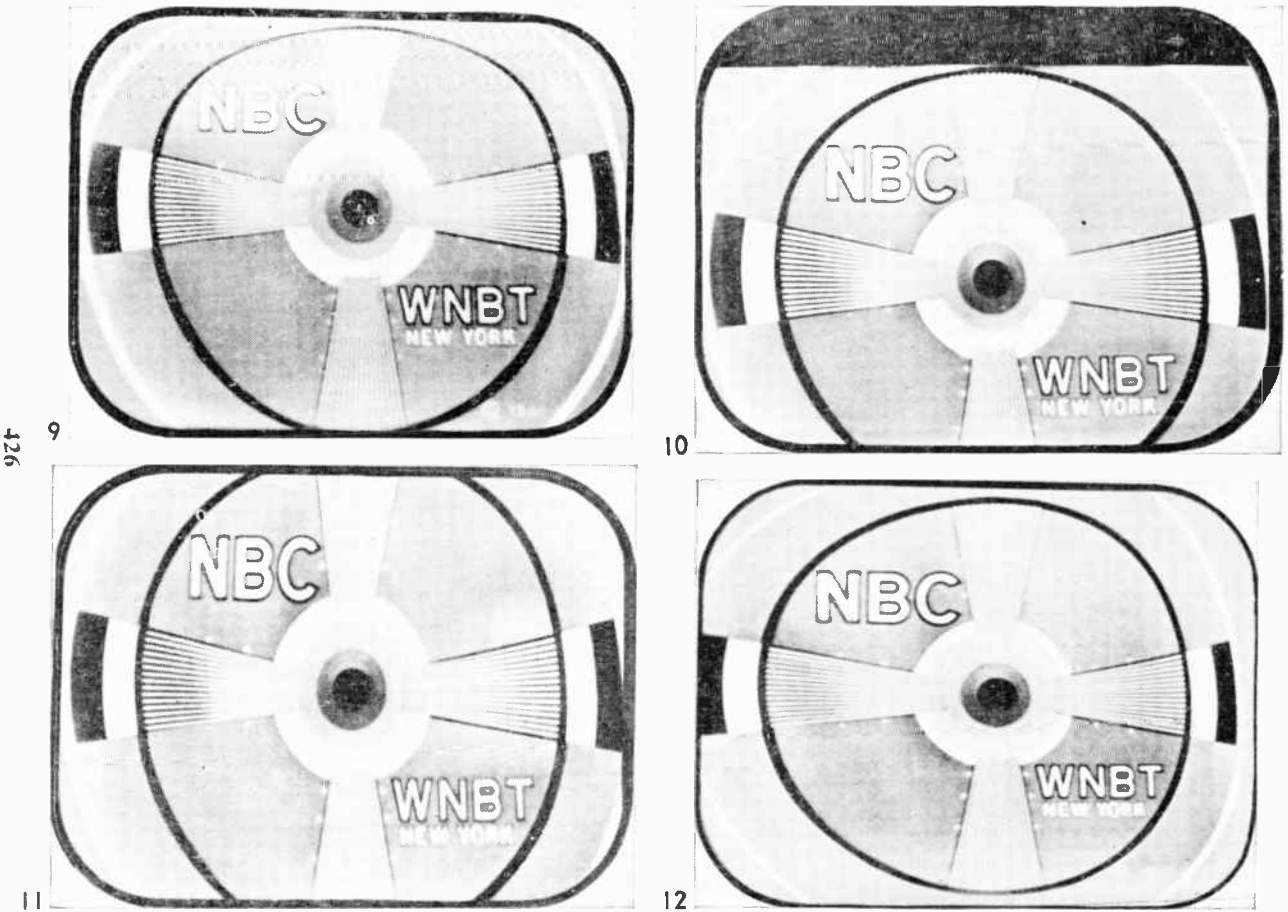


FIG. 281 Installation Test Patterns

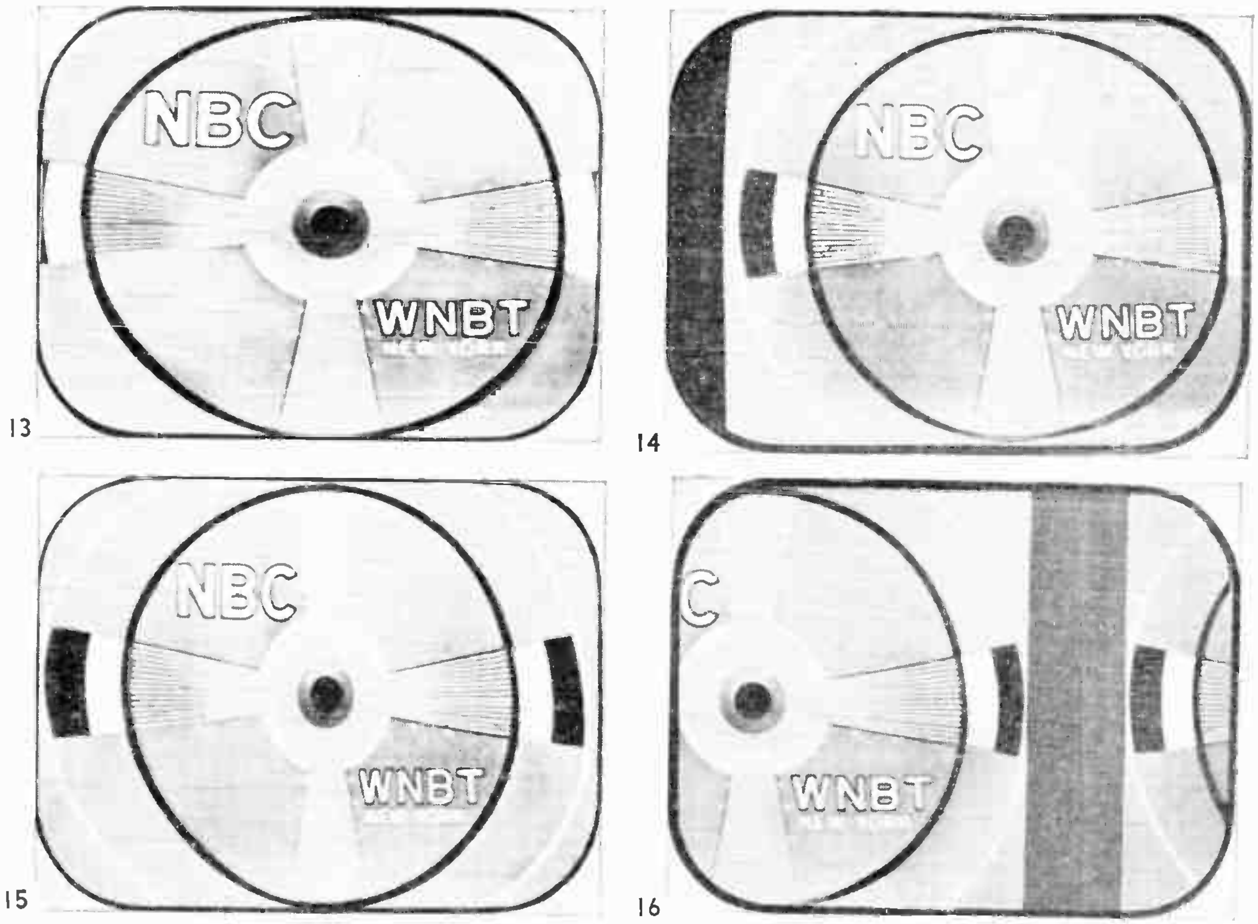


FIG. 282 Installation Test Patterns

11. Adjust "height" control until height of pattern extends to end of the mask at top and bottom. Set "vertical linearity" control for best linearity vertically (use linearity squares present in station chart or observe any large circle in pattern and adjust control until circle is not flattened at top or bottom). Distortion of the circle in the station chart by improper setting of the linearity control is shown in photo 9, Fig. 281.

"Height," "linearity," and "vertical centering" controls interact on each other, and it is necessary to readjust all three controls to find the optimum setting. Displacement of pattern vertically by incorrect setting of "centering" control is shown in photo 10. Improper height-setting is shown in photo 11.

12. Adjust "width" control until the image just fills the mask horizontally. Adjustment of horizontal linearity requires manipulation of a number of controls. One typical procedure for a magnetic deflection system is as follows:

a. Turn up "horizontal drive" control as far as possible without crowding picture at right (photo 12). "Width" control is readjusted to have the image just fill the mask.

b. Adjust "horizontal linearity" control until the image is linear, left to right.

c. Some receivers have two linearity controls in the horizontal output stage—one control affects linearity on the left, the other at the center.

Image defects created by incorrect settings of horizontal deflection controls are shown in other photographs. Photo 13, Fig. 282, shows too much sweep width; photo 14, improper setting of "horizontal centering," and photo 15, incorrect setting of "horizontal linearity."

13. When an automatic sync system is used in the horizontal, a number of additional adjustments must be made. Adjust "horizontal hold" control until picture locks in. If the horizontal blanking bar appears on screen (photo 16) it is necessary to adjust the "horizontal phase" adjustment until retrace of the horizontal sawtooth occurs at same instant horizontal blanking pulse arrives on grid of the picture tube.

14. The final preinstallation check to be made is to observe the performance of the receiver on all active channels in the area. Make certain the "fine tuning" control can tune in each station and is not crowding the end of its capacity range when it does so. If an a-f-c system is used it is necessary to precisely tune the r-f oscillator on each station. In the Philco model receivers a convenient output jack is incorporated for this purpose, measuring the d-c component of discriminator voltage. The oscillator inductor is tuned for proper reading on all channels either with received or crystal-calibrated signal generator set on sound-carrier frequency of channel.

The photos of Fig. 283 demonstrate spurious signal defects which may hamper receiver adjustments. The dual image showing in photo 2 indicates arrival of a reflected signal at the antenna. Photo 3 shows the effect of interference from near-by diathermy equipment, while photo 4 shows the effect of hum on picture such as might be caused by power-supply ripple.

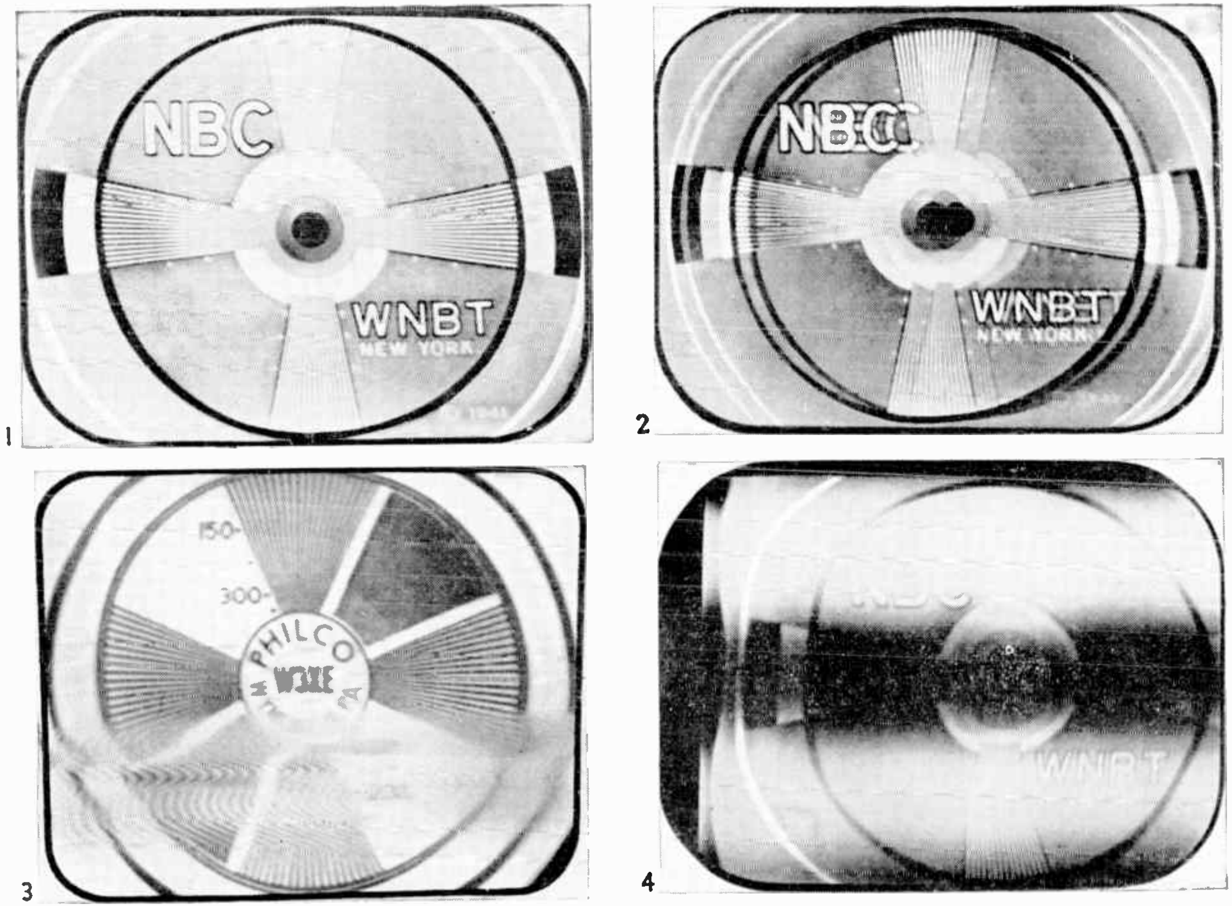


FIG. 283 Spurious Signal Defects

140. *Operating Instructions for the Layman*

If the customer is to obtain satisfactory performance from his receiver, considerable time should be spent instructing him and his family in the proper operation of the receiver. This not only benefits the customer but prevents many call-backs attributed to improper tuning procedure. Not only should the technician demonstrate the operation of the receiver thoroughly but should guide various members of the family as they go through the very same operations. A satisfactory operating and demonstrating procedure is as follows:

1. Turn the receiver on and put the channel "selector switch" on the channel to be received. Show the customer the positions of the various stations which can be received in his area. Demonstrate the procedure of going from station to station. Turn "contrast" control fully off and turn up the "brightness" control until a visible scanning raster appears on the screen. Then back off the "brightness" control until illumination just disappears. Vary the "brightness" control through its entire range showing the customer the effect it has on presentation of the illuminated area of the picture-tube screen.
2. Advance "contrast" until presentation appears on the screen.
3. If "fine-tuning" control is used vary it until the picture contains the greatest amount of black. Again demonstrate to the customer, showing him when the "fine-tuning" control is not properly set and what to expect when "fine-tuning" control is set at the proper point. The "fine-tuning" control in most receivers is adjusted for best sound reproduction, though occasionally there is only a limited variation of tuning over which the sound is satisfactory and, at times, a better picture can be obtained by closely observing the fluorescent screen until the picture has the best resolution.
4. Now readjust the "contrast control" for the most satisfactory picture. Demonstrate to the customer the improvement in picture which can be obtained by slightly readjusting the "contrast" and "brightness" controls for the most presentable picture. Demonstrate the tune-up procedure on the program material and also on the station chart. If station chart can be received, demonstrate the use of resolution wedges in adjusting fine tuning and the use of contrast squares for contrast and brightness settings.
5. If the focus control is a front-panel adjustment show the effects of improper setting of "focus" control on the reproduced picture and how the resolution wedge in the test pattern can be used properly in setting this control. When the front-panel "hold" or "frequency" controls are a part of the receiver, demonstrate their functions in obtaining a stationary pattern on the screen. Show the effects on the reproduced picture when one or the other or both of these frequency controls are out of adjustment.

After the operating procedure is demonstrated have the customer and members of the family go through a similar operating procedure. Advise them of any mistakes or improvement which can be made in their methods. Finally leave thorough and complete operating instructions with the customer and impress him with the importance of using them until he is familiar with receiver operation.

If at all possible, demonstrate the operation of the receiver when there is noise interference to show the effects of such interference on the reproduced image, particularly ignition noises or any other sparking noises prevalent in that area. Finally, go through operating procedure once more and at the same time show the appearance on the screen when any one of the controls is not properly set.

141. *Antenna Types*

The choice of antenna type, design of the antenna, and quality of the component parts of which the antenna is constructed is of utmost significance in obtaining good receiver performance in weak-signal areas, noisy locations, and for extended-range reception of television signals. In weak-signal and extended-range areas the choice of antenna type and the quality of that antenna are the difference between a satisfactory signal and no signal. It has been the author's experience, in one long-range test where two antennas of exactly the same type were used, that one antenna received the signal and the other did not because of the quality of the insulation that isolated the two sides of a folded dipole. This demonstrates the importance of the dielectric material which is used to insulate some portion of the antenna from other sections. Of course, the same two antennas positioned within the primary area do not produce any noticeable difference in signal strength, but out in the fringe areas of reception it means the difference between receiving signal and not. The choice of antenna and the antenna type is dependent on the following factors: direction and range of the station or stations to be received, channel frequency of the station, noise and interferences prevalent in the area, space-loop distribution, reflections, and relative signal strength. Thus, there is no definite rule to be laid down in the choice of an antenna and in most conditions compromise tactics must be employed. Let us treat each factor as a separate subject and indicate the influence the other factors have on optimum choice of antenna for that factor.

CHANNEL FREQUENCY AND SPACE-LOOP DISTRIBUTION

The antenna type, and particularly its dimensions, are dependent on the distribution of channels in a given area. The farther separated the channel frequencies are (for example, channel *2* and channel *13*, which are widely separated), the more difficult it is to obtain a single antenna which will perform satisfactorily on all channels. If channels *2* and *22* were assigned in your area a satisfactory antenna for one channel (54 to 60 megacycles) would not always be satisfactory for the other (518 to 524 megacycles), and either one signal or the other would suffer in respect to signal strength delivered to the input of the receiver.

Antenna dimensions are a function of the channel frequency allocations in the area and the relative signal strengths of the stations received in your district. Still another factor is the space-loop distribution of these signals in the immediate vicinity of the antenna erection site. Stations can be definitely favored by correct positioning of the antenna at a space loop. This is a particularly important factor when trying to obtain the very best signal and performance possible from a normally weak station.

RANGE

Locations within service area of the television stations generally fall into one of the following categories:

1. In a strong-signal area of all stations any type antenna may be purchased (such as a simple mismatched dipole) to deliver sufficient signal for all stations (if the antenna is cut for any frequency in the low-frequency set of channels). In fact, in strong signal areas in which all stations deliver a substantial signal, a simple antenna would permit reception on all channels. The one stipulation which applies in this case is that the stations would have to be in the same general direction, or at least in a general direction broadside to either side of the dipole elements.

2. In a strong-signal area where perhaps one or two of the stations are delivering a substantially weaker signal it is advisable to cut the antenna for the weaker station and to direct it for maximum signal strength in that direction. Thus, the antenna would be most sensitive to the weaker station and still would receive substantial excitation from the stronger stations on differing frequencies.

If all stations are delivering approximately the same signal strength to a given location the antenna is often cut to the low-frequency side of the extremities of the frequency allocations in that areas. For example, if channels *3* and *12* were the extremities for a given area the antenna would be cut for a median frequency between 60 and 66 megacycles, being tilted to permit reception of channel *12*. If signal strength in the area favors a high- or low-frequency station it is advisable to cut the antenna to a frequency somewhere near the end of the frequency spectrum for the weaker. Likewise, if possible, the antenna could be directed toward those stations which deliver the weaker signals.

3. In suburban districts where signal strength is still substantial, stations are most often found in the same general direction, and it is at times possible to purchase a so-called "broadband antenna" which has a wide acceptance band and delivers to the receiver substantial signals from all stations. Otherwise same considerations apply as per above example.

4. In weaker signal areas where most signals are weak and perhaps one or two stations deliver a medium or strong signal, a broadband antenna with a high directivity in the direction of weaker stations is satisfactory but oftentimes very elaborate. Another requisite, of course, is that the weak stations must be in same general direction from the receiving location. If signals are weak and

the stations are not in the same general direction, more than one antenna must be employed. This expedient does not necessarily make the antenna installation unduly elaborate; the antennas can be stacked on the same mast with the higher frequency antenna at the top. In most cases two separate antennas will suffice and only occasionally is it necessary to use as many as three. Generally, one antenna cut for the mean frequency of the low-channel stations and a second cut for the high-frequency channel stations perform satisfactorily when directed toward weaker stations. The stronger station or stations will deliver satisfactory signal although in a direction which might not always be the most sensitive angle for the particular antenna. Be certain to position the antenna at the space loops of the weak stations.

In all weak signal areas it is advisable to cut the antenna dimension as near as possible to the dimensions which indicate optimum performance for a given channel or channels and to make the antenna as sharply directional as possible (parasitic directors or reflectors). Also reduce the sensitivity of the antenna to noises arriving from other directions to increase signal-to-noise ratio (parasitic elements and stacking).

DIRECTION

A confusing problem, which in many cases can be answered only by actual tests, is the choice and orientation of the antenna which is to receive signals from a number of directions. The problem is not too difficult in strong signal areas and can be best decided by actual test. The antenna to be used is either a wideband antenna or one which has been cut to give satisfactory performance on the channels allocated within the area. The antenna itself, after it is positioned on the building, is oriented to give satisfactory performance on all stations and at the same time have best rejection of noise and absence of reflections. Thus, it is more or less of a cut-and-try method to find the correct point of positioning and orientation.

If any particular station is considerably weaker than others in a given location, the antenna is positioned, cut, and directed toward the weaker station and performance is checked for the other stations in the area. In case the antenna is found unsatisfactory on one of these stronger stations because of insensitivity or reflections in that direction, the antenna is shifted slightly until better performance is obtained without sacrificing too much sensitivity in the direction of the weaker station. In using test tactics, particularly in built-up areas, changing the antenna's position a short distance occasionally has an appreciable effect on signal distribution. An increase in antenna height under most circumstances means an increase in signal strength. However, in some few locations a lowered antenna will still maintain line of sight and give improved performance in the presence of surrounding buildings or objects, space pattern, or a change in the vertical angle of directivity of the antenna (number of wavelengths above the ground). Medium signal strength areas require more careful choice of the antenna type, its position, and orientation, particularly if the antenna is erected in a noisy district. Fortunately again, most stations are located in the same general direction, and one good antenna (properly matched to the transmission line, and the transmission line matched to the receiver) will give good performance if it is a wide-band antenna (broad and/or with harmonic relations) or cut for a mean frequency, preferably favoring the weaker stations. If stations are not in the same general direction an experimental approach can be used in the orientation of the antenna, favoring the weaker station again.

If a single antenna does not produce satisfactory performance a dual-antenna installation must be made, both antennas mounted again on the same mast (with high-frequency antenna at top and each antenna directed toward the station or stations for which it has been cut). Two such separate antennas, if mounted on the same mast, are generally spaced at least a half-wave at the lowest frequency to be received. There is no objection to using two entirely separate antennas on separate masts except for the additional elaborations necessary.

In weak-signal areas antenna design and choice are extremely important if a good signal and good signal-to-noise ratio are to be attained. If stations are not in the same general direction, the only recourse is a different antenna for each direction. Fortunately, most weak-signal areas receive all stations from the same general direction, and a good highly directional wide-band antenna is generally found to be satisfactory although at times an additional booster amplifier is necessary. The use of Directronic-type antennas or antenna rotators is especially effective in areas where stations are not in the same direction.

NOISE AND INTERFERENCE

In noisy locations it is important that the antennas have maximum sensitivity toward the stations to be received, maximum rejection of signals arriving in other directions and angles. Inasmuch as nearby sources of noise generally feed signal from beneath the antenna and occasionally from above, a stacked antenna in a noisy location always produces an improvement in signal-to-noise ratio. This is particularly important in areas where the signal is weak and every expedient must be used to advantage to bring that signal out of the masking noise.

As far as horizontal directivity is concerned, it must be maximum in the direction of the station. Parasitic directors and reflectors are not only an improvement with relation to signal strength. They are also an improvement in areas in which the signal is relatively strong. The sharply directional antenna has an improved signal-to-noise ratio, because in addition to making the antenna more sensitive in the given direction it also makes it less sensitive at other angles and, therefore, less sensitive to noise arriving from these angles.

In certain localities there is interference from nearby television receivers (local oscillator radiation and proximity of antenna), harmonics of FM stations, and other sources. In these areas it is also advisable to keep the antenna sharply directional toward the desired stations and also to choose an antenna type and dimension that is least sensitive to the frequency of interference. In some sections, therefore, a wide-band antenna is not desirable because of the sensitivity to stations on other frequencies which enter the receiver and beat with local oscillator fundamental or harmonics. It is expected within the next few years that considerable improvement will be made in the front ends of television receivers themselves to improve spurious frequency rejection and to make them more sensitive on the respective channels to which they are set and less sensitive to other frequencies. At present the interference problem must be overcome by properly orienting and designing the antenna or by means of trap circuits within the receiver or at the point the transmission line is attached to the receiver.

REFLECTIONS

In weak-signal areas the problem is to extract enough signal from the various stations to excite the receiver, while in the strong-signal areas the problem is prevention of interference between stations and multipath reception. It is odd that improvement in both areas involves the same changes in an antenna system. The more sensitive the antenna is in a given direction and the less sensitive it is in other directions, the more effective the antenna system is in delivering a strong signal (of primary importance in weak-signal areas), and likewise the more effective it is in preventing reflections and interference in the strong-signal areas.

Thus, in strong-signal areas it is often necessary to employ a highly directional antenna to prevent reflections from surrounding structures and consequent multiple images. Maximum suppression of reflections is, in most cases, a matter of experimentation, positioning, and orientation of the antenna until the reflected signal is at a minimum and the desired signal is substantially higher in level.

SIGNAL STRENGTH

It is a fact that medium-to-weak signal areas sometimes produce the least problems in antenna choice and orientation. In such an area reflections are generally at a minimum and antennas by and large can be oriented and designed for maximum sensitivity toward all stations. In weak-signal areas the problem is getting the signal out of the noise while in strong-signal areas the problem is the prevention of interference between stations. Thus, antennas in the strong-signal areas, while not always more elaborate than antennas used in the medium-signal areas, are much more difficult to erect and orient to obtain ideal performance on all stations that may be of varying signal strengths and in a number of differing directions. In weak-signal areas the antennas are, of course, more elaborate, and must be designed to attain maximum sensitivity to signal and maximum rejection of noises if satisfactory performance is to be had.

LONG-RANGE RECEPTION

Satisfactory reception of a television signal over distances in excess of 50 miles taxes the ingenuity of the television technician. Not only must he cope with weak signals but with noise and the surrounding terrain. Each new station to be received, although in the same general direction, is a problem in itself, particularly if there is a wide separation in frequency between stations.

The only satisfactory antenna is one with exceptional horizontal directivity, such as obtained with reflector and director assembly. It is very important to mount it at a space loop. In addition, the antenna should be stacked with at least two sections (occasionally as many as four are employed), which improve antenna sensitivity by concentrating it at low vertical angles. At the same time noise pickup is substantially reduced and the signal can be brought out above the adjacent noises. It is very important, if maximum signal is to be delivered, that the antenna matches the transmission line. So far as choice of lines is concerned, if the lead-in must cover a substantial span, a parallel line (or shielded parallel line) is preferable because of its higher impedance and consequent lower distributed capacity and attenuation. In a particularly noisy location it is necessary to take advantage of the shielding characteristics of coaxial lines. Choose a line with the least capacity, least attenuation, plus good quality dielectric. There is no point in having an elaborate, sensitive antenna and lose the signal between the antenna and receiver. Likewise, it is important that we choose the antenna of proper dimension and height to deliver a strong signal and with that choice, make sure that the dielectric of the insulating portion of the antenna and mast be of the very best for high-frequency application. Again, an antenna type which has unusual sensitivity is valueless unless the component parts which make up that antenna are low in loss.

In extended-range reception a wide-band antenna is not always satisfactory because it does not deliver peak performance on any one channel. It is very difficult to design an antenna which would have a uniform and, at the same time, peaked sensitivity for the wide range of frequencies associated with the television spectrum (low set and high set of channels). Thus, for extended-range reception involving stations on both the high- and low-band set of channels it is often advisable to use separate antennas, one ideally designed for the low set of channels, and the second for the high set. It is generally feasible to mount a highly directive three- or four-element parasitic antenna for the high-band set on top of the mast which supports an array used for the low-band set of stations. At any great distance it is better to use separate transmission lines for both antennas instead of a mid-point tap to reduce loading effects by the antenna not being used. To go a step further, if a wide-band antenna has been purchased which is designed for reception on both high- and low-frequency channels, and it is only desired to receive the low-frequency channels, removal of the high section of the wide-band antenna (such as a folded dipole with high-frequency wings) produces an increase in signal strength for the low-band stations.

A booster amplifier is almost essential in most extended-range locations if the signal is to be brought above the masking noise or so-called "snow." The choice of a booster amplifier is an important consideration because its function is to amplify a very weak signai before it is applied to the input of the television receiver. To make full use of the signal the antenna system must be matched by the input of the booster amplifier and, likewise, the output of the booster amplifier should match the receiver input.

If the booster amplifier is to be effective on a weak wide-band signal its inherent tube and thermal noises must be extremely low; otherwise the booster itself will introduce noise which is comparable or stronger than received signal.

Just amplifying signal, therefore, is not the only function of the booster, but also improved signal-to-noise ratio if masking noise is to be cut down. This it cannot do if the booster noise itself is great. Although the stage gain is higher when a pentode r-f amplifier is used for narrow-band work, in wide-band operation it is possible to realize as much gain from a triode because of the relatively lower impedance of the output circuit. A noted advantage of the triode is the lower level of tube noise and consequent improvement in signal-to-noise ratio. Thus, a low-capacity high-g_m triode is often preferable to a pentode in an r-f stage used to amplify an extremely weak wide-band signal.

142. *Space-Loop Positioning and Antenna Orientation*

Correct positioning and antenna orientation are important factors if the very best performance is to be obtained. Not only must the antenna be oriented in the direction of the station, but at this angle the antenna should be moved along an imaginary line toward the station until it is located at a space loop. Thus, before the exact mounting position is finally decided the antenna or a portable test antenna should be moved about the tentative site while receiving signal reports from an observer at the receiver or test set. A weak station, therefore, can be definitely favored by the moving antenna until a strong signal position is found.

The space loops are of course recurrent, depending on signal frequency. Inasmuch as the loop spacing differs with the channel frequency, it is very possible that one position can be found which is at a space loop or near a loop for two or more stations. It is definitely possible with only a little effort in positioning to favor a single weak station. Space-loop positioning is every bit as important as correct antenna orientation in favoring weak stations.

It is advisable to employ a two-man crew to position and orient the antenna system, one man observing the effects on the television screen and a second man shifting the position of the antenna. Self-powered phones can be attached to the transmission line at each end and communication can be established.

In orienting the antenna, rotate the mast until maximum signal strength is obtained as indicated by best contrast.

In many localities the antenna has to be oriented for optimum performance on all stations insofar as signal strength, noise pickup, and reflections are concerned. Seeking a position which will give satisfactory performance on all stations being received reduces to an experimental procedure.

In suburban or long-range reception, all stations are often in the same direction and antenna can be oriented simply with a compass. Be certain to consider magnetic declination in your area. A most troublesome defect which appears on the television screen, particularly in built-up areas, in multiple image or ghosts on the screen when reflections are present. At times as many as two or more images will appear on the screen. Actually, the displaced images are caused by arrival of identical signals at short intervals later. Thus, as the scanning beam travels across the screen a double signal is supplied to the grid of the picture tube, one signal delayed with respect to the other and, therefore, a double image will appear on the screen. When the reflection is from a nearby point the reflection affects the resolution of the picture, although it is not always evident as a double image. The antenna should be located at a point at which reflections are absent or at a minimum.

One common source of reflected signal is a signal bouncing off a reflecting surface (metallic surfaces in particular) in the rear of the antenna (Fig. 284, drawing A). The actual distance over which the reflection is carrying can be calculated by measuring the separation between the images on the screen of the picture tube. At times this procedure in calculating distance is helpful in locating source of a reflection. Another type of reflected signal is one which arrives in an indirect path, such as shown in drawing C. In this example, the reflected signal arrives at an angle from some large, massive reflecting surface, such as a bridge or an extremely large metallic structure. In this case, a difference in arrival time is the difference in time required to travel from the transmitter to the reflecting surface and on to the receiving antenna compared to the time required for the direct wave to travel between the station and the antenna.

For example, on a 10-inch picture tube, the active scanning width is approximately 8 inches, and it requires a scanning beam 53.3 microseconds to travel this distance. Thus, the time required to travel 1 inch is approximately 6⅔ microseconds. A radio wave travels through the air at the rate of 186,000 miles per second and is, therefore, traveling at the rate of 0.186 mile per microsecond. Thus, if the reflected signal requires 6⅔ microseconds longer to reach the antenna it means it has traveled over a path length 1.24 miles longer than the direct path between station and receiver (6⅔ multiplied by 0.186). If the source of reflection, therefore, is directly behind the antenna it would be separated from the antenna one-half of this distance, or 0.62 mile. Insofar as the type of reflection indicated in Fig. 284, drawing B, is concerned, a difference between reflected and direct paths (sum of A, B, and C, minus A-C, the direct path) is 1.24 miles. Thus, if there is a massive structure in the area, it is possible by means of a calibrated map of the city, and with tests and calculations, to ascertain where the reflection is coming from. The easiest figure to remember as far as the 10-inch tube is concerned is that the beam travels 1 inch in 6⅔ microseconds, and a 1-inch displacement between direct and reflected image on the screen represents a path difference of this amount,

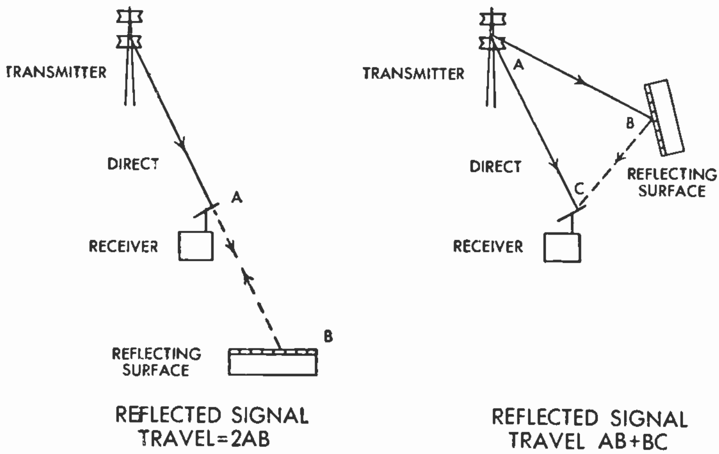


FIG. 284 Reflected Signal Paths

or 1.24 miles; while on a 7-inch tube the beam travels one inch in 9.6 microseconds, and for a 20-inch tube, requires 3.9 microseconds per inch. As is to be expected, reflected signal path displacement of the image and reflection on the screen are also dependent on size of the screen. It is important, therefore, in making any measurements that the width control of the receiver is adjusted to scan the full active width of the scanning surface of the fluorescent screen.

143. *Antenna Installation*

A television antenna installation, in addition to being electrically effective, must be rigid and as unobtrusive as possible. Physical size of the antenna generally is such that it can be a single-mast-supported system, lending itself to ease of mounting on most all types of structures and in various positions.

One of the first jobs to perform in erecting the antenna is positioning of the mounting base or bracket, the style of the bracket depending on the position at which the antenna is to be mounted on the roof or side of the building. Two types of base mounting systems (Fig. 285) show the mast mounted at different points on a building. In drawings A and B properly shaped brackets are used to support the masts to sloping portions of a roof or to peak; the actual mounting base of each mast is then bolted to bracket. A universal

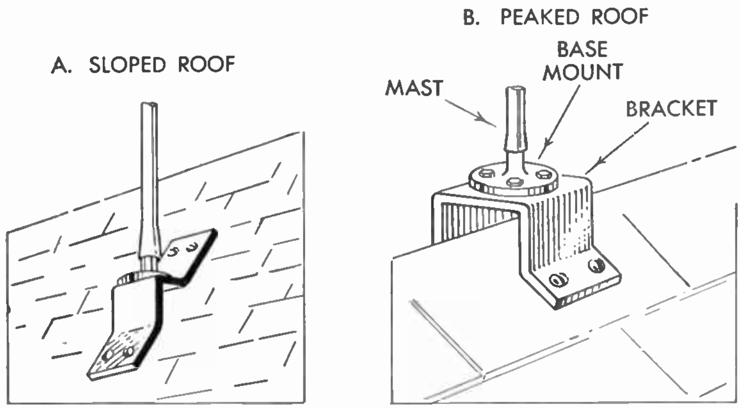


FIG. 285 Roof Mounts

base mount by Brach is shown in Fig. 286, along with a view of it attached to the side of a building. This universal base mount serves as both bracket and base and can be swung into various positions for mounting on the side or top of the building. It can be fastened to roof or side of a house, chimney, or beneath the eaves, and can be tilted and secured in any position to hold the mast upright. On stone or masonry buildings the base mount is first attached to a 2-by-4 block of wood and then secured to the masonry. Another feature of the base is that the antenna itself can be rotated for correct orientation even after the mount has been secured.

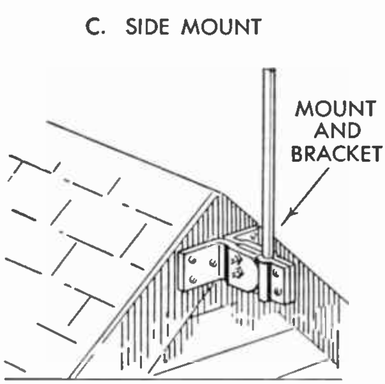


FIG. 286 Side Mount

After the mounting site has been chosen the bracket has to be secured to the structure—the method being dependent on composition of the material which must support the bracket and antenna mast. If the antenna is to be supported on a flat roof the mast base or bracket is secured to a wooden platform, Fig. 287, and the mast is then guyed properly. This method is preferable because drilling in the roof is not necessary. A platform can also be used on the apex of a roof if the proper lengths (according to slope of roof) of wooden dowels are attached to the four ends of the platform.

If no guying is to be used and a metal mast base is to be secured to the roof (either on slope or apex) drill holes with care—the bolts should preferably pass through the roof and be secured tightly on the underside with lock-washers and nuts. Seal lube should be poured into all holes, and bolts or screws should be coated before they are inserted. No drilling or antenna mounting should be attempted on a tile roof (use side-mounting bracket) and proceed cautiously when erecting antenna on a slate roof. When drilling slate use light pressure; at the same time hold the drill firmly to prevent cracking.

If it is necessary to use a side mounting in brick or masonry, special tools are needed to drill the holes for the antenna support. When attaching a support to a brick wall or chimney the holes must be drilled into the brick and not the cement between bricks, to prevent water leaks and loosening of the

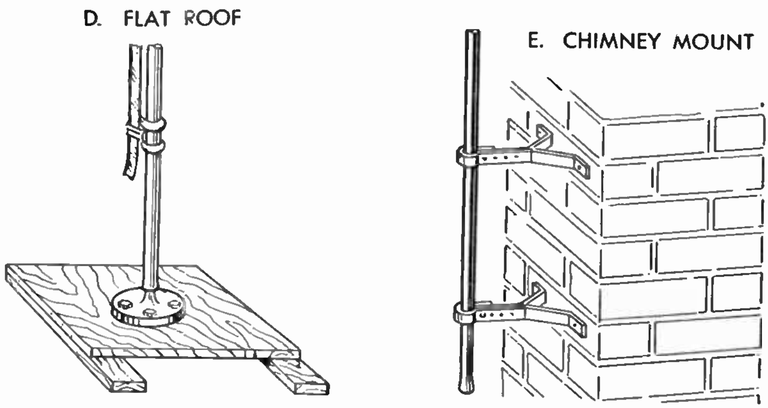


FIG. 287 Supporting Antenna on Roof

bricks. Either a star drill or a slow-speed electric drill with a masonary bit is used. When using an electric drill the hole can be started with an awl tool. Next, a lead expansion fitting is now inserted into the hole and the lead expansion sleeve hammered (use an Ackerman-Johnson tool) until it becomes embedded in the hole. After the holes are drilled and fitted the mast clamps can be bolted to the lead expansion fittings. Smaller holes for mounting insulated screw eyes, etc., can be made with awl tools and then supplied with an awl plug for insertion of the screw eye.

144. *Eliminating and Reducing Reflections*

As the desired signal arrives at the receiving antenna from a given direction, it is often a relatively simple matter to eliminate a reflection by making the antenna more highly directional in the direction toward the station and to reduce its sensitivity at other angles. Thus, a dipole element with parasitic elements, particularly a reflector, will eliminate reflections coming from rear of the antenna and will make the direct signal dominate the reflected signal to the extent that the reflected signal is invisible. Likewise, if the reflected signal is arriving at an angle other than from the rear of the antenna, it is possible, by shifting the antenna position or adding parasitic elements, to reduce reflection by placing the most insensitive angle of the antenna toward the reflected signal path, the antenna remaining reasonably sensitive in the direction of the station.

The most difficult reflections to eliminate are those encountered when the direct signal is severely attenuated by an intervening structure. For example, in Fig. 288, the intervening structure is immediately in front of the antenna and severely attenuates the signal arriving from the station, while the reflected signal is nearly as strong as, and sometimes stronger than, the direct signal. The type of reflection in which the two arriving signals are of approximately the same signal strength is the most difficult to eliminate. The solution, of course,

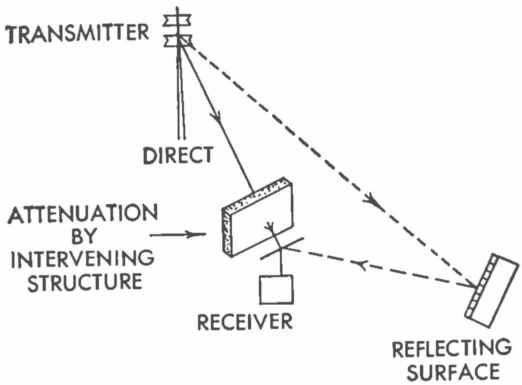


Fig. 288 Attenuation of Direct Signal by Intervening Structure

is to try and make the antenna as highly directional as possible toward the transmitter station. At the same time the angle of arrival of the reflected signal should be ascertained and the antenna system made least sensitive in that direction. In some cases, the easiest solution under these conditions is to direct the antenna toward the reflected signal, hoping that it will have little sensitivity toward the direct path. Of course, the changing of position of the antenna in many cases will assist in finding a spot in which the reflection is at a minimum and where it can be overcome by the direct signal. However, if the reflected signal is coming off a massive structure and the attenuation of direct signal is severe, changing antenna position over a limit set by the building is not always sufficient. Elimination reduces to an experimental method, and best results can be obtained with a two-man crew and observation of the screen as changes are made.

The elimination of reflections in a built-up area where the stations are located at various angles from the receiving antenna is more difficult. If reflection cannot be removed on the various stations by orientation or shifting the position of the antenna a multiple antenna installation must be made. With such an installation the antenna can be cut with peak sensitivity in the direction of the station or stations and, therefore, will have a decided improvement in sensitivity toward the various stations and a decrease in sensitivity at the various angles at which reflections are arriving. Another important factor in the reduction of reflections is properly matching the antenna system and occasionally using a coaxial line because of its shielding properties when the transmission line is long. This expedient eliminates those reflections which enter the system by pickup of the transmission line itself when it is not properly matched or shielded.

A very peculiar reflection encountered in apartment houses or hotels in which an antenna distribution system is used to supply signal to the various receiver outlets is a "leading reflection." This defect is caused by a signal arriving directly at the input to the receiver which is actually leading the signal striking the antenna and arriving at receiver outlet through a lengthy distribution system. Oftentimes, particularly in the higher floors of a building, the signal which arrives directly at the receiver is comparable in strength to the signal which arrives through the distribution system. It is often possible to eliminate this stray pickup by properly shielding the transmission line system and receiver input at its location. Thus, a distribution system to be effective must be properly designed and matched at all points and should employ matched and shielded transmission lines to prevent excitation of the transmission line by path other than through the antenna itself.

In summation, the elimination of reflections reduces to the following procedure:

1. Design the antenna system with maximum sensitivity toward the station and least sensitivity at other angles to reject the arriving reflection.

2. Position the antenna high and clear not only to extract the most signals from the direct path but to get it free of intervening buildings and other obstructions from which the reflected waves bounce. Position the antenna on the building at a point where the desired signal most completely dominates reflection.

3. Design the antenna system to properly match the transmission line, and properly match transmission line to the receiver. This reduces sensitivity of the receiver system to stray pickup through the transmission line or at the receiver input.

4. When a single antenna installation will not eliminate reflections, treat each station and its associated reflection as an individual problem. At times this means resorting to an individual antenna for each station troubled by a reflection.

145. *Interference Problems*

Interference problems can be separated into signal interference and impulse noises. The sources of signal interference are many: adjacent television channels, commercial FM signals, mobile, amateur, special services, local oscillators of nearby receivers, and interference in the i-f frequency range. Radiation from diathermy equipment can also be considered signal interference, although the signal radiated serves no useful purpose whatsoever and should be completely suppressed. The most common impulse noises which affect television receiver performance are: auto ignition systems, sparking machinery and motors, elevators, high-voltage leakage associated with neon and germicidal lamp installations, and various home electrical appliances. There are two basic methods of combating interference, by filtering the source of the interference and by making the receiver insensitive to the interfering signal.

ADJACENT CHANNEL INTERFERENCE

The television receiver is subject to adjacent channel interference because of its extremely wide bandwidth and poor selectivity, which means that the sideband tapers off slowly over a substantial span of frequency spectrum. Inasmuch as its selectivity at the portion of the spectrum to which it is tuned is low in comparison with what can be obtained with a peaked response, it is also subject to interference from signal frequencies far removed. So far as adjacent television channels are concerned it is necessary to employ resonant traps in the i-f system to remove interference from this source. Other interfering services are radio amateurs near channel *2*, the FM stations between low and high television bands, plus the various government and mobile services between both high and low bands and below and above the television spectrum. If a transmitter in this spectrum is very near a TV receiver often the only manner in which interference can be held down is by means of trap circuits inserted into the television receiver.

IMAGE RESPONSE

The sensitivity of the television receiver to image frequencies is high, but fortunately the use of the higher i-f frequencies has removed some of the potential image-response difficulties. For example, with the 25-megacycle and higher i-f spectrum, as shown on the chart, the local oscillator frequencies in the television receiver are such that image frequencies are outside of the television channels and outside a greater portion of the FM band.

Image frequency is calculated simply by adding the i-f frequency to the frequency at which the local oscillator is set on a specific channel. For example, if we are tuned to channel *2* and the picture i-f frequency is 25¾ megacycles, the image frequency will be the oscillator frequency, 81 megacycles on channel *2*, plus the picture i-f frequency of 25¾ megacycles, to give an image frequency of 106.75 megacycles. Thus signal from any station which transmits on 106.75 megacycles beats with the local oscillator of the receiver set on 81 megacycles to produce an i-f of 25¾. Thus an FM station located near this frequency (FM band extends to 108 megacycles) produces interference on the pattern, depending on its signal strength plus proximity and sensitivity of the receiver to the image frequency. Of course, all the television channels have their own image frequencies, and if the receiver is near any station using that image frequency there will be interference on the received image which can be eliminated with a trap system at the receiver or highly directional antenna.

TRANSMITTER PROBLEMS

Another source of interference is the harmonic output (second, third, and higher) of various transmitters. For example, if channels *6* and *7* are assigned in a locality the second harmonic of the channel-*6* transmitter would fall into channel *7*'s spectrum. The only solution to harmonic interference is to suppress the harmonic at the transmitter. This can be successfully accomplished except in a limited area near the transmitter. Here even low-percentage second-harmonic radiation enters the receiver and interferes with the desired signal because harmonic and desired signal both arrive in the same frequency spectrum and a trap system cannot be employed to tune out the harmonics. So far as the high set of television channels is concerned, second harmonics (of commercial FM stations) and some harmonics (third) of the old low-band transmitters fall into this spectrum. Likewise, harmonics of the amateur band and other services fall within the frequency spectrum of the low-frequency set of television channels. Here again, it is a matter of suppressing harmonic radiation from the transmitter, and in the case of a nearby transmitter it is often impossible to suppress it sufficiently to eliminate all interference.

LOCAL OSCILLATOR RADIATION

In a wide-band amplifier, the sensitivity of the system to the frequency spectrum to be received is relatively low. At the same time it is more sensitive to frequencies removed from the desired spectrum compared to a more selective system. Thus, it is possible to have signals enter the receiver far removed from the channel to which the television receiver is tuned, and it is also possible to have the local oscillator signal in part pass through the mixer or r-f stage on to the antenna, where it is radiated. If any other television receivers are in the immediate vicinity they will pick up this local oscillator signal and produce interference on the channel to which the second receiver is tuned. Local oscillator radiation is a difficult problem to overcome where many television receivers are grouped, as in a crowded section of the city or apartment building. Not only is the fundamental frequency of the local oscillator a source of interference, producing loss of resolution on some other receiver, but its harmonics can also develop a disturbing pattern on a television receiver set on some other channel. For example, if the local oscillator on channel *3* is set on 87 megacycles any other receiver in the vicinity tuned to channel *6* will pick up this radiated signal. Likewise, its second harmonic produces a signal at 174 megacycles, or on channel 7. If a higher frequency i-f is used, such as 26.4, the second harmonic of the local oscillator frequency is almost on the picture carrier frequency of channel 7. This particular problem is best licked at the source by suppression of local oscillator radiation both by proper receiver design and proper transmission line system between antenna and receiver.

HARMONIC MIXING OF LOCAL OSCILLATOR AND SIGNAL

Inasmuch as the front end of the television receiver is not yet an ideal system, the receiver has a certain amount of sensitivity to frequencies far removed from the channels to which it is tuned. In fact, it is possible for a signal to enter the receiver and beat with the second harmonic of the local oscillator to produce the proper i-f frequency to interfere with a signal on the screen. For example, the channel *10* picture-carrier frequency is 193¼ megacycles. If our television receiver is tuned to channel 6, the local oscillator is on 109 megacycles and has a second-harmonic component on 218 megacycles. This second-harmonic component of 218 megacycles beats against the picture-carrier frequency of channel *10,* which enters the receiver when it is set on channel *6.* The second harmonic produces an i-f frequency of exactly 24¾ megacycles and superimposes the picture of channel *10* on that of channel *6.* The problem, again, is not so much the second harmonic of the local oscillator but the fact that the receiver itself is sensitive to a frequency so far removed from the desired frequency. This type of interference can be suppressed by means of wave traps in the front end of the television receiver. This same type of interference can be caused by double conversion. For example, 109 megacycle local oscillations reaching r-f amplifier can beat with incoming channel *10* signal to produce a difference of approximately 84 megacycles creating interference in channel *6.*

Channel	LOCAL OSCILLATOR FREQUENCIES *I-F 25.75*	*I-F 26.4*	*I-F 26.6*	*Image Frequency 25.75 MC. PIX I-F*	*2nd Harmonic of Local Oscillator 25.75 MC. I-F*	*Transmitter 2nd Harmonics*
3 60–66	87	87.65	87.85	112.75	174	122.5 131.5
6 82–88	109	109.65	109.85	134.75	218	166.5 175.5
10 192–198	219	219.65	219.85	244.75	438	386.5 495.5

FIG. 289 Signal Frequencies for Typical Allocation

These spurious frequencies produced by mixing action with the second harmonic of the receiver local oscillator not only cause interference between television channels but between television channel and the FM spectrum again. For example, the harmonic of the local oscillator, when the receiver is on channel *2,* will beat with the stations at the high end of the commercial FM channel to produce an i-f frequency component in the picture i-f system of the receiver.

DIATHERMY

Diathermy radiation produces a broad-band signal with hum modulation, producing a herringbone pattern on the screen, as shown in photo 3 of Fig. 283. Diathermy is of such a broad-band and indefinite frequency spectrum that a trap system is not particularly useful in the receiver. Instead, suppression of radiation should occur on the diathermy equipment since radiation serves no useful purpose in diathermy treatment.

146. *Suppression of Signal Interference*

The many possibilities for signal interference are demonstrated in the frequency spectrum chart of Fig. 290. With the receiver set on channel 3 it is possible to obtain interference from any of the frequencies so indicated. It is helpful to draw up such a chart for each station in your locality to assist in the diagnosing of signal-interference possibilities.

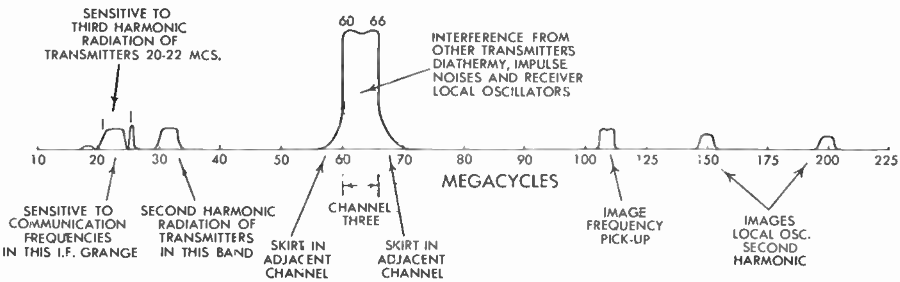


FIG. 290 Sources of Interference with Receiver Set on Channel *3*—25 Megacycle-I-F

Interference by stations on the same frequency as the channel to which the receiver is tuned, such as would be obtained from the fundamental of an interfering signal, second harmonic of an interfering station, or local oscillator radiation from a nearby television receiver, is often difficult to suppress.

The antenna system that is being interfered with, of course, should be directed away from the source of interference, if at all possible, and made more sensitive in the direction of the desired station. However, in the case of local oscillator interference, more than likely both antennas are directed at the same angle. In this case, booster amplifiers between antenna systems and receivers add another stage of rejection to each receiver, reducing the amount of local oscillator radiation with the addition of one or more tuned circuits. The ultimate solution to image interference and mixing with the second harmonics of the local oscillator is to design the front end of the television receiver with a much improved signal sensitivity and improved image rejection. Again, a high-gain, highly directional antenna is preferable for rejection of image and spurious signal interference because it not only has peak sensitivity for the desired frequencies but it can be directed in a manner which will reduce the sensitivity of the antenna system to interfering signals which arrive at other angles.

INSERTION OF WAVETRAPS

Wavetraps can be inserted into the transmission line at the receiver termination or in some cases at the mixer grid circuit to suppress undesired signals (Fig. 291). They are often series resonant and inserted in one or both conductors of the transmission line and tuned to any undesired frequency. Resonant circuits can be in the form of suitable lumped constants (drawing A) or

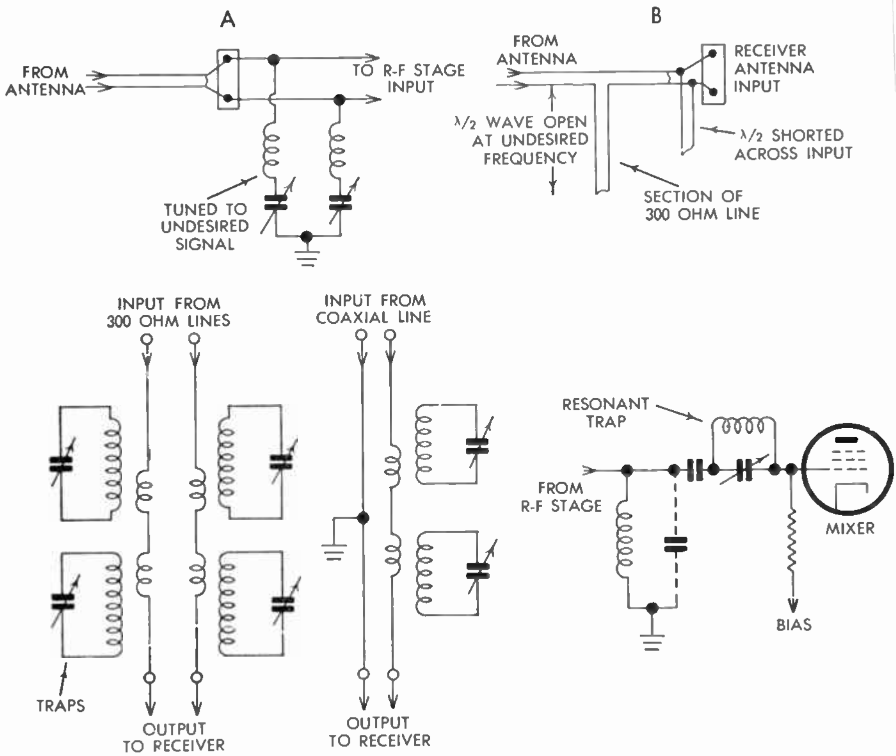


FIG. 291 Input Traps to Remove Spurious Signals

sections of actual transmission lines (drawing B). In the case of the interference previously mentioned between channels *6* and *10* (picture carrier of channel *10* beats with second harmonic of local oscillator on channel *6* to produce a signal in the i-f spectrum of the receiver), it can be eliminated by a half-wave section of line open, tuned to the picture carrier frequency of channel *10*. It forms a parallel resonant circuit in series with the signal path to the r-f amplifier of the receiver. It prevents the entrance of channel *10* picture carrier into the receiver. This system would also attenuate channel *10* signal when the receiver is set on channel *10*. However, at that time the end of the half-wave section transmission line can be shorted together to form a series-resonant circuit tuned to the picture carrier frequency of channel *10* and, therefore, the passage of the signal into the r-f stage is not impeded. A simple shorting switch on the end of the transmission line would do the job. Keep in mind that the actual physical half wave of the transmission line section being used as a tuned circuit is shorter than a half wave in space by the velocity constant of the line. A quarter-wave section of line open or half-wave shorted can be attached directly across antenna input to do the same job—shorting input at undesired frequency.

It is possible to employ groups of resonant circuits in such a suppression system to eliminate a number of spurious signals which interfere with operation of the receiver (drawing C). Suppression circuits can also take the form of a parallel resonant circuit in series with a single path between the r-f amplifier and the mixer grid; the parallel-resonant circuit (drawing D) prevents the transfer of the undesired signal to the grid of the mixer and, therefore, curbs the mixing action which produces the interference in the i-f system.

Concerning interference from international short-wave stations and other transmitters in the frequency spectrum between 20 and 25 megacycles, many receivers use an inductance which is center-tapped and grounded at the input to the receiver. Such a reactance is very small at the lower i-f frequency ranges, therefore shunting off to ground any of these undesired short-wave signals. At the higher frequencies of the television band the inductive reactance would be very large and, therefore, its effect insignificant. Again it is a matter of designing the front end of the receiver with a peak sensitivity at the desired frequency and good rejection at undesired frequencies as well as proper shielding of the receiver to prevent direct pickup into the i-f system.

In summation, the following steps can be taken to reduce interference from spurious signals:

1. Orient and design antenna to have peak sensitivity toward the desired signals. This means a wide-band antenna is not always feasible because of its sensitivity to signals over a wide frequency spectrum. In the elimination of interference from FM stations and other services and their harmonics, it is better to use an antenna which is only sensitive over the spectrum of the frequencies desired. Again, if interference is severe it is a matter of treating each station desired as a separate problem and perhaps using a separate antenna on those stations which are the most interfered with.

2. If the difficulty is a result of second-harmonic radiation, try to cooperate with the station if at all possible in the suppression of their second harmonics. If the second harmonic is particularly strong and station cooperation is not forthcoming, report to the FCC. All cases of diathermy interference should be reported immediately if this type of interference is to be eliminated.

3. Interchannel interference can be eliminated with wave traps. First, calculate the frequency of the interfering wave and construct or purchase a wave trap which can be inserted and tuned until the interference disappears.

Booster amplifiers, because of their added selectivity, are also very effective in the removal of spurious signal interference. In strong-signal areas you will occasionally find a booster amplifier in operation for this reason.

147. *Suppression of Impulse Noises*

The reduction of interference from impulse noises, which cause streaks of black and light to run across the screen in a random pattern, and in the case of severe noise cause the receiver to go out of synchronism, is obtained in a number of ways. So far as the receiver itself is concerned effective rejection of impulse noises is attained with a properly designed and matched antenna system as well as a shielded lead-in which is routed away from sources of noise. Inasmuch as many noises can enter the receiver through the power line, a filter inserted between outlet and receiver does a lot to cut down noises from home appliances or equipment operating on the same power lines. The antenna should be positioned high and clear and well away from any source of noise. In addition to that, if located in a noisy district a coaxial feed-in should be used to shield the transmission line from noise. This coaxial line can be in the form of a simple line with the proper matching system at the termination to match the input to the receiver. Or a dual coaxial line can be used (Fig. 292). If the coaxial line covers a considerable span its shield should be grounded at various points to reduce any tendency for line pickup. It is, of course, extremely important that a good low-loss line be used because with a poor quality high-attenuation line, reduction of noise pickup is confronted by additional attenuation of the signal and nothing is gained. If 300-ohm line is used noise pickup on the line can be reduced considerably by twisting the line every foot or two.

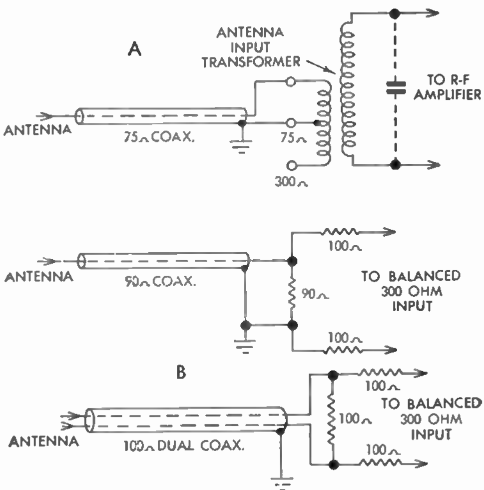


FIG. 292 Matching Coaxial Lines to Receiver Input

Impulse noises can also be attacked at the source by means of proper suppression systems like those used on neon lights and other high-voltage installations. After locating a source of noise try to cooperate with the owner of the radiating equipment to suppress noises with a filter or specially designed commercial suppressor. The manufacturers of such equipment should be notified so that they may include as a part of the particular device a means of suppressing radiation. So far as home appliances are concerned, as well as electric cash registers, office equipment, electric shavers, etc., a considerable reduction in the noise component which reaches the receiver can be obtained by simply inserting a small 0.03-microfarad capacitor across the line at the input to the device. For complete suppression of noises from heavier industrial equipment there are several types of commercially available filters consisting of series-inductor and shunt-capacitor combinations which prevent r-f energy from feeding back to the power line and radiating or feeding directly into radio equipment.

In summation, the procedure for eliminating impulse-noise interference at the receiver is as follows:

1. Erect the antenna high and clear, of course, and as far away as possible from sources of noise such as a busy street, elevator shaft, etc. Again, design the antenna to have low-angle directivity and, therefore, have least sensitivity to noise arriving at other angles. Design the antenna to extract the most signal possible and shield the transmission line which feeds the antenna to the receiver, thereby improving the signal-to-noise ratio.

2. Insert high-frequency filters in the power mains between the outlet and the receiver proper to prevent direct excitation of the receiver by noises.

3. Try to locate filters at the point where the a-c line feeds the device which is causing the noise. So far as auto ignition is concerned, capacitor spark suppression can eliminate this source of interference.

The television receiver of today, because of its wide acceptance, makes it susceptible to interference and noise. As the number of stations increases in all high-frequency services and as the number of receivers increases, an understanding of noise-prevention methods and interference will earn dividends for the television technician.

QUESTIONS

1. Explain in detail relation between frequency response and lines of resolution.
2. Differentiate between vertical and horizontal resolution in respect to: a) limitations and b) methods of checking resolution on a test chart.
3. What apparent defect does poor low-frequency response produce in reproduction of test chart?
4. What apparent defect does poor high-frequency response produce?
5. List performance checks which can be made with RETMA test chart.
6. Give a step-by-step procedure in attempting to minimize reflections at a receiving site.
7. Detail all considerations which should be considered in installation of an antenna in a weak signal area.
8. Detail a logical installation procedure to be followed in the interest of customer satisfaction.
9. Discuss importance of antenna positioning and orientation.
10. List many sources of signal and impulse interference.

Chapter 13

ALIGNMENT AND TROUBLE SHOOTING

148. *Alignment Objectives*

Alignment and trouble shooting of a television receiver are greatly facilitated with proper and accurate test equipment. A vacuum-tube voltmeter or high-resistance voltmeter is necessary to make various electrode and circuit measurements, particularly those in conjunction with high-impedance circuits in the sweep and sync systems of the television receiver. An a-c vacuum-tube voltmeter, one which can be used to measure peak amplitudes of various non-sinusoidal waveforms, assists in location of more obscure troubles. A versatile oscilloscope is a timesaver in trouble shooting the video amplifiers and sync and sweep systems. Even an inferior oscilloscope in the hands of a capable technician can be used to isolate most of the defects which arise in the television receiver.

A number of test instruments are necessary to perform a complete alignment of a television receiver. A sweep oscillator is used to align or make the final alignment adjustments on most all television receivers. An oscilloscope is used to obtain a visual presentation of the response curves. A simple and inexpensive oscilloscope can be used to obtain these response curves because of the extremely low rate at which they are presented on the screen (60 or 120 cycles). A reliable and accurately calibrated signal generator is a very important instrument in adjusting the critically tuned traps and oscillator of the television receiver. A complete alignment of a television receiver is a tedious process and, fortunately, most of the systems are wide band and stable, requiring only infrequent alignment. Those circuits which do require occasional alignment can be adjusted simply. The sections of the television receiver which occasionally require alignment or an alignment check are: the picture and sound i-f systems, the r-f section, and some special circuit such as an a-g-c or automatic sync system.

Insofar as trouble shooting is concerned, a simple signal-tracing method can first be employed in localizing defective i-f or r-f stages with a signal generator. An oscilloscope is used to isolated trouble in the video amplifier or sync and sweep systems; it locates the more obscure troubles in the sync and sweep systems by comparing the fidelity of the wave with manufacturer's recommended standard waveforms. The most convenient device for trouble shooting the television receiver is the picture tube itself because every defect will evidence itself as a deformed or irregular presentation on the screen of the picture tube. As for speed of trouble shooting, the capable technician who understands thoroughly the theory of operation of the various component circuits of the receiver will find his knowledge enhancing his ability to locate troubles efficiently and quickly.

In summation, the following equipment is necessary for alignment and effective trouble-shooting of the receiver:

SWEEP OSCILLATOR

A sweep oscillator assists in alignment of the picture r-f and i-f sections of the receiver. It can also be used for alignment of the narrower band sound i-f system, particularly alignment of the discriminator.

VACUUM-TUBE VOLTMETER

A vacuum-tube voltmeter can be used to check the d-c voltage at various points in the circuit; because of its high impedance, it is particularly useful in checking the voltages of high-impedance plate and grid circuits. It also assists in alignment of various circuits in the receiver; namely, discriminator, wave traps, a-f-c, and so forth. A high-frequency probe attached to the voltmeter can be used to advantage in checking the a-c voltages at various points in the receiver. If the meter is designed to read peak voltage of non-sinusoidal wave forms, it can be used to measure the amplitudes of the various sync and sweep waveforms.

OSCILLOSCOPE

An oscilloscope is necessary when alignment is performed with a sweep oscillator. It also can be used with a modulated signal generator to align wave traps and other narrow-band circuits. An oscilloscope can also be used in the sync and sweep systems of the television receiver to observe the fidelity of the various waveforms or to ascertain their presence at various points. A good oscilloscope with a wide-band vertical amplifier can be used to check the absolute fidelity of various waveforms and thus assist in locating obscure trouble. An inexpensive oscilloscope can be used to obtain approximate waveforms; in the hands of a good technician, it can be used to isolate most defects.

SIGNAL GENERATOR

An accurately calibrated signal generator is necessary in the alignment of critically tuned traps and other sharply tuned circuits. In addition, if the sweep oscillator does not incorporate an internal marker system, an accurate signal generator must be used to calibrate the response curves on the oscilloscope screen. A r-f signal generator can also be used to signal-trace the r-f and i-f systems to locate a dead stage.

149. *Test Oscilloscope*

The characteristics of a test oscilloscope itself must be considered when using a particular model for alignment and trouble shooting of a television receiver. Although a relatively inexpensive oscilloscope can be used for a great number of checks and procedures, a more advanced design oscilloscope permits more precise observation of operating characteristics. Some of the characteristics to be considered in using an oscilloscope for television applications are discussed below.

OSCILLOSCOPE LOADING

Oscilloscope loading must be considered in use of an oscilloscope for checking waveforms across a high-impedance circuit. For example, if an oscilloscope is said to have an input impedance of 1 megohm and 10 micro-microfarads, it means that the presence of oscilloscope across any circuit adds an impedance consisting of a 1-megohm resistor shunted by a 10-micro-

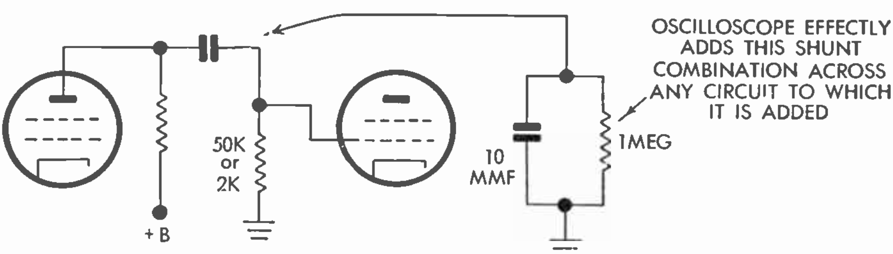


FIG. 293 Loading Effect of Scope

microfarad capacitor (Fig. 293). Inasmuch as reactance of capacitive element varies with frequency, the oscilloscope will have more of a loading effect at high frequencies than it does at lower ones. Actual amount of loading is also dependent upon the impedance of the source across which observed waveform is present. For example, as demonstrated in Fig. 293, the oscilloscope across a 50,000-ohm source will reduce the impedance of this source 30 per cent at the frequency at which the reactance of the oscilloscope capacity equals the ohmic value of the resistor. This frequency is approximately 320 kilocycles. Thus if a 320-kilocycle signal is present across the 50,000-ohm source, the presence of the oscilloscope reduces the amplitude of this frequency 30 per cent. If this signal is a part of a non-sinusoidal waveform, the loss in amplitude affects the fidelity of the composite waveform; consequently, a true indication is not shown on the oscilloscope screen. Higher frequencies are more severely attenuated. Likewise, if oscilloscope is connected across a higher impedance circuit, its effect will be felt at even lower frequencies. In summation, although an oscilloscope has a high impedance so far as its resistive component is concerned, the high-frequency loading effect is caused by the low capacitive reactance that the oscilloscope shunts across the circuit under test. It is important, therefore, when observing signals or waveforms which contain high-frequency components, that they be observed across as low an impedance point as possible.

If the same oscilloscope is attached across a 2,000-ohm source, the 30 per cent loading point is not felt until the frequency is 8 megacycles or higher. It is apparent therefore that although an oscilloscope can be attached across high-frequency or high-impedance circuits when observing television waveforms, which contain high-frequency components, to preserve absolute fidelity of waveform, it is necessary that these waveforms be observed across as low an impedance point as possible. If a certain television waveform contains high-frequency components and if the point at which it is observed is of relatively high impedance, the presence of the oscilloscope will distort the waveform and possibly affect operation of the circuit.

If the oscilloscope is attached across a very high impedance circuit, not only the reactive component is important but also the resistive one. For example, if the oscilloscope is attached across a 1-megohm resistive source, the resistive component of the oscilloscope shunted across same source reduces its effective resistance to 500,000 ohms. This reduction exists at all frequencies. For example, if we are measuring the frequency of a vertical blocking oscillator, operating at 60 cycles, the grid circuit has resistance in excess of 1 megohm, and if we attach the oscilloscope to this point, the resistive component of the grid circuit will be substantially lowered; therefore, the frequency of the oscillator itself will change and along with it the shape of the grid waveforms. It is important, therefore, that the television technician know the characteristics and limitations of the oscilloscope he uses and that he understand the loading effect of the resistive and capacitive components presented by this oscilloscope. It is a wise policy to locate the lowest impedance point from which a particular waveform can be observed to prevent distortion of the waveform or changing of the operating characteristics of the circuit under observation.

The more expensive oscilloscopes incorporate a high-frequency probe and generally contain a large value resistor shunted by a very small capacitor which effectively isolates and reduces the loading of the oscilloscope. At least it makes the loading of the oscilloscope a known quantity and not dependent on the proximity and the manner in which the lead connects to the oscilloscope. When a scope without a high-frequency probe is used, the loading it presents to the circuit is not a fixed quantity but is dependent on the capacity of the test leads used—their length and their proximity to other component parts and ground. The high-frequency probe, with its isolating components at end of the probe which attaches to circuit to be observed, keeps loading constant. In addition, stray pickup is reduced because probe lead is shielded.

OSCILLOSCOPE SENSITIVITY

Another important factor when choosing an oscilloscope for a specific purpose is sensitivity—the ability of the oscilloscope to cope with the amplitude levels of the signals to be observed. If an oscilloscope has an exceptional sensitivity, it is also important that an attenuator or switching arrangement be present to permit observation of higher voltage signals. If an oscilloscope is to be used for alignment in particular, it must have a very good sensitivity to be able to observe the very weak output of a video detector when a sweep oscillator is being used as a signal source. The output of the average sweep oscillator is sufficiently high to permit observation on the conventional oscilloscope only after signal has passed through two or more stages. However, in some cases it is necessary to check the response of a stage or two which has relatively low amplification; therefore, if the response pattern is to be obtained, the oscilloscope must be capable of amplifying a very weak signal.

Oscilloscope sensitivity is generally measured in rms volt-per-inch vertical deflection on the screen. For example, if a specific oscilloscope has a sensitivity of ½-volt rms per inch when the gain control for the vertical amplifier is set on maximum, it means that upon application of a signal of ½ volt, a 1-inch high deflection is observed on the screen. Sensitivity of commercial oscilloscopes for television and special servicing use vary from approximately a few millivolts rms per inch to a few volts rms per inch.

In television, it is also necessary to observe complex waveforms which may be in the order of a few hundred volts peak; therefore, it is necessary that the oscilloscope also be capable of handling a relatively high amplitude voltage by insertion of an appropriate attenuator.

FREQUENCY RESPONSE

The frequency response of the vertical amplifier of the oscilloscope must be broad if the various complex waveforms associated with a television signal are to be observed with fidelity, because the individual waveforms not only contain their fundamentals but harmonic components up into the megacycles. If the waveform is to be reproduced on the oscilloscope screen as it occurs, the vertical amplifier of the oscilloscope must be capable of transferring all its frequency components.

So far as alignment alone is concerned, the oscilloscope need not have good high-frequency response because the repeating waveforms, which represent the response characteristic of a specific amplifier, occur at a very low frequency. An oscilloscope with a rather limited response can be used to trace most of the defects inherent in the complex waveform circuits of television receivers, particularly if the technician makes certain that he applies his oscilloscope to low-impedance circuits which are not severely loaded by the oscilloscope. His knowledge of what constitutes the high- and low-frequency components of a complex waveform will assist him in finding those deformities which are likely to be present on observed waveform because the oscilloscope has a limited frequency response. For example, when an extremely sharp leading edge is applied to an oscilloscope with poor high-frequency response, it is to be expected that the leading edge will appear rounded off on the oscilloscope screen. Nevertheless, a rather limited oscilloscope with poor response can be used to obtain an approximate waveform at various circuits in the receiver and to still give a satisfactory indication of correct or incorrect operation.

Commercial oscilloscopes, dependent on price and design, have a low-frequency limit, which is set at some value between 10 and 100 cycles, and a high-frequency limit, which is set at some value between tens of thousands of cycles and one to two megacycles. Although the oscilloscopes are approximately linear over the frequency ranges mentioned, many oscilloscopes are capable of reproducing a signal at much higher frequencies, although with considerable attenuation, and it is not unusual to find an oscilloscope that will amplify, to a certain extent, frequencies as low as 5 cycles and as high as 5 megacycles.

Although an oscilloscope is said to have a specific frequency response, it does not necessarily mean that you can apply that particular frequency from the source and expect to get it on the screen of the oscilloscope tube unless specific impedance requirements are met. It is true that if a specific frequency is delivered to the amplifier, it will be reproduced on the screen. However, if that signal is attenuated before it reaches the amplifier, frequency response of amplifier is of no benefit. Thus, if the signal to be observed has high-frequency components, try to observe that waveform across some low-impedance circuit where the input impedance of the oscilloscope is substantially larger. Under these conditions, the entire signal is delivered to the vertical input of the oscilloscope, and the frequency response of the amplifier will be instrumental in producing a true likeness of the waveform on the screen.

If the technician owns an oscilloscope with a limited frequency response not good enough for reproducing the television waveforms, it is possible to improve response at a sacrifice in gain by shunting the vertical amplifier plate load resistors with smaller value resistors. Fidelity of the waveform reproduction is thereby improved though restricted to a small fraction of the screen height. When attempting to observe the relatively higher amplitude waveforms encountered in the sweep circuits, it is also possible to use an attenuating cable and probe which reduce the input capacity and loading to the scope although again sacrificing gain. Such a homemade cable could consist of a series resistor of approximately 1 megohm shunted by a 5-micromicrofarad capacitor mounted in the probe, with the other end of the cable connecting into the regular scope input.

HORIZONTAL SWEEP RANGE

Sweep range and fidelity of horizontal sawtooth are important if the waveform to be observed is to be reproduced with fidelity. For example, many complex waveforms are checked with reference to time which should be laid out linearly by the horizontal sweep line. To do so, the horizontal sawtooth must be linear to keep the beam moving from left to right across the scope screen at a constant velocity. Also, if the waveform to be observed is a repeating signal, it is necessary that the horizontal sweep be adjustable to its frequency or some subharmonic of it. For television application alone this does not place any strict requirements on the horizontal sweep system of the test oscilloscope because the field rate is only 60 cycles per second and the line rate, 15,750. Thus, if the oscilloscope has a linear sawtooth from 30 cycles to approximately 15,000 or even 8,000 cycles, it will be quite satisfactory because for true observation of waveform it is always best to have 2 or 3 complete cycles showing on the screen to counteract any loss of waveform by the retrace of the oscilloscope.

In television practice, for observation of a frame or two fields of a television signal, the oscilloscope is set on 30 cycles, producing two fields of the 60-cycle repeating vertical rate on the screen. For observation of two lines of the television signal, a horizontal sweep is set on one-half of 15,750-line rate or 7,785 cycles. Thus, in the time required for the beam to sweep left to right across the screen and back, two lines of picture occur and can be observed on the screen.

150. *Typical Commercial Oscilloscope*

An excellent oscilloscope for routine television servicing is the Triplett 3441, Fig. 294, which can be used as an alignment scope and as a signal-tracing instrument in the sync and sweep systems of the television receiver. It contains push-pull vertical and horizontal amplifiers to obtain balanced deflection, high resolution, and high sensitivity. A cathode-coupled multivibrator generates the horizontal sawtooth that is applied to a two-stage resistance-coupled amplifier and then to the push-pull output stage that presents a balanced feed to the horizontal deflection plates. The vertical amplifier consists of a high-impedance cathode-follower input stage, a pentode amplifier, a triode phase-inverter, and balanced vertical output stage that supplies the signal to be observed to the vertical deflection plate of the scope. A *Z*-axis input provides a means of applying signal to the cathode of the scope tube and therefore permits intensity-modulation of the scanning beam with specific applied signals.

The vertical amplifier input has an impedance of 2 megohms shunted by 22 micromicrofarads of capacity with a 10-millivolt-per-inch deflection sensitivity at a 2-megacycle bandwidth and a 20-millivolt-per-inch deflection sensitivity when the vertical amplifier is set on the 4-megacycle bandwidth position. Bandwidth of the amplifier is controlled by a switch at the rear of the scope and is normally set on the 2-megacycle position for routine use and on the 4-megacycle position for more precise observation of waveform. The bandwidth control is in the plate circuit of the vertical output stage (Fig. 295) and adds resistance to the plate load on the 2-megacycle position in order to increase voltage output to the deflection plates. Of course, the added plate resistance reduces the bandwidth of the amplifier, as compared to the 4-megacycle setting. The vertical gain control is located in the cathode output circuit of the input stage, tube V3, and supplies signal to the pentode amplifier, tube V5, which employs series-shunt peaking in order to obtain a satisfactory frequency response. This video amplifier supplies signal to a triode stage having cathode and plate output terminals, thus permitting a choice of signal polarity (plate and cathode output are of opposite polarity) for excitation of the two grids of the balanced output stage with switch *S7* choosing proper polarity.

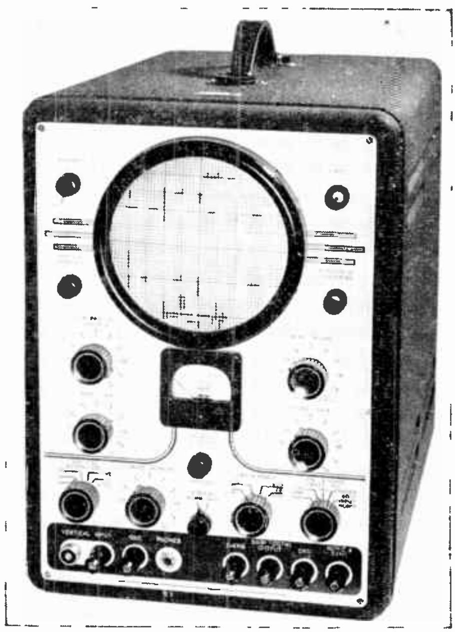


FIG. 294 Triplett Oscilloscope

The vertical amplifier is conveniently calibrated with a built-in sine-wave source and an a-c meter; it can be used to measure sine-wave input as well as peak amplitude of non-sinusoidal waveforms. Calibration is accomplished by applying a known sine-wave voltage to match the amplitude of any signal to be measured. This voltage is then measured on the voltmeter, calibrated to measure peak voltage. The voltage-measuring circuit is very useful in checking the peak amplitude of the various television waveforms in the television receiver. In operation, a signal to be measured is applied to the vertical input; the attenuator and gain control are then set to obtain a suitable vertical deflection on the screen. The scope is next switched to calibrate, and the calibration control *R13* is adjusted until the calibrating sine wave occupies the same vertical spacing on the screen. Peak voltage of the applied signal is indicated on the voltmeter scale.

The sawtooth generator has a frequency range from 10 cycles to 60 kilocycles, and there is a 60-cycle sine-wave source for alignment use. Some of the other features of the scope are as follows—retrace blanking, sawtooth-

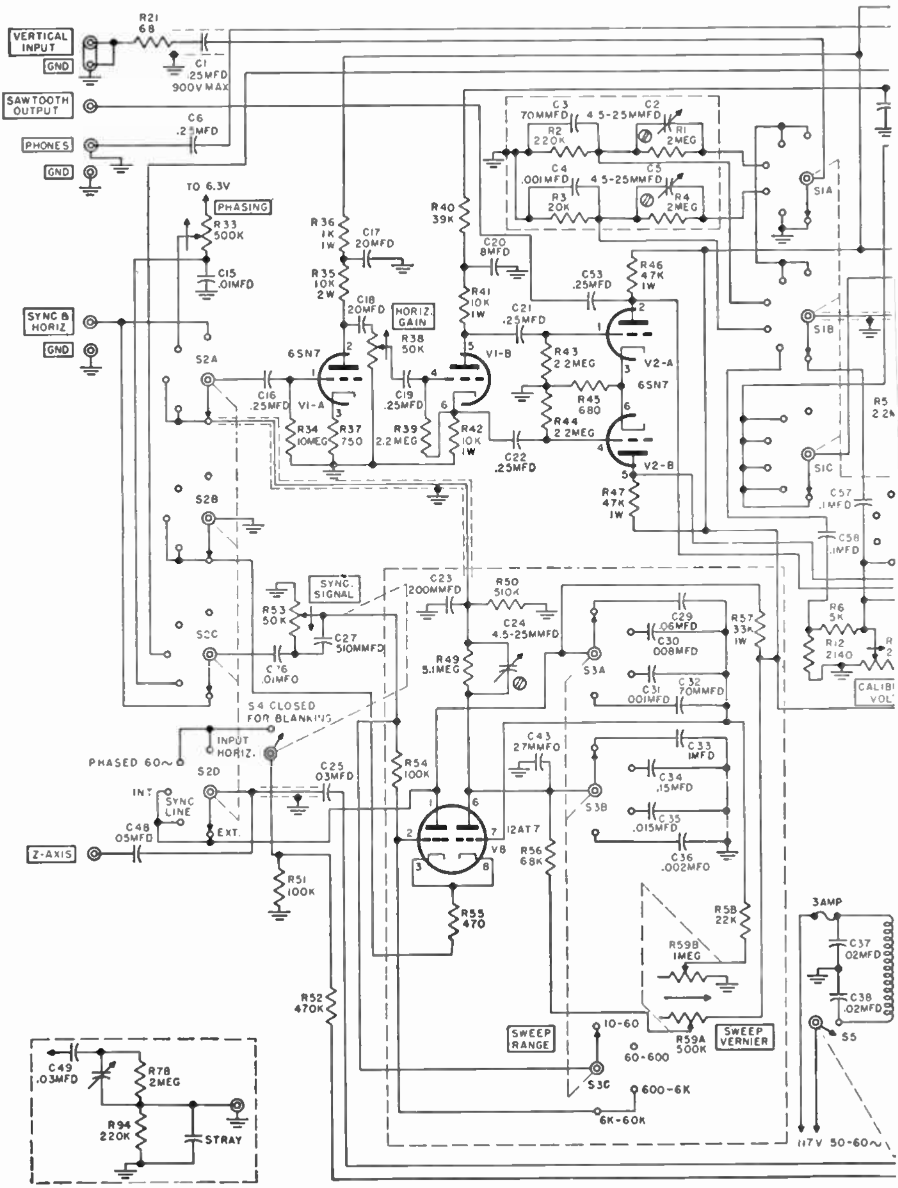
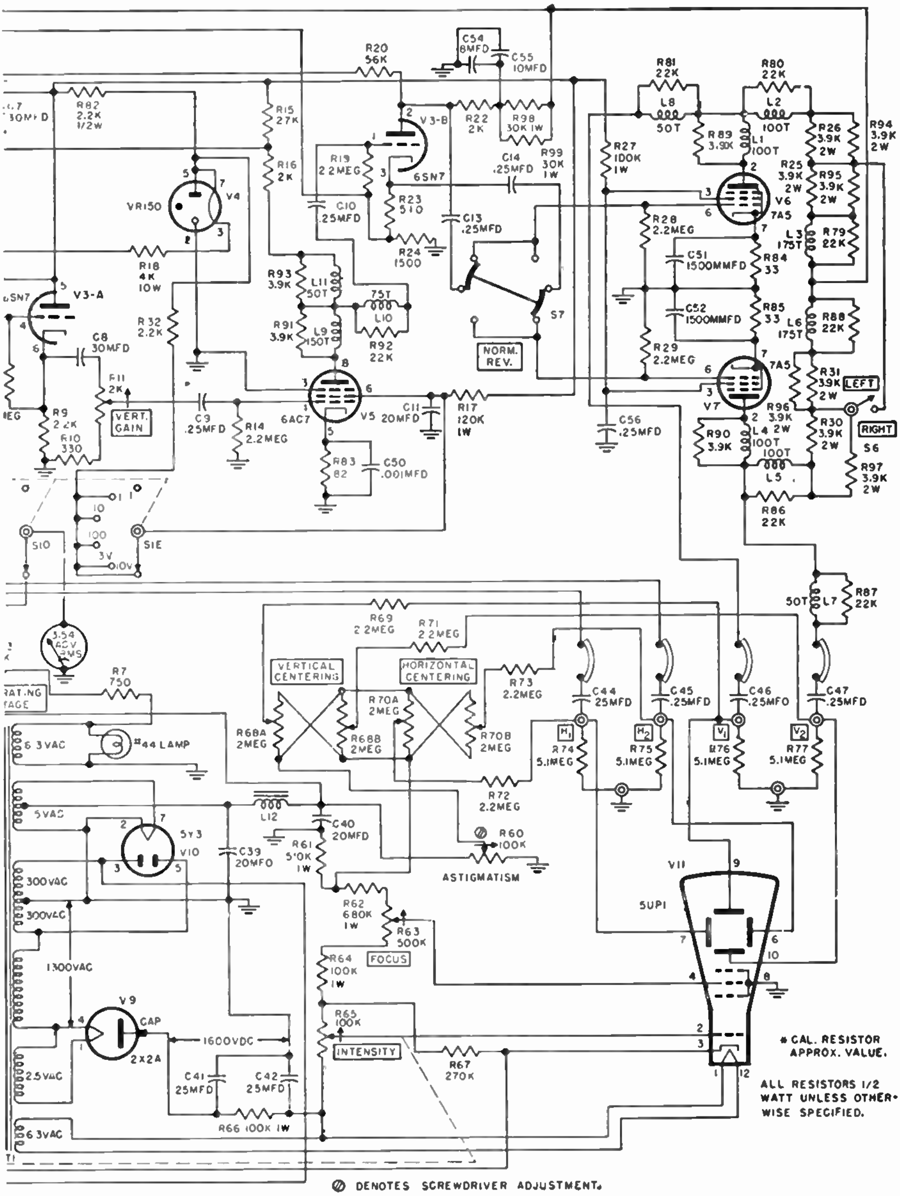

VERTICAL INPUT
GND
R21 68
C1 .25MFD 900V MAX
SAWTOOTH OUTPUT
C6 .25MFD
PHONES
GND
TO 6.3V
PHASING
R33 500K
C15 .01MFD
SYNC & HORIZ
GND
S2A
S2B
S2C
S2D
C3 70MMFD
C2 4.5-25MMFD
R2 220K
R1 2MEG
C4 .001MFD
C5 4.5-25MMFD
R3 20K
R4 2MEG
S1A
S1B
S1C
R40 39K
R36 1K 1W
C17 20MFD
R35 10K 2W
C20 8MFD
R41 10K 1W
C18 20MFD
HORIZ. GAIN
R38 50K
C21 .25MFD
6SN7
V1-A
V1-B
C16 .25MFD
C19 .25MFD
R34 10MEG
R37 750
R39 2.2MEG
R42 10K 1W
R43 2.2MEG
R45 680
R44 2.2MEG
C22 .25MFD
C53 .25MFD
R46 47K 1W
V2-A
6SN7
V2-B
R47 47K 1W
C57 .1MFD
C58 .1MFD
R5 2.2
SYNC. SIGNAL
R53 50K
C27 510MMFD
C26 .01MFD
C23 200MMFD
R50 510K
C24 4.5-25MMFD
R49 5.1MEG
C29 .06MFD
C30 .008MFD
C31 .001MFD
C32 70MMFD
R57 33K 1W
R6 5K
R12 2140
CALIB VOL
S3A
S4 CLOSED FOR BLANKING
INPUT HORIZ.
PHASED 60~
INT
SYNC LINE
EXT.
C25 .03MFD
C48 .05MFD
Z-AXIS
R54 100K
C43 27MMFD
C33 1MFD
C34 .15MFD
C35 .015MFD
C36 .002MFD
S3B
12AT7
V8
R56 68K
R51 100K
R55 470
R58 22K
R59B 1MEG
R52 470K
SWEEP RANGE
10-60
60-600
600-6K
6K-60K
S3C
R59A 500K
SWEEP VERNIER
3AMP
C37 .02MFD
C38 .02MFD
S5
117V 50-60~
C49 .03MFD
R78 2MEG
R94 220K
STRAY

FIG. 295



Oscilloscope Schematic

output terminal, phonejack, and phased 60-cycle sweep position with phasing control.

151. *Vacuum-Tube Voltmeter*

A vacuum-tube voltmeter or at least a 20,000-ohms-per-volt voltmeter is used to make the general run of voltage and resistance measurements necessary when trouble-shooting a defective stage of a television receiver. It is important that the meter be a high-resistance one to permit measurement of d-c components of voltage in high-impedance circuits. To improve versatility of measurements which can be made, an a-c probe can be of considerable benefit because it can be used to measure a-c components of signal over a prescribed frequency range.

In the use of an a-c probe and vacuum-tube voltmeter, it is necessary to know the impedance of the instrument at various frequencies and to prevent loading of circuits under measurements, resistive and reactive components. The actual capacitive component changes with frequency because of Miller effect; resistive components change with transit time. Again, it is very important to know your instrument and all its features and characteristics if you are to use it to advantage.

So far as frequency response and loading are concerned, when using the instrument, observe the same cautions you would in the use of an oscilloscope. When measuring high-frequency signals for a true reading and for a comparison with measurements at other frequencies, attempt to obtain the readings across a low-impedance circuit to prevent loading. Again, if a high frequency is to be measured, it is important that it is delivered to the instrument and not shunted by loading effects of meter.

METER SENSITIVITY

Another important characteristic of a meter is its sensitivity and range scales; choice of instrument, again, is dependent on what measurements are to be made. If weak radio-frequency signals are to be measured, the instrument must be extremely sensitive. This is a very important consideration when the instrument is used for alignment purposes and for measurement of received signals.

152. *Commercial Vacuum-Tube Voltmeter*

A typical commercial voltmeter (Fig. 296), which can be adapted for many television applications, is the RCA WV–75A. This instrument includes an external diode probe for measurement of a-c signals and can be used as an a-c or d-c vacuum-tube voltmeter or as an ohmmeter.

A voltmeter consists of a vacuum-tube bridge; tubes *V1* and *V2*, which have a constant plate voltage when no signal is applied to either grid; meter

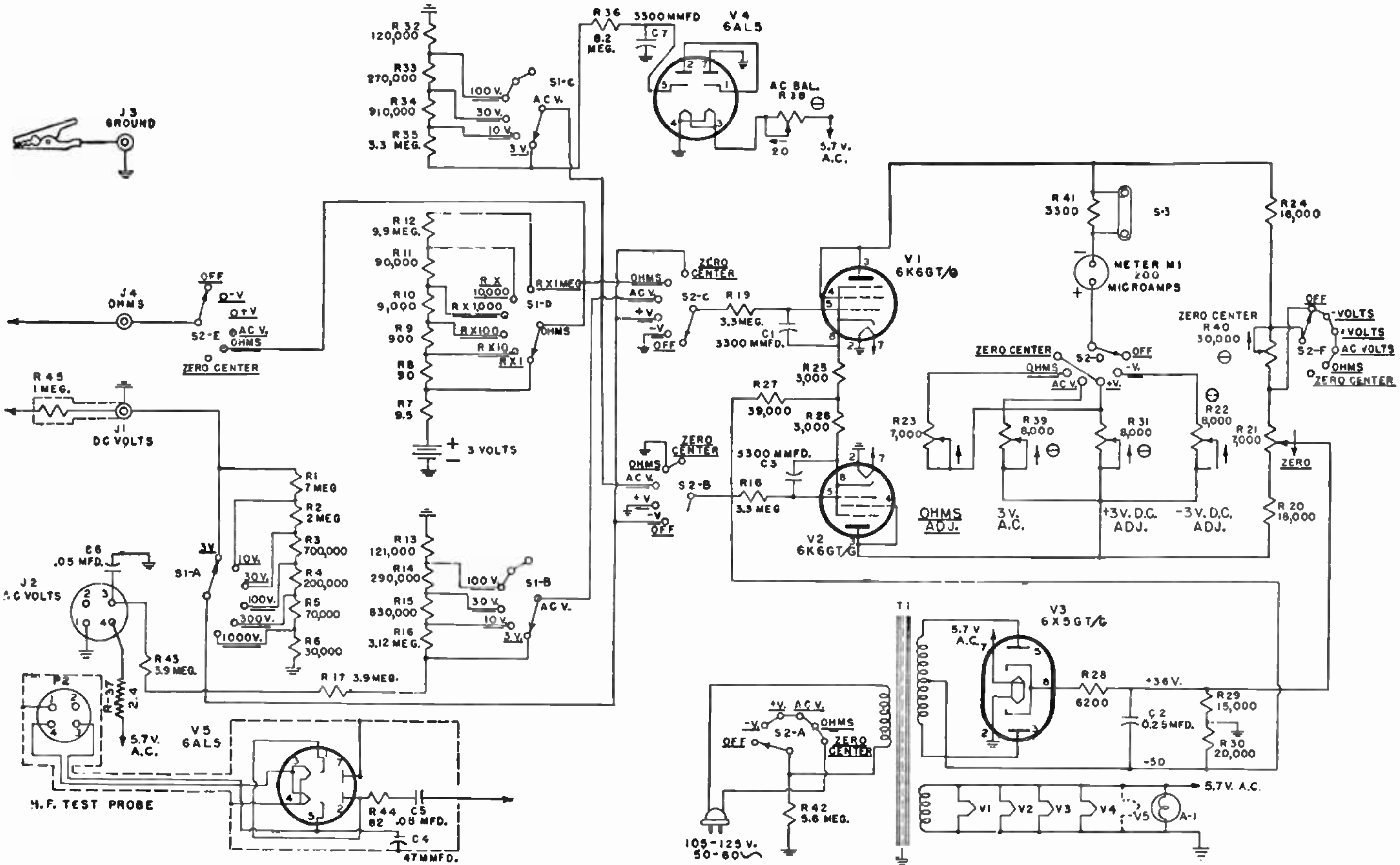


Fig. 296 RCA Vacuum-Tube Voltmeter

connected from plate to plate reading zero. Upon application of a d-c component of signal positive to grid of *V1* or negative to grid of *V2*, the plate voltages change and a d-c component of current flows through the meter in accordance with the potential applied to one grid (which grid is dependent on whether the voltage to be measured is positive or negative). An unusual feature of the instrument is the zero center adjustment, *R40* and associated circuits, which permits adjustment of the plate circuit of the bridge to have the meter indicator read at the center for certain d-c measurements (particularly for alignment of a discriminator). In this position of the meter, the indicator deflects to right or left depending on the polarity of the direct voltage to be measured. The meter is centered with no signal applied by slightly unbalancing the bridge with potentiometer *R40*.

In measuring a-c, a diode miniature tube, built into a probe, is attached almost directly to the point from which the signals to be measured are taken; thus, the high-frequency signal is rectified immediately, and the lead between the probe and meter does not affect the loading of the a-c signal source. The rectified voltage is applied to the grid of tube *V1*, and meter is calibrated to read an rms equivalent of the peak voltage rectified by the diode probe.

To prevent fluctuations in circuit components and line voltage from affecting zero centering position of the meter and d-c component of rectified current contributed by the a-c probe rectifier, an additional rectifier tube similarly connected, but internal, supplies a d-c component of current to the lower tube of the bridge which balances out any average change in the d-c component of current contributed by the probe.

A specific advantage of the vacuum-tube voltmeter is that the diode probe rectifies a peak current and develops voltage proportional to peak voltage of the signal measured and, therefore, would also rectify a voltage proportional to the peak amplitude of a transient waveform. Thus, the peak amplitude of a complex waveform can be obtained by simply multiplying the rms reading on the meter by 2.83 (1.414 times 2). Consequently, with this instrument, if duration of various pulse waveforms is held at a minimum with respect to their interval spacing, an accuracy within 2 per cent is obtained. Most of the television waveforms likely to be observed fall within these requirements; therefore, the instrument can be used to measure complex waveforms with reasonable accuracy. Some of the characteristics of this instrument are as follows.

1. Eleven-megohm input resistance for all d-c scales. On lowest voltage scale (0 to 3), sensitivity is 3.7 megohms per volt.

2. The input impedance presented when using the voltmeter for a-c measurements with the diode probe connected directly—at 1 megacycle, 625,000 ohms, and 15.6 micromicrofarads; at 10 megacycles, 32,000 ohms and 14.5 micromicrofarads; and at 25 megacycles, 100 ohms and 13 micromicrofarads.

It is apparent, therefore, that the loading effect of the meter becomes more pronounced at the higher frequencies. Consequently, it is important to understand that if a higher frequency signal is to be measured and delivered to the input of the meter, it must be taken off at a relatively low-impedance point if readings are to be entirely accurate and if loading effect on the device to be measured is to be insignificant.

3. The frequency response of the instrument, when the diode probe is used directly, is 30 cycles to 200 megacycles; when the diode probe is used with supplied leads, 30 cycles to 30 megacycles. It is apparent, therefore, that instrument is well adapted to work on video amplifiers and can be conveniently used to take frequency-response measurements on a video amplifier because it is capable of reading with accuracy those frequencies likely to be present in a video amplifier.

4. As an a-c vacuum-tube voltmeter—zero to 3, 10, 30, and 100 volts with the a-c probe; with a special multiplier—zero to 300 and 1,000. As a d-c meter—zero to 3, 10, 30, 100, 300, and 1,000 volts.

153. *Sweep Oscillator*

A sweep oscillator is an essential and effective test instrument for wide-band amplifier alignment. In conjunction with an oscilloscope, a visual representation of the response of such an amplifier is obtainable. This response curve can be conveniently calibrated in frequency with an accurate marker system, which will permit immediate observation of any one frequency with respect to all other frequencies of the bandpass; likewise, it is possible to locate instantly frequencies of various abnormalities, dips, or rises in the response characteristic. A sweep oscillator and its marker system, therefore, permit spot frequency checks as well as over-all response characteristics and indicate, at a glance, bandwidth (range of frequencies which the amplifier can pass) and linearity of the response (flatness of response over a desired spectrum).

To obtain a visual representation of a response characteristic, it is necessary that a wide band of frequencies be applied to the input of the wide-band amplifier and that this frequency band be not applied simultaneously but in a progressive sequence, starting from some high frequency and proceeding to a lower one, or vice versa. Frequencies are again presented from low to high, and then a new cycle of frequency sweep begins. The rate at which the frequencies are scanned is generally 60 or 120 times per second. A simple method to generate such a varying frequency signal is to use a variable capacitor driven by a motor. Thus, as the capacitor plates rotate from an open position to a fully meshed point at a prescribed rate (capacitor is part of oscillator tuned circuit), the frequency generated by an associated oscillator varies from high frequency to low frequency; as the plates begin to move toward full open position, the frequencies generated vary from low to high. Other methods of generating a sweep frequency are by means of a reactance tube modulator or a speaker motor which drives a metallic surface toward and away from the inductor of a tuned oscillator circuit.

An important consideration in generating a sweep of frequencies is that amplitude of all frequencies generated should be uniform to obtain a true response curve. Likewise, the rate at which the frequencies are varied must be such to ensure a reasonably linear time spacing on the oscilloscope screen.

The sweep frequency output of the oscillator is applied to the input of the wide-band amplifier under test, and its output frequencies are made to sweep over a range of frequencies which the amplifier should be capable of passing, plus an additional band of frequencies on each side of the bandpass of the amplifier to obtain a complete response. A sweep oscillator, therefore, must incorporate two adjustments—one to set a center frequency or the frequency about which deviation occurs and one to control the extent of the frequency deviation on each side of this center point.

When the sweep frequency is applied to the amplifier under test, each frequency, as it is instantaneously presented to the amplifier, is amplified in accordance with the gain of the system at that particular frequency. If this signal is then presented to a detector, a signal component is rectified in accordance with the strength of the signal at each instant. Thus, the signal output of the video detector varies as each new frequency having a different gain is presented to the amplifier and remains constant if the gain is the same at the new frequency. If the amplifier has no gain at a certain frequency at that instant, there will be no signal component rectified by the diode detector. If the amplifier has a maximum gain at a specific frequency or frequencies, maximum diode current will flow; for frequencies at which the amplifier has less than maximum gain, a lower amplitude diode current will flow. It is apparent, therefore (Fig. 297), that instantaneous diode current varies in accordance with the gain of the amplifier at the many frequencies presented to it. If these frequencies are presented at a constant amplitude to the input, the variation in the diode current output (and voltage across diode resistor) represents the variation in gain of the amplifier over a prescribed frequency range.

For example, if a conventional signal generator were set on 25 megacycles and a d-c vacuum-tube voltmeter placed across the diode detector of an amplifier having response of drawing A, Fig. 297, maximum diode current would flow and there would be maximum voltage across the diode load resistor. If the signal generator is now shifted toward 24 megacycles and the same output amplitude from the generator retained, a similar amount of diode current and diode voltage would be read. However, if the output of the signal generator were placed on 25 megacycles and its output held again at its constant value, meter reading would be less. If the signal generator were put on 28 megacycles, at which frequency the amplifier has no gain, minimum current would be read on the meter attached across the diode load resistor. Now, if the signal generator were made to sweep over the range of frequencies from 30 to 20 megacycles and back again, the instantaneous diode voltage would vary in accordance with the response of the amplifier. Furthermore, if the signal presented to the input of the amplifier were to sweep over this range of frequencies at a prescribed rate, the diode voltage would trace out the response curves that many times per second. Thus, if the frequency at the output of the sweep oscillator varies from high to low and back to high again 60 times per second, the instantaneous voltage across the diode resistor traces the response curve that many times per second.

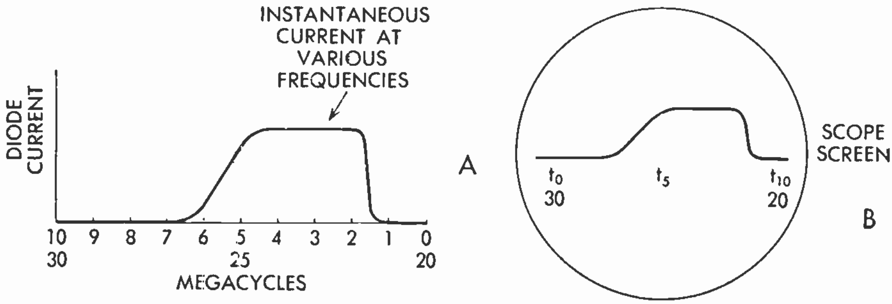


FIG. 297 Diode Current and Response Curve on Scope

If this signal is applied to the vertical deflection system of an oscilloscope, the beam is deflected vertically in a corresponding manner this many times per second. What is more important, if oscilloscope beam moves left to right across the screen at same rate the frequency at output of the sweep oscillator varies from high to low, the horizontal axis of the oscilloscope becomes calibrated in time and frequency. For example, if it takes the oscilloscope 1/120 second to move from left to right across the screen (it would do so if the horizontal waveform increased from minimum to maximum in 1/120 of a second), time t_0 on the oscilloscope trace would occur when output frequency of the sweep oscillator is 30 megacycles. Exactly 1/120 of a second later, at time t_{10}, the output frequency of the sweep oscillator would be 20 megacycles. At time t_5 frequency of the output of sweep oscillator was 25 megacycles. Thus, as the beam sweeps horizontally from left to right, it is setting off equal segments of frequency; if scope vertical input is attached to output of the video detector, signal output of the video detector deflects the beam vertically in accordance with gain of amplifier for each frequency instantaneously applied.

It is very evident that to meet these relations it is necessary for the horizontal motion of the beam to be synchronized in frequency and phase with the sweeping frequencies at the output of the oscillator. For example, in a typical sweep generator, the oscillator is frequency-modulated by a sine wave of 60 cycles; consequently, the oscillator frequency is deviating about center frequency sinusoidally. When the sine wave is at minimum (Fig. 298) time t_0, let us assume that the output of the sweep oscillator is at its highest frequency (for a prescribed center frequency setting). At the instant this highest frequency is applied to the amplifier under test, a similar 60-cycle sine wave is being applied to the horizontal deflection system of the oscilloscope, which is also at minimum and is holding the beam on the far left side of the screen at that instant. As time progresses from time t_0 during the period of the sine wave, it is frequency-modulating the sweep oscillator in such manner as to have frequency decrease. Likewise, the similar sine wave applied to the oscilloscope horizontal is increasing in amplitude and therefore moving the beam left to right, horizontally. In like manner, the output of the video detector is changing in accordance with the gain of the amplifier at the frequency being applied instantaneously, the variation being applied to the vertical deflection system of the scope, causing the beam to deflect vertically.

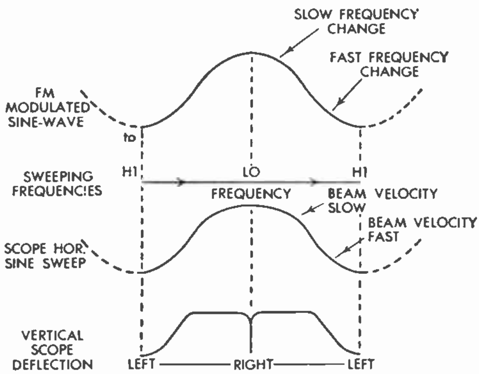


FIG. 298 Timing of Sweep Frequencies and Scope Sweep

When the frequency-modulating sine wave is crossing its zero axis, sweep oscillator frequency, at that instant, is the center frequency to which the oscillator has been adjusted. As its sine wave continues to increase in amplitude, the frequency at the output of the sweep oscillator now decreases on the other side of center the same amount as the deviation on the high-frequency side. In one cycle of the modulating sine wave, frequency has varied the oscillator from high frequency to low and back to high again. In the same time, the sine wave applied to the horizontal deflection system of the oscilloscope has varied from minimum to maximum and back to minimum; therefore, the beam has moved from left to right across the screen and back again at the same rate from right to left. Thus, the beam traces across the screen and back again over the same path.

To make the trace and retrace precisely overlap, and therefore properly phased with respect to the output of the video detector, a phasing adjustment is generally included with the sweep oscillator to be adjusted until the two patterns overlap precisely Actually, the phasing adjustment is setting the phase of the sine wave at the horizontal deflection plate to match the phase of a similar sine wave which, at the sweep oscillator, is generating a sweep frequency by frequency modulation. Inasmuch as the horizontal sweep for the oscilloscope is also generated in the sweep oscillator, it is occurring at the same frequency as the rate at which the oscillator is being swept; consequently, when properly phased, a stationary, rigid pattern is formed on the screen of the oscilloscope.

Another consideration is the shape of the frequency-modulating wave and the wave at the horizontal deflection system. For example, if a sine wave is used to frequency-modulate the oscillator, the rate of frequency change when the sine wave passes through zero is faster than the rate of change at the peak of the positive and negative alternations of the sine wave. Thus, if a linear frequency scale is to be set off on the oscilloscope screen, it is necessary that the velocity of the beam horizontally vary identically. This is actually the case when a sine wave is also used to deflect the beam horizontally, for the rate of horizontal motion of the beam is lower at the peak of the sine-wave alternation than it is when the sine wave is going through zero or when the beam is crossing the center axis of the scope screen. In the older, motor-driven type of sweep oscillator in which a variable condenser was driven by a motor, it was necessary to distort horizontal sweep waveform intentionally to obtain a linear pattern. The amount that the horizontal waveform was distorted to produce a linear frequency scale was dependent on the changing rate at which the rotating capacitor varied the oscillator frequency. Most capacitors have a faster rate of frequency change as the capacitor plates approach the open position, and thus the horizontal deflection wave form had to be compressed and the beam velocity speeded up whenever the rotating capacitor approached the fully open position.

MARKER SYSTEM

The response waveform on the screen of the oscilloscope is only approximately calibrated, the center of the trace representing the center frequency at which the sweep oscillator has been set. Extremities of the sweep can be approximated by adding and subtracting from the center frequency the maximum frequency deviation for which the sweep oscillator has been adjusted. A more accurate means of calibrating the response curve is to employ an accurately calibrated signal source, which can be internal or external to the sweep oscillator. Sometimes this calibrated source takes the form of crystal oscillators of various frequencies which can be switched in and out to calibrate the curve with a low-amplitude oscillation or notch or it can be a continuously variable signal generator (accurately calibrated, of course) producing a tunable notch which can be run across the actual response curve, its position on the response curve representing the frequency to which it is set. Thus, any frequency point on the response curve can be ascertained by setting the signal generator on that frequency. The presence of the continuous-frequency signal in the wide-band amplifier under test adds to or subtracts from the sweep frequency whenever it sweeps past this frequency. The marker frequency is generally set at a very low amplitude in comparison to the sweep oscillator signal, producing only a slight niche or two-sided spike in the response curve. Whenever the signal generator is varied, the niche moves about the response curve, calibrating the various frequency points.

It is also possible to put a niche in the curve by simply positioning a resonant circuit near the output circuit of the sweep oscillator to absorb a small amount of signal at the frequency to which it is tuned whenever the sweep oscillator passes through this particular value.

In summation, the sweep oscillator and associated components have the following features:

1. The sweep oscillator emits a frequency-modulated signal which varies above and below the prescribed center frequency at a low audio rate. It is essential that the output of each frequency over which the oscillator sweeps have a constant output, thereby applying frequencies of constant amplitude to the input of the amplifier under test. It is also important that the sweep oscillator be well shielded and spurious radiation kept at a minimum to prevent interference in nearby receivers.

2. The sweep oscillator should generate an audio frequency which has the same wave shape as the frequency modulation which deviates the sweep oscillator. This waveform will cause the horizontal motion of the oscilloscope beam to follow the variation in the rate of frequency change, setting off a linear frequency scale on the oscilloscope. A means should be incorporated to set the phase of the horizontal deflection waveform to have it phased properly at the deflection plate with the frequency swing at the output of the sweep oscillator.

3. The output of the sweep oscillator must have proper attenuation control to supply sufficient signal to check response of a low-gain system as well as a weak signal output so as not to overdrive a high-gain system.

4. If possible, an accurate marker system should be incorporated to permit precise calibration of the response waveform appearing on the scope screen. An accurate center frequency adjustment of the sweep oscillator output has many advantages.

5. A sweep width adjustment must be incorporated to control the deviation of a sweep oscillator in accordance with the bandwidth of the system under test. It is very advantageous to have this calibration accurate also.

6. The sweep oscillator should have a low-impedance output for television application because it prevents any circuit to which it may be attached from affecting the linearity of the sweep frequency output.

154. *Sweep Oscillator Types*

A number of basic oscillator systems are employed. They can be segregated into two specific categories—a mechanical system and an electronic frequency-modulation system. The two basic mechanical systems are shown in the block diagram of Fig. 299.

The first mechanical sweep oscillator type consists of a high-frequency oscillator, the tuned circuit of which contains a motor-driven capacitor which, as it rotates, varies the resonant frequencies of the tuned circuit. The output of the oscillator sweeps in frequency from some minimum value to the maximum value at a rate set by the number of revolutions the motor makes. By means of a selector switch which changes the inductance of the tuned circuit, oscillator output on the VHF television channels can be obtained.

When it is desired to sweep in the i-f frequency range, the output of the oscillator sweeps over a range from 80 to 90 megacycles. This output is applied to a detector mixer. To this same mixer another output from a fixed oscillator is also applied and beats with the swept frequencies to produce an

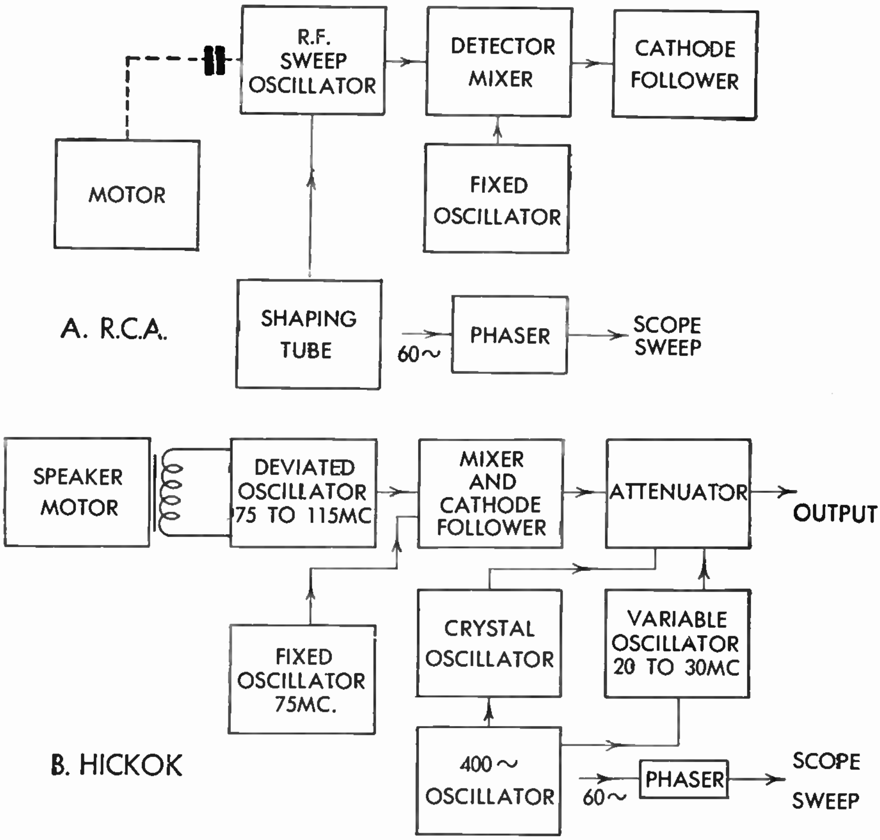


FIG. 299 Mechanical Sweep Oscillators

i-f frequency sweep. This new frequency sweep is applied to a cathode follower clipper and then to an output jack. By proper choice of fixed oscillator frequencies to mix with the sweeping frequencies from the main oscillator, the various television and commercial FM i-f ranges can be obtained.

The output frequencies of the sweep oscillator are swept over the desired frequency range at a prescribed rate depending on the rotation of the motor-driven capacitor, which is generally 60 or 120 times per second. A second output from the sweep oscillator, therefore, must be a 60- or 120-cycle sine wave, which is applied to the horizontal input of the oscilloscope. Its phase is properly regulated to have the start of the oscilloscope sweep from the left-hand side of the screen coincide with the start of the r-f or i-f frequency sweep at output of sweep oscillator.

A second mechanical system is used by Hickok, block diagram of which is shown in drawing B. In this system the deviated oscillator has its tuned-circuit inductor positioned near a dynamic-speaker motor drive which moves a metallic surface away from and toward the inductor. Consequently, the frequency of the oscillator is deviated. Center frequency of the oscillator is controllable over a range of 75 to 115 megacycles.

To permit setting output frequency at any point over the television spectrum, an additional 75-megacycle oscillator is incorporated. This oscillator mixes with output of the deviated oscillator and either sum or difference frequencies, or harmonics can be chosen at output. A cathode-follower output of the mixer-limiter stage applies the signal to the attenuator which controls the signal output amplitude over a wide range.

An internal marker system is incorporated in this unit; either a choice of crystal oscillator signals on chosen frequencies or a variable oscillator tunable from 20 to 30 megacycles can be used for the television i-f spectrum. Either of these oscillators can be tone-modulated and used to obtain (amplitude modulation) a continuous frequency output from the generator, obtained by turning off the dynamic drive. Speaker drive, of course, is under control of a 60-cycle sine wave; a similar 60-cycle sine, after passing through a phaser, contributes the horizontal deflection waveform.

ELECTRONIC FM SWEEP OSCILLATORS

Two basic electronic types are shown in Fig. 300. In the U.S. Television oscillator (drawing A), a reactance tube is used to frequency-modulate a 135-megacycle oscillator, actual frequency modulation consisting of a 60-cycle sine wave. Deviation of the oscillator is controlled by regulating the amplitude of the modulating sine wave applied to the reactance tube. To permit center frequency adjustment over a wide range, a tunable oscillator, which can be set at any frequency between 135 and 245 megacycles, is applied along with output of the frequency-modulated oscillator to the mixer, difference frequency constituting the output of the sweep oscillator.

An internal marker system is incorporated. It consists of two crystal oscillators, one on 1 megacycle and another on 10 megacycles. Each or both oscillators can be used to supply marker calibration, putting a calibrating niche every 1 or 10 megacycles along the response curve.

An extremely versatile sweep oscillator (drawing B) has been developed by Kay Electric. It consists of two 3-centimeter (10,000 megacycles) klystron oscillators. One 3-centimeter oscillator is called the *signal oscillator* and sets the center frequency of the output; a second 3-centimeter oscillator, called a *local oscillator,* supplies the frequency-modulation signal. Local oscillator is frequency-modulated by applying a sawtooth sweep to its repeller plate. This is the customary and a convenient method to frequency-modulate a klystron.

Klystron local and signal oscillators feed through appropriate attenuators to a crystal-mixer stage which passes the difference frequencies to the output. Thus, output of sweep oscillator is actually a difference frequency between the two microwave oscillators. This output is the useful signal output of the sweep oscillator, and it is, of course, frequency-modulated because the local oscillator is frequency-modulated by a low-frequency sawtooth voltage.

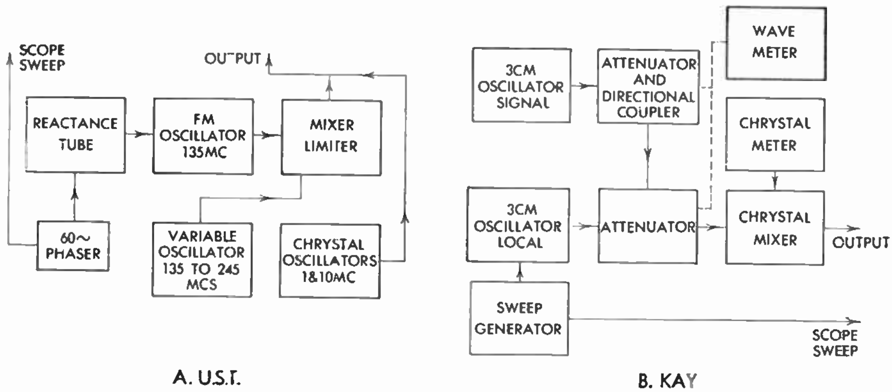


FIG. 300 Electronic FM Sweep Oscillators

Associated with the attenuator at the output of the signal oscillator is a so-called "directional coupler" which attenuates the signal oscillator output substantially and prevents interaction between the two oscillators. This expedient prevents the local and signal oscillators from interacting on each other and consequent frequency pulling. A reduction in amplitude of the signal contributed by this oscillator makes it dominate in the output, as the weaker signal always does in a mixer circuit. Thus, it is the signal oscillator frequency which is varied manually to determine the difference frequency that appears in the output to act as center frequency, and it is the local oscillator which is frequency-modulated and usually set at a fixed frequency. This fixed frequency is frequency-modulated and in turn frequency-modulates the difference frequency of the output, center frequency of which is controlled by setting of signal oscillator.

A crystal meter is associated with the crystal mixing circuit to permit setting of oscillator frequencies for efficient operation. An absorption coaxial wavemeter is used to measure the oscillator frequencies which appear at output.

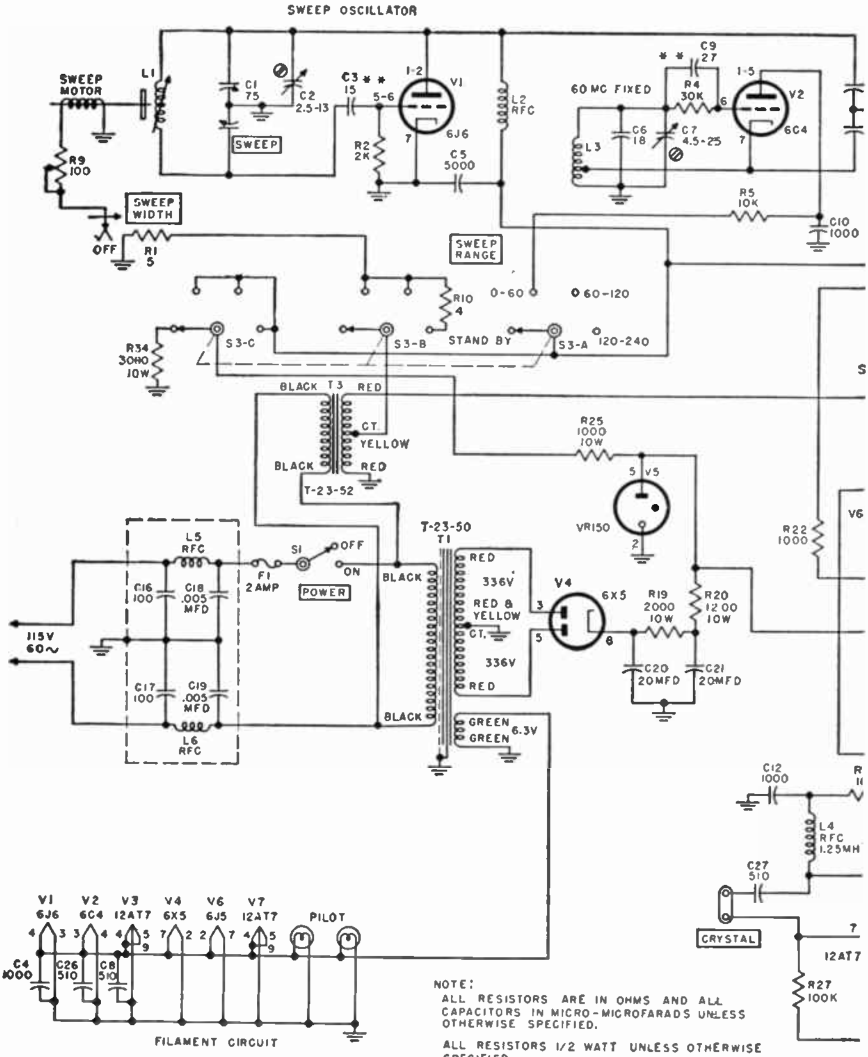
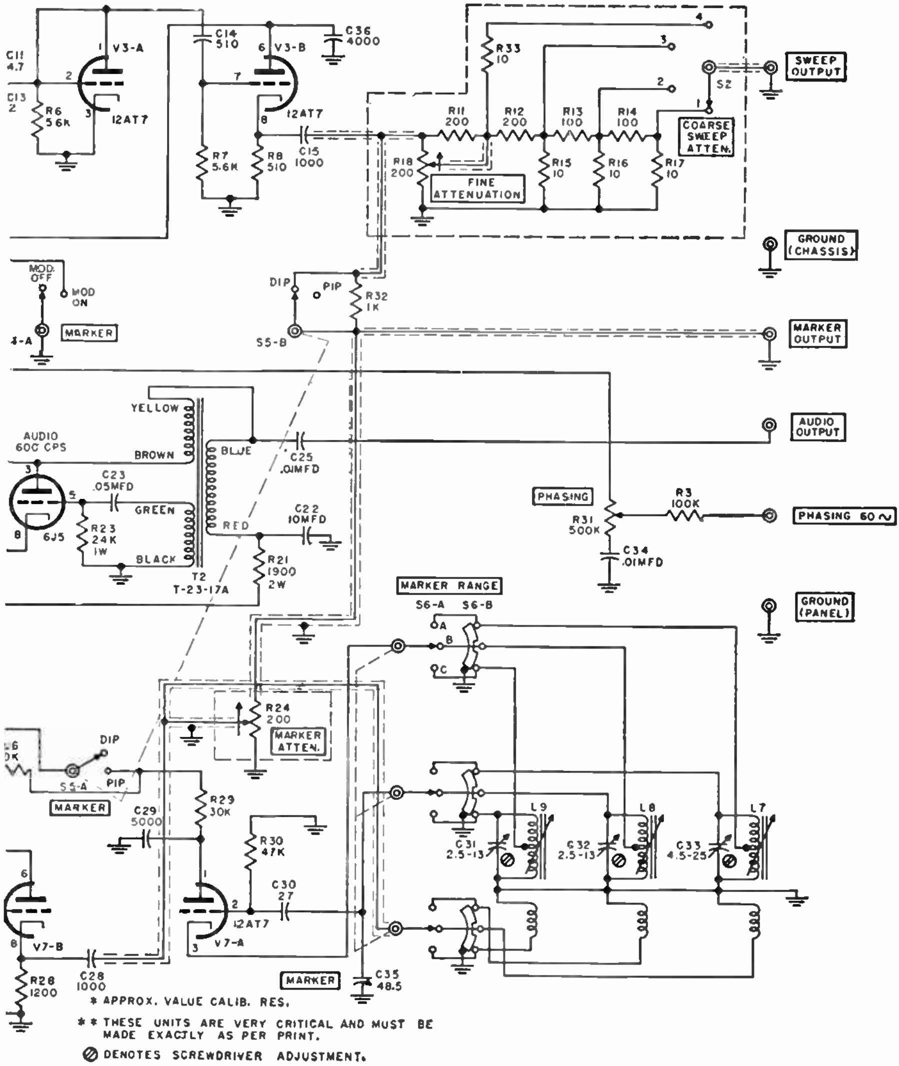


FIG. 301 Triplett Sweep-



Signal Generator Schematic

155. *Commercial Sweep Oscillators*

An effective sweep oscillator for the television technician is the Triplett model, shown in Figs. 301 and 302. One-half of a 6J6 miniature tube is used as a frequency-modulated oscillator with its center frequency adjustable to any frequency between 60 and 120 megacycles. It is an ultra-audion oscillator, and its plate-tank circuit (tuned circuit between plate and grid) is placed in close proximity to a moving metallic surface driven by a dynamic motor.

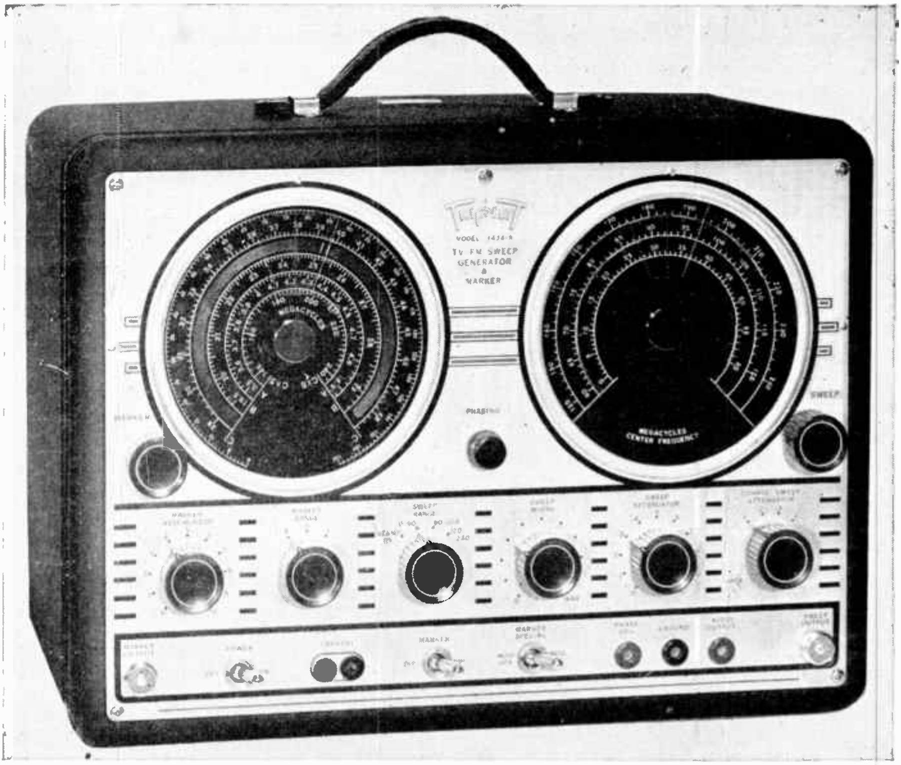


FIG. 302 Triplett Sweep and Marker Generator

When there is no excitation to the motor, the oscillator has a single-frequency range between 60 and 120 megacycles. However, with excitation applied (the extent of which can be controlled by resistor *R1* that connects to a 60-cycle sine-wave source), this oscillator can be deviated to a maximum of 12 megacycles. The oscillator is capacitively-coupled to a diode mixer and clipper, tube V3, which limits amplitude variations of the sweep-frequency output. The sweep signal is then applied to a triode clipper-limiter with a cathode-follower output that feeds a low-impedance attenuator circuit. Sweep-frequency output is available on the fundamental range from 60 to 120 megacycles and over the second harmonic range from 120 to 240 megacycles. A 0- to 60-megacycle sweep range is obtained by beating the output of the sweep-oscillator tube with a 60-megacycle fixed-frequency oscillator, tube V2, at the input to the diode clipper—the signals meeting at the junction of capacitors *C11* and *C13*.

A fundamental marker-oscillator is provided to supply an accurate marker signal over the useful ranges required in television alignment—3.5 to 4.9 megacycles, 19.5 to 29 megacycles, 24 to 48 megacycles, and harmonic ranges up to 240 megacycles. Marker output is supplied through resistor *R32* to the sweep-signal output. A separate marker-output terminal is also provided for applying the marker signal directly into specific circuits for stronger indication. The variable marker is augmented with both a Pierce crystal oscillator and its external crystal jack into which specific crystals can be inserted for calibration and for dual accurate marking of response curve. Fundamental crystals with oscillating frequencies in the 30-megacycle range can be inserted also, providing harmonic output well up into the VHF- and low UHF-frequency ranges.

Inasmuch as a 60-cycle sine wave is used to excite the motor drive, a similar 60-cycle sine wave must be applied as horizontal sweep to the oscilloscope if the scope does not have an internal 60-cycle sine-wave source. Furthermore, the phase of this sine wave applied to the horizontal sweep must be correct with respect to the phase of the deviation of the oscillator. A phasing control *R31* permits adjustment of the phase of the sine wave applied to the scope and consists of resistor *R31* and capacitor *C34,* their relative reactance and resistance ratios determining the phase of the sinewave. This type of generator is an ideal instrument for alignment, because it is built as a compact self-contained unit which incorporates the various types of signals necessary in the alignment of a television receiver. It contains sweep oscillator output in r-f and i-f ranges and, in addition, can be used as a single-frequency generator with accuracy over the same frequency ranges. Additional features are a 600-cycle audio oscillator for modulation of marker and added clarity and an absorption-marker circuit that can be used on the low-frequency marker range with the oscillating marker turned off.

RCA SWEEP OSCILLATOR AND TELEVISION CALIBRATOR

The television sweep oscillator and calibrator (Figs. 303 and 304) form a versatile combination which can be used for alignment adjustment and trouble shooting of television receivers. The sweep oscillator has a frequency sweep on each of the 13 television channels, an i-f frequency sweep between 5 and 15 megacycles, an i-f frequency sweep between 20 and 30 megacycles, frequency sweep between 20¼ and 22¼ megacycles for sound i-f alignment, a 10-to-11½ megacycle sweep for commercial FM receiver sound i-f.

A 6-megacycle sweep can be obtained at any point within the 25-to-50 megacycle range with center frequency adjustable to any point within that

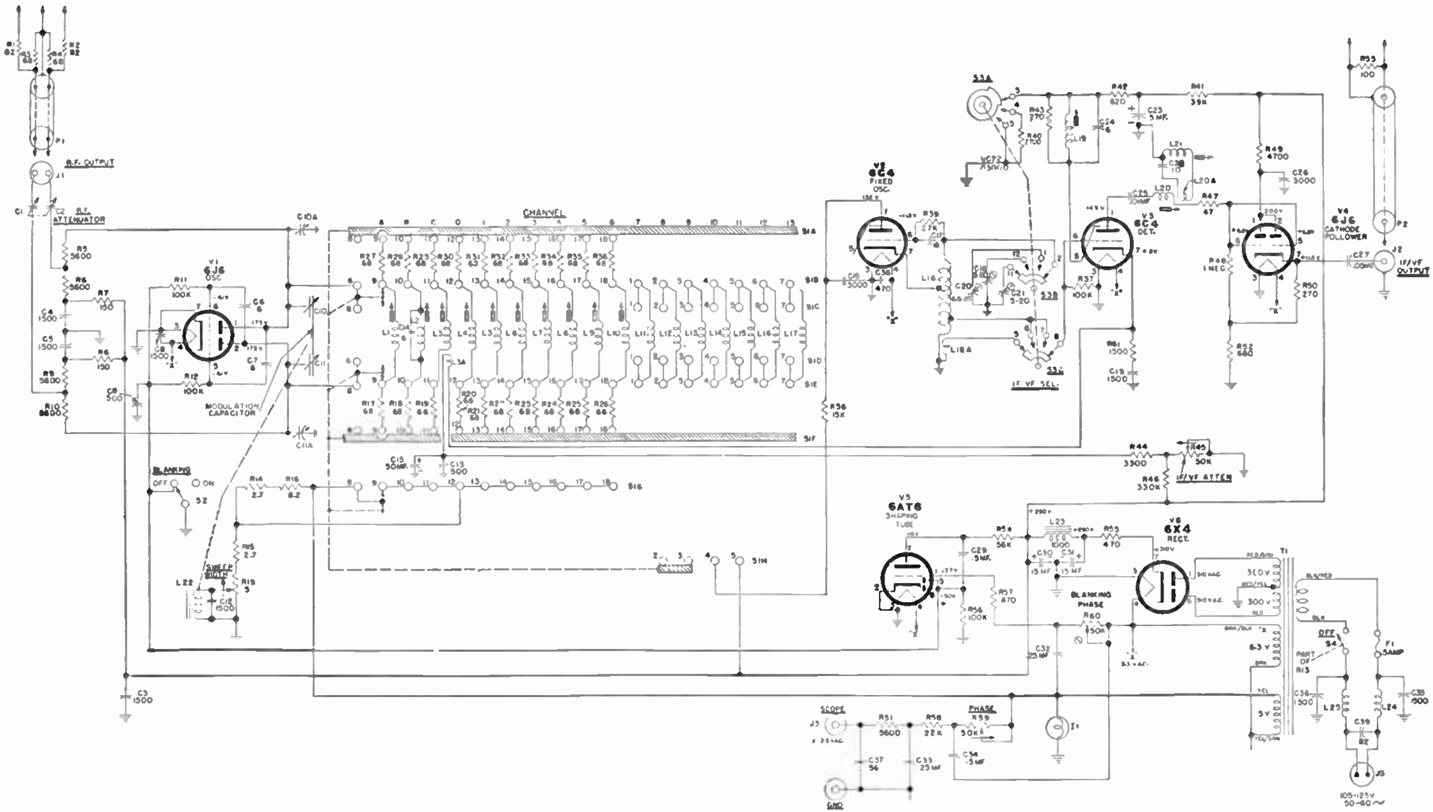


FIG. 303 RCA Sweep Oscillator

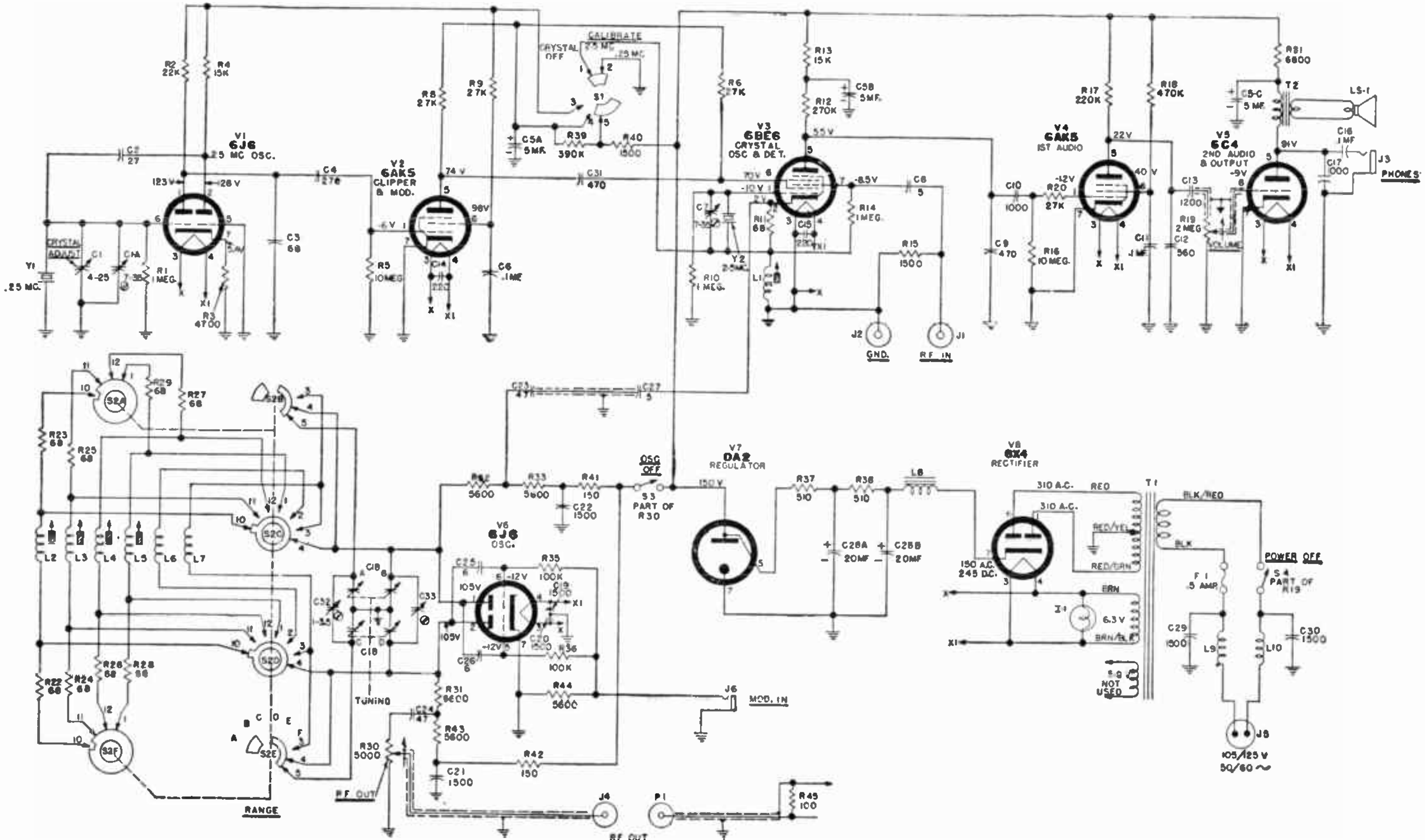


FIG. 304 RCA Signal Calibrator

range. To check video amplifier response, a special zero-to-10 megacycle sweep is also available and can be used to obtain the response curve of a video amplifier in a television receiver. The output voltage of the sweep oscillator is at some level between 0.1 and 0.18 volt.

The main oscillator consists of a 6J6 connected as a push-pull oscillator with a tuned-plate circuit consisting of a motor-driven capacitor and a series of inductors that are switched across this capacitor in accordance with the frequency range desired. Thus, on each new channel a new inductor is switched across the rotating capacitor to produce the frequency sweep required. Sweep width can be controlled by regulating the current through coil *L22*, which determines the frequency change caused by the rotating capacitor. Output of the oscillator is taken off a resistor divider and coupled through a capacitive attenuator to the r-f output jack. Output cable for the r-f section is terminated in a 300-ohm balanced resistive load.

A special shaping tube, *V5*, is excited by a sine wave applied to its grid which is amplified and applied to the diode rectifier section of the tube. During the positive alteration, the rectifiers draw current and place a negative charge on capacitor *C9* which cuts off the oscillator. Thus, the frequency sweep is from some minimum frequency to a definite maximum value. After it reaches this maximum, the tube is cut off while the driven capacitor continues to rotate until it reaches the point at which the oscillator is again turned on to begin a new frequency sweep at the minimum value. Consequently, the frequency sweep is in only one direction, simplifying adjustment and preventing a dual trace on the oscilloscope screen. The blanking phase control, *R60*, must be adjusted (a pre-set adjustment) to have the oscillator turn on when the capacitor is at its minimum frequency point and to have the oscillator turn off when the capacitor rotates to its maximum frequency point. A 60-cycle output is also present at jack *J3* for application to the horizontal sweep of the oscilloscope. This phase is properly adjusted with phase control *R59*, which ensures that the start of the sweep on the left-hand side of the screen coincides with the minimum frequency point of the frequency sweep at the output of the oscillator.

When the selector is turned to one of the i-f ranges, the 80-to-90 megacycle sweep frequency signal is coupled through coil L3A to the cathode circuit of the 6C4 detector tube, *V3*. At this setting, the fixed-oscillator tube, *V2*, is also turned on and supplies signal to grid of detector tube. This frequency is chosen to have the proper i-f frequency present in the plate circuit of the detector (as a difference frequency) when mixed with the r-f frequency sweep being applied to its cathode. The tuned-plate circuit of the detector is coupled to the grid of the cathode-follower output tube, which has a low-impedance output to match the 100-ohm termination of the i-f and video cable. The i-f attenuator control is in the cathode circuit of the detector, setting the level of the positive external bias applied to its cathode.

One unusual and very helpful characteristic of this sweep oscillator is the zero-to-10-megacycle sweep range which is obtained when an 80-megacycle signal from the fixed oscillator is applied through the detector to beat with the 80-to-90 megacycle swept range applied from the main oscillator. Difference frequencies represent, of course, a range of frequencies from zero to 10 megacycles. The zero-to-10-megacycle sweep range is calibrated with a 3-megacycle marker by means of absorption trap *L21* and *C28* which absorb some of the 3-megacycle components through *L20A*, putting a dip in the response curve at 3 megacycles.

A companion piece to the sweep oscillator is the television calibrator which serves as an accurate marker generator for the sweep oscillator for sweep alignment procedures and as an accurate single-frequency generator with or without audio modulation. It can be used to calibrate the sweep oscillator properly.

It can be set accurately on any frequency between 19 and 110 megacycles or between 170 and 240 megacycles. The output voltage is at least 1/10 volt rms at 100 ohms. The unit contains a variable-frequency oscillator plus two crystal oscillators on 2.5 megacycles and 0.25 megacycle to be used as standards. These crystal stages can be used to check frequency of the variable-frequency oscillator, permitting a very accurate marker frequency signal when the generator is used for sweep oscillator alignment procedure. A heterodyne detector and audio amplifier permit use of the instrument as a heterodyne frequency meter to check unknown frequencies which fall within the frequency range of the instrument.

Scale accuracy of the instrument is within ¼ megacycle from 19 to 110 megacycles and within ½ megacycle from 170 to 240 megacycles. In addition, with the 2½-megacycle crystal oscillator turned on, a beat note will be heard (small built-in loudspeaker) every 2½ megacycles which can be used to set the oscillator precisely on any 2½-megacycle harmonic over the entire frequency range of the variable-frequency oscillator. In addition, after locating one of these 2½-megacycle points, the ¼-megacycle crystal oscillator can be turned on and an intermediate point accurately located. Thus, it is possible to set variable-frequency oscillator precisely on the picture- or sound-carrier frequency of the various television channels quickly and accurately. For example, in locating the picture-carrier frequency of 61¼ megacycles for channel 3, the 2½-megacycle calibrating crystal is turned on, and the variable-frequency oscillator is shifted until the beat note is found at approximately 60 megacycles. Now the ¼-megacycle calibrating crystal is turned on and the frequency of the variable-frequency oscillator is increased until the fifth beat from the 60-megacycle point is found on the dial scale; this should be exactly on 61¼ megacycles. The same method can be used accurately to locate any frequency point over the entire frequency range of the instrument.

The second schematic diagram (Fig. 304) shows the component parts of the calibrator. The first tube at the top left is the ¼-megacycle oscillator connected in a cathode-coupled arrangement and output of which is coupled to a 6AK5 clipper and modulator. The clipper and modulator square the wave to produce a ¼-megacycle signal which is rich in harmonics and therefore produces beat notes to a very high frequency. Output of the clipper ties the ¼-megacycle component to a grid of the multigrid 6BE6 tube. The same tube also incorporates the 2½-megacycle crystal oscillator and a means for inserting an external r-f signal to one of the other grids of the combination crystal oscillator and detector.

The variable-frequency oscillator is a push-pull connected 6J6 with a band-switching and variable-tuning system to cover the desired frequency range. Output is taken off the plate 2 side of the oscillator and through an isolating resistor, *R31*, is coupled to the r-f output jack and cable. Oscillator can be modulated by an external source of audio or video through jack *J6*. The second output is removed from the top plate 1 of the oscillator and is coupled to the cathode of the 6BE6 crystal oscillator and detector.

It is interesting to note that the crystal oscillator and detector tube has four signal input possibilities. When any two frequencies are simultaneously applied to this tube, a difference frequency will appear in the plate circuit which is coupled through the two-stage audo amplifier to the small loudspeaker. For example, if the 2½-megacycle crystal signal and the ¼-megacycle crystal signal are both applied the tenth harmonic of the ¼-megacycle crystal, frequency will compare with the 2½-megacycle crystal frequency to produce an audible difference frequency in the output. The ¼-megacycle oscillator is adjusted to precise frequency by zero-beating this harmonic with the 2½-megacycle crystal frequency. When the variable-frequency oscillator output is beat against either one of these two crystal frequencies, the harmonics of the crystal frequencies will zero beat with the variable-frequency output at various points, the zero-beat point, of course, representing calibrating points throughout the entire variable-frequency range. Furthermore, it is possible to apply an unknown frequency through the r-f input jack, *J1*, and then zero beat the variable oscillator frequency against it to determine the frequency of the unknown signal. This is the method by which the local oscillator in a television receiver can be precisely set on frequency. Do so by first setting variable oscillator on proper frequency and then make the local oscillator adjustment until a zero beat is heard in the output of the small speaker. The combination of the television sweep generator and television calibrator is ideal for the television technician and along with an appropriate oscilloscope can be used for adjustment, alignment, and trouble-shooting of the r-f and i-f sections of the television receiver.

156. *Alignment Checks*

The sweep oscillator is not only a test instrument by means of which a wideband amplifier can easily and effectively be aligned, but, in addition, it can be used to perform alignment checks simply. In most wideband amplifiers, actual alignment because of drift in the resonant circuit frequencies is seldom required. In most cases, there is a defective component part or some part has changed in its capacity or inductive characteristics. For example, a defective tube might cause a change in distributed capacity thereby affecting the band-pass of the amplifier. Replacement of the bad tube with a good one results in proper response curve again. Thus, if there is an incorrect response curve, as indicated by a simple alignment check, the defective component can be isolated and, if replaced, response curve returns to normal. It is very important for the technician to realize that an incorrect response curve is generally not an indication of incorrect alignment but of a defect produced by a bad part.

Occasionally if a tube burns out and is replaced by a new one with a differing interelectrode capacity, the response curve changes shape and realignment may be necessary. Often it is possible to replace the defective tube with a number of substitute tubes and to choose one which does not change the shape of the response curve. By means of alignment check, therefore, it is possible to ascertain if alignment is necessary by observing the over-all response of the amplifier under test and by checking it with the response curve recommended by the manufacturer. When the curve is found to be defective, make absolutely certain that alignment is necessary and that the defective response curve is not a result of a defective part. In practice, a good service man will always touch up the alignment of an AM receiver to obtain maximum performance whenever a receiver is repaired. However, do not attempt realignment of a television receiver unless it is necessary. More than likely it is not.

Successful application of sweep oscillator for alignment, alignment check, or even as an instrument to trouble-shoot for obscure defects is a function of proper utilization of the oscillator and an understanding of the curves presented on the oscilloscope screen. If the instrument is improperly used and the technician's knowledge of what to expect is vague, a sweep oscillator is useless as a test instrument.

PRECAUTIONS FOR TESTING

In connecting the sweep oscillator for operation, the following precautions should be taken if desirable response curves are to be obtained. First connect the test equipment to the amplifier under test as indicated in Fig. 305.

1. Use the balanced or unbalanced output of the sweep oscillator in accordance with the input needs of the amplifier under test. Never apply unbalanced output of a sweep oscillator to a balanced input. A few wide-band amplifier response curves are affected by the input termination, and therefore it is necessary to match output of the sweep oscillator to the input circuit. A 100-ohm unbalanced output sweep oscillator can be approximately matched to a 300-ohm balanced input as shown in Fig. 305. The output of the sweep oscillator feeds into 100 ohms, but the total resistance across the terminals of the input circuit is 300 ohms.

2. A solid and complete grounding system must be used if uniformity of

response curve and accuracy are to be obtained. At the very high television frequencies, even a rather short interconnecting lead has appreciable inductance and is subject to a stray pickup. If such a ground is not available, it is absolutely essential that a heavy ground wire be used to interconnect all units; the grounding system should be confined to as small a space as possible. This is one definite advantage to a sweep oscillator which has all its essential components self-contained, reducing the number and length of interconnecting leads. A poor ground is indicated when shape of response curve changes when ground points are touched or when output amplitude of sweep oscillator is varied (not enough to cause overload).

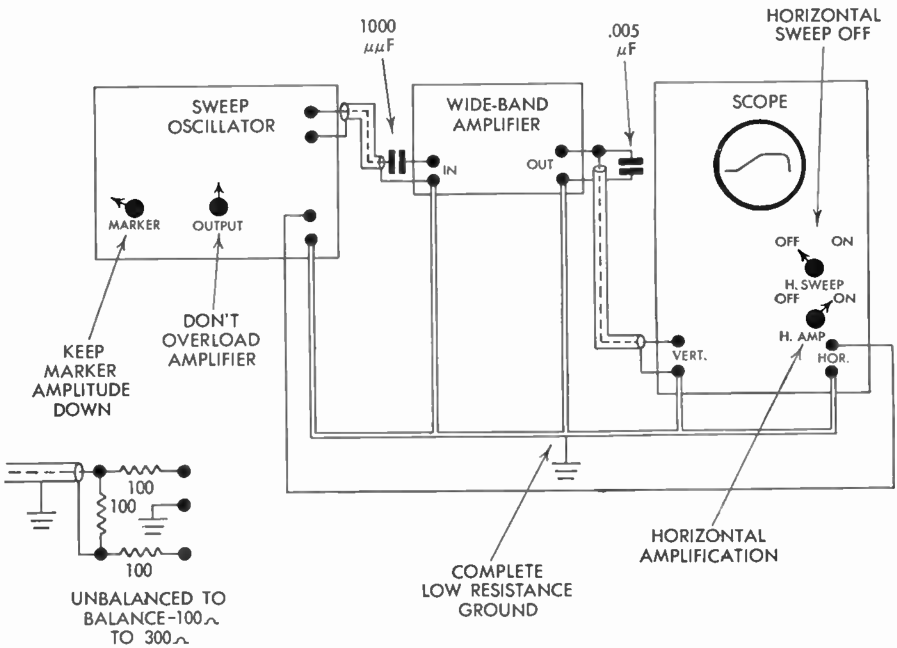


FIG. 305 Interconnection of Units for Sweep Oscillator Alignment

3. Be certain that oscilloscope controls are set for proper operation. The internal sweep must be shut off because the sweep signal is generated by the sweep oscillator. Only occasionally will a sweep oscillator be used which generates a pulse to synchronize the internal sweep of the oscilloscope. If so, it will be indicated on the instrument. The now available television alignment oscillators generate their own sweep voltage. It must be applied to the horizontal input of the oscilloscope, and the internal scope sweep must be turned off. Do not turn off the horizontal amplifier of the oscilloscope.

4. Inasmuch as the response curve is painted on the screen of the oscilloscope at a very low-frequency rate (at the most 120 times per second), the output of the instrument under test can be shunted with a rather large capacitor and can thereby prevent pickup of spurious signals which would disturb the

clarity of the response curve. Try to kill all other signals which might enter the amplifier under test and disturb or add fuzziness to the pattern. For example, in the alignment of the i-f system of television receiver, removal of local oscillator tube as well as horizontal and vertical sweep oscillator tubes helps to improve clarity of the curve.

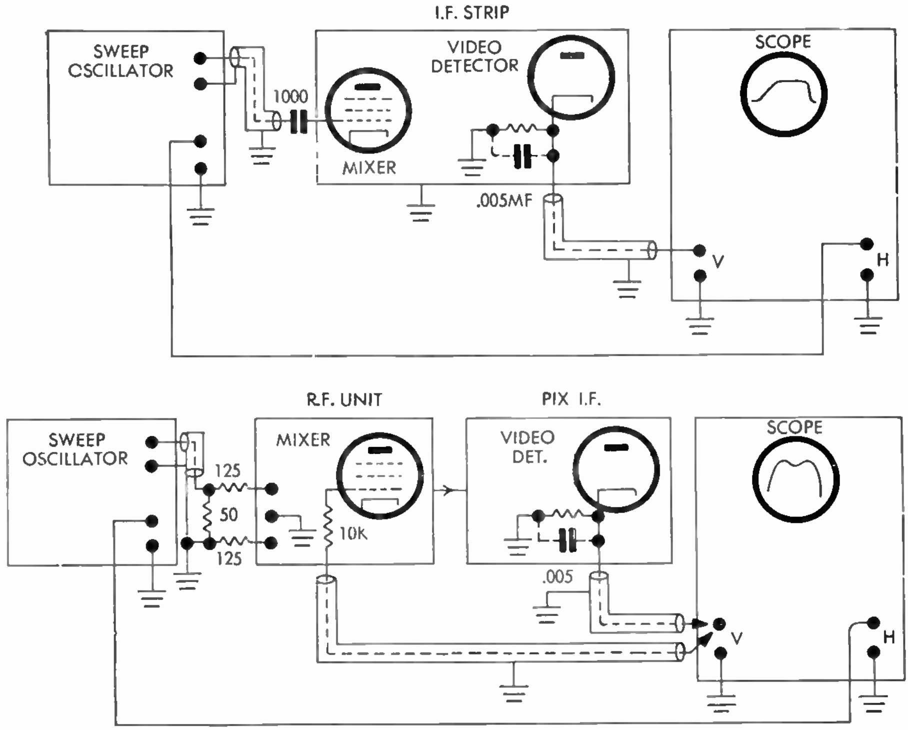


FIG. 306 Sweep Oscillator Interconnections for TV R-F and I-F Alignment

5. Do not overload amplifier under test by applying too much signal from the sweep oscillator. Apply only a minimum of signal required to obtain a clearly visible curve. Overloading of the amplifier will cause an untrue curve which is flattened off or actually compressed.

6. Do not apply too much marker signal, which will also distort pattern and give an incorrect curve. It is best to have marker amplitude at absolute minimum until response curve is obtained and then to apply just sufficient marker amplitude to produce a marker which is visible and which will not disturb fidelity or amplitude of the desired response curve.

Interconnection of equipment to obtain a television picture i-f alignment check curve is shown in the block diagram of Fig. 306. In this circuit, the unbalanced output of the sweep oscillator is attached to the grid of the first picture i-f stage or, whenever possible, to the grid of the mixer tube. Energy is coupled through a small capacitor to prevent shunting any bias circuits with low resistance output of the sweep oscillator. Horizontal output of the sweep oscillator is applied to the horizontal sweep input of the scope horizontal amplifier. Signal, representing the response curve, is taken off the video detector load resistor. It is shunted by a large capacitor to filter out spurious higher frequencies.

In the second drawing, the sweep oscillator is applied to a TV receiver to obtain r-f unit response curve. A 50-ohm unbalanced output is balanced and matched to a balanced 300-ohm input with resistors. Signal for oscilloscope is generally removed at the grid of the mixer. In some receivers the response is checked from antenna input terminals through to video detector. When an oscilloscope is attached to the mixer to prevent shunting that mixer with capacity of the oscilloscope and probe, a 10,000-ohm series resistor is always inserted for isolating purposes.

When an external marker generator is to be used (accurately calibrated signal generator or resonant trap), it must be applied through a small capacitor and series isolating resistor. Proper isolation prevents the presence of the marker generator from affecting the response of the stages in the amplifier. It is also important that the marker oscillator output is kept at the minimum necessary to give a just visible indication on the response curve to prevent an improper indication. The marker oscillator is often attached at the same point that the sweep signal is being applied although occasionally, for a clearer marker indication, it is applied to an earlier stage.

OPERATING PROCEDURE

Typical operating procedure for use of a sweep oscillator is as follows:

1. Permit amplifier and sweep oscillator to warm up for at least 15 to 20 minutes.

2. Properly interconnect equipment and make certain a satisfactory ground has been incorporated. Set sweep oscillator on desired center frequency.

3. Turn up sweep oscillator output to about ¾ full. Turn up deviation in accordance with approximate bandwidth of amplifier under test. Marker amplitude should be at minimum.

4. Adjust oscilloscope controls until pattern appears on screen. Adjust phase control to obtain overlap of dual pattern. Now turn down sweep oscillator output to minimum and gradually increase output, observing pattern on screen. Pattern should increase in amplitude, but its shape should remain uniform. If shape of pattern changes appreciably as the sweep oscillator output is gradually increased, the grounding system is not satisfactory. After the amplitude is increased too far, pattern will be seen to flatten and eventually compress, showing that some of the stages of the amplifier under test have been overloaded. Never keep the sweep oscillator output at this high level. It is preferable in the alignment of most wide-band amplifiers that the minimum sweep oscillator output be used to give a visible pattern on the oscilloscope screen (least attenuation by the vertical amplifier of the oscilloscope).

5. Turn up marker amplitude until marker indentation is visible on the

response pattern. Keep marker amplitude at a level which will keep it discernible and which will not affect fidelity of the response curve.

6. Adjust center frequency output of sweep oscillator and deviation until entire response curve is visible on the screen, extending to a point where the response drops to a minimum. It is important to consider the number of amplifiers present in the oscilloscope as well as the polarity of the response curve at the point it is taken off amplifier output. Under some circumstances, the response curve will appear negative on the screen of the oscilloscope; at other times, depending on polarity of output and number of vertical amplifiers in the scope, it will appear positive.

If the television test bench is properly organized, it should be a simple matter to set up the equipment quickly for performing alignment check, wide-band amplifier trouble-shooting, or alignment. The advantage of using a sweep oscillator method of checking the characteristics of r-f and i-f systems of a television receiver is that the system can be maintained in operation as changes are made and the effect is immediately visible on the curve present on the oscilloscope screen. Thus, suspected parts can be removed and equivalent components substituted and the immediate effect will be apparent on the screen. Likewise, the suspected tube can be removed and a new one substituted and the results observed on the response curve. The response curve should not only give you an indication of the bandwidth and fidelity of the response but also, in case of a weak tube, for example, the amplitude of the response curve should rise when the weak tube is replaced. To see if alignment is necessary, it is only a matter of comparison with the curve recommended by the manufacturer of the amplifier under test. It might be a good policy for the manufacturer not only to indicate the shape of the response curve but also to give an idea of what amplitude should be expected for a given level of applied signal of the sweep oscillator. It will assist in checking the over-all gain of the amplifier as well as the fidelity and bandwidth of the response.

A similar procedure can be used to check the alignment of the sound i-f and discriminator of a television receiver (Fig. 307). When aligning the sound i-f, the sweep oscillator output is applied to the grid of the mixer, if accessible, or to the grid of the first i-f amplifier in the sound section. The response curve can be observed across the limiter grid resistor at which point the input to the vertical amplifier of the oscilloscope is attached. Inasmuch as the bandwidth of the sound i-f is very much narrower than the bandwidth of the picture i-f, deviation of the sweep oscillator is cut down to a very low value (approximately ± 100 kilocycles). To check the response of the discriminator, sweep oscillator output is attached at the same point and the oscilloscope is attached to output of the discriminator. When a sine-wave sweep is used in the sweep oscillator, a single S-curve appears (Fig. 307) on the screen; when aligning a discriminator, it is only necessary to adjust the phase control to obtain a coincident curve. In many of the FM sweep oscillators in which a back-to-back sawtooth sweep is used, a double S-curve shows on the screen.

In most television receivers the sound i-f section is aligned first and, if exactly aligned, can later serve as a standard for proper alignment of the r-f unit of the receiver. Generally, the next in order are the various wavetraps and special tuned circuits in the picture i-f and in the coupling system between

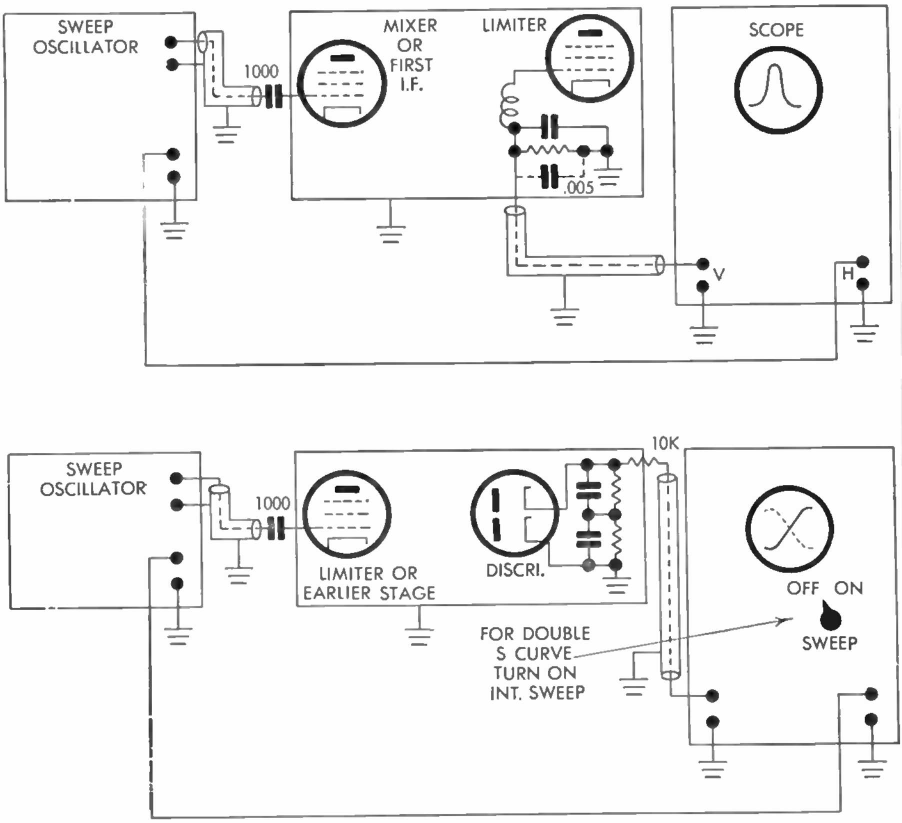


FIG. 307 Sound Channel Alignment with Sweep Oscillator

the picture and sound i-f channels. Next is the alignment of the picture i-f and finally of the r-f unit of the receiver. If the receiver is a projection type, alignment of optical unit follows complete electrical alignment. The above sequence, although not always used, presents a logical manner in which to approach the problems of alignment.

157. *Alignment of Sound I-F Channel*

In the alignment of the sound i-f, a sweep-oscillator or output-meter procedure can be followed. It is advisable to use the procedure recommended by the manufacturer. Equipment for meter alignment of the i-f section of the sound channel is interconnected as shown in Fig. 308.

Procedure for performing meter alignment of the i-f section is as follows:

1. Apply signal generator set to the sound i-f frequency, to the grid of the first sound i-f tube. It is very important that the signal generator be accurately calibrated because alignment of sound i-f not only affects the signal as it passes through the sound system but also influences the alignment of the r-f unit and the various traps in the picture i-f. One plan to be recommended highly is the use of a signal generator which has crystal-controlled frequencies at the sound i-f for the various receiver models handled. If this procedure is used, the technician can be certain of sound i-f alignment and its influence on the alignment of other sections of the receiver.

2. Attach vacuum-tube voltmeter across grid limiting resistor at limiter stage. Adjust various sound i-f transformers and tuned circuits starting at the limiter tuned circuit and proceeding back to the grid circuit of the first i-f amplifier. Always be certain to apply signal generator through a small capacitor whenever it is applied to a circuit which has a d-c potential on it.

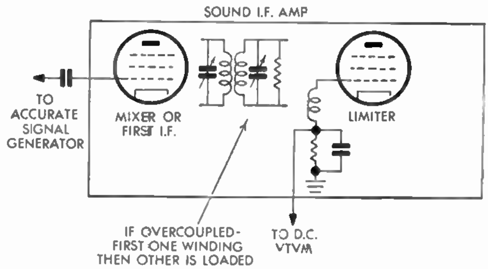


FIG. 308 Meter Alignment of Sound I-F Stages

3. If a single-tuned or sharply tuned i-f transformer is used, each resonant circuit is tuned for maximum on the meter. Be certain to keep output of signal generator at a low value to prevent overloading of the i-f stages. If overcoupled transformers are used in the sound i-f section, a special procedure must be used to align the overcoupled winding in order to obtain maximum gain at proper bandwidth. Customary procedure when aligning an overcoupled transformer using the meter method is to employ loading resistors which are placed across first the primary winding of the transformer under alignment and then the secondary. Indicating meter is still at the limiter grid resister, and signal is being applied to the grid of one of the sound i-f stages ahead of the transformer under alignment. At times, because the loading reduces the gain of the stage, it is necessary to use a substantial signal from the output of the signal generator. First, the loading resistor of a small value, as recommended by the manufacturer (200 to 2,000 ohms), is placed across the primary winding, and the secondary trimmer is tuned for maximum with the i-f center frequency being applied. Next, the loading resistor is transferred from primary to secondary winding, and the primary trimmer is tuned for maximum output at the i-f center frequency.

When using a sweep oscillator and oscilloscope to do the alignment, particularly in the case of the alignment of an overcoupled transformer system, the bandwidth and amplitude adjustment can be observed simultaneously on the response curve. General procedure using the sweep oscillator and oscilloscope method is as follows:

1. Apply sweep oscillator output to grid of the first sound i-f stage or in accordance with recommendations of manufacturer. Attach oscilloscope across grid resistor of limiter.

2. Adjust sweep oscillator for proper output level, center frequency, and narrow-band deviation as required for FM. Apply just enough signal to get a clean pattern on the oscilloscope screen when the vertical amplifier is adjusted for minimum attenuation.

3. Align the i-f stages in a progressive manner, starting at the limiter and proceeding to the point at which the sweep oscillator signal is applied. Align for maximum output, but at the same time keep the response curve essentially flat over the required bandwidth of the receiver. In the alignment of overcoupled stages, the effects of the tuning are immediately apparent on the curve, and adjustments should again be made for maximum amplitude, equal amplitude peaks, and not too deep a valley between the two overcoupled humps.

When it is necessary to align the sound i-f channel of the receiver which does not use a limiter, the output of the discriminator must be used as a point to which the meter can be attached. In this procedure, the discriminator must first be properly and accurately aligned. A vacuum-tube voltmeter is attached across the voltage stabilizing circuit at the output of a ratio detector or across one of the output resistors in a conventional discriminator. With the meter in the proper position, each of the i-f transformers is tuned for maximum indication.

158. *Discriminator Alignment*

In the alignment of a discriminator, a visual or vacuum-tube voltmeter procedure can again be used. Actual procedure is also dependent on what type of discriminator is used—conventional Seeley-Foster, ratio detector, or a modified unbalanced detector. The essential differences in the procedure for the various detectors are only in the placement of the output meter or oscilloscope. Correct positions for the various FM detectors are indicated in Fig. 309. Signal generator or sweep oscillator signal is applied to the grid of the i-f tube ahead of the discriminator or to the grid of an earlier stage. Always be certain, again, not to overload the i-f system when making the alignment. To be sure of this, always apply the minimum signal strength necessary to produce a discernible pattern on the screen when oscilloscope vertical amplifier is set for maximum gain or least attenuation.

Meter procedure for alignment of discriminator is as follows:

1. Apply signal generator to grid of limiter. Set frequency to precise center frequency. Apply just enough signal to obtain a discernible reading.

2. Attach vacuum-tube voltmeter across one of the resistors at the output of a conventional discriminator or across a part of the stabilizing network of a ratio detector (Fig. 309, arrows marked "PRI").

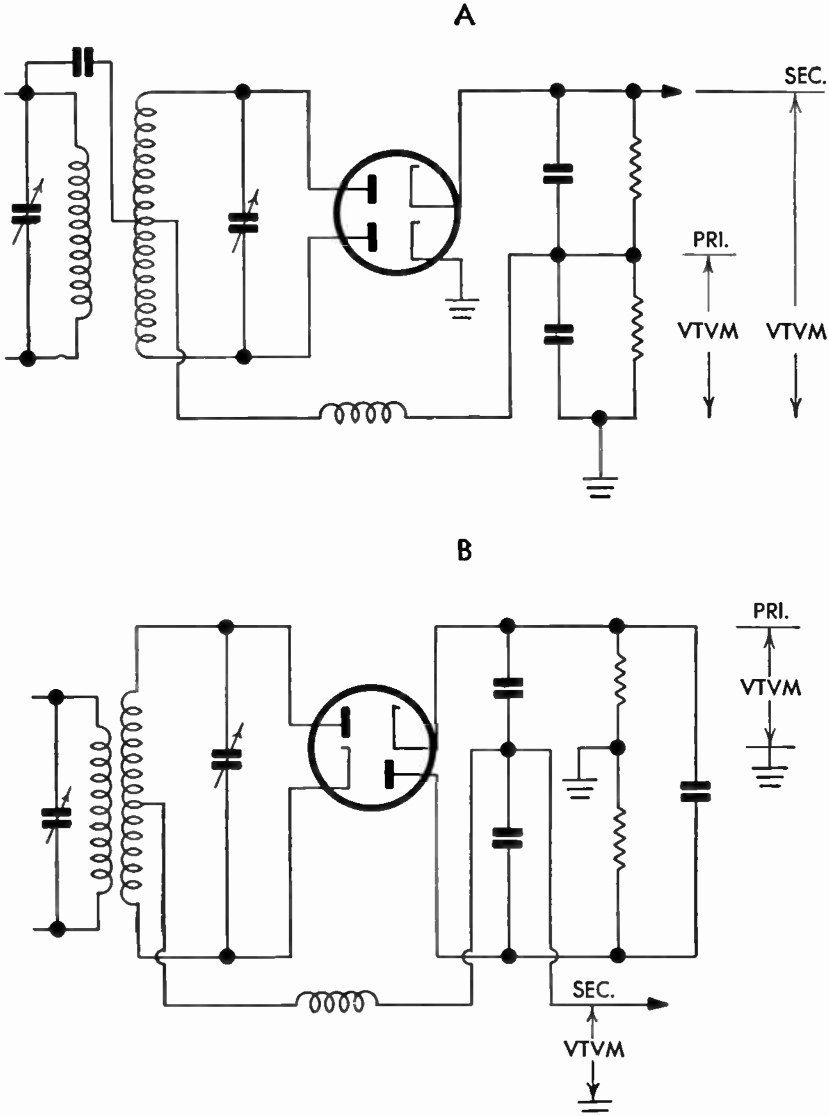


FIG. 309 Meter Alignment of Discriminator and Ratio Detector

3. Adjust the primary of the discriminator transformer for maximum meter reading.

4. Attach the meter across the output of either type of FM detector (arrows marked "SEC").

5. Adjust the secondary trimmer for zero or minimum reading on the meter. To make certain that the proper zero is obtained, vary the signal generator frequency on each side of the center frequency. On one side of the

center frequency, the meter reading should be positive and on the other side, negative. This check makes certain that the zero reading obtained is the one that occurs because the secondary is resonant at the center frequency and because the secondary is tuned off the correct frequency.

6. To make an approximate check of linearity, set the signal generator on a frequency exactly 25 kilocycles above the center frequency and take a reading; then apply a signal generator frequency 25 kilocycles below the center frequency. The readings should be identical but opposite in polarity if the linearity of the discriminator is satisfactory. When aligning a ratio detector which has inherent limiting qualities, it is preferable to align the input circuit with a weak applied signal. Therefore, the best limiting efficiency occurs on the reception of a weak signal; for stronger signals, for which the discriminator is not tuned precisely because of Miller effect, limiting action must not be as effective.

VISUAL ALIGNMENT OF DISCRIMINATOR

The following procedure is used to perform visual alignment of discriminator:

1. Attach sweep oscillator to stage preceding FM detector or to an earlier stage. Attach oscilloscope to output of detector.

2. Set sweep oscillator on exact center frequency with assistance of accurately calibrated marker signal. Keep output of sweep oscillator at the minimum value necessary to obtain a discernible pattern on the scope screen and with just enough deviation to sweep over bandwidth required.

3. When a sine-wave sweep is used on the oscilloscope, a single S-curve is produced on the screen indicating the response of the discriminator. If there is no center line on the scope to use as a reference point, it is often helpful to use a double S-curve (Fig. 310). Such a double S-curve is obtained by turning on the internal sawtooth sweep of the oscilloscope and using the sine-wave signal of the sweep oscillator to synchronize it. To obtain a true reproduction of the curve, it is necessary that the sawtooth sweep be double the sine-wave rate at which the output of the sweep oscillator is deviating. Thus, on the first sawtooth wave, the frequencies at the output of the sweep oscillator may be deviating from high to low in frequency; during the second sawtooth wave, the

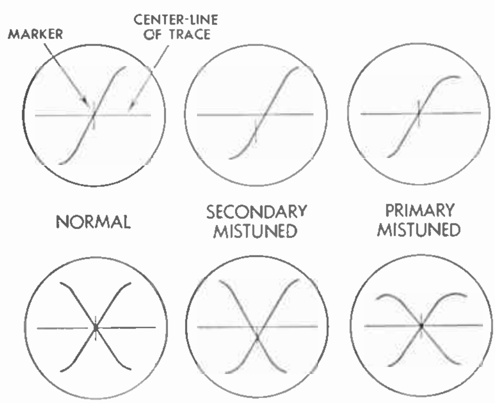


FIG. 310 Discriminator Curves

next half of the sine wave occurs and the frequencies are deviating from low to high. Inasmuch as the output amplitude is negative on one side of center frequency and positive on the other, the response curve is first traced from high to low and then again from low to high, low frequencies always producing negative voltage, let us say, and high frequencies, positive voltage.

4. Tune the primary of the discriminator transformer for a maximum vertical height, reducing the gain of the vertical amplifier on the oscilloscope at the same time to keep the pattern visible on the screen.

5. Adjust the secondary of the discriminator tuned transformer to put the center frequency marker at the center of the S-curve.

6. The correct shape and position of the marker on the S-curve is shown on the first drawing; in the second and third drawings of the same line, defects in tuning are shown. In the second drawing, the secondary is mistuned and the center frequency marker is not on the center line of the scope or on the center of the linear part of the response curve. This is understandable when we consider that, if the secondary is not properly tuned and the center frequency is applied through it, the output will not be nil but some positive or negative voltage depending on what side of resonant frequency the applied signal center frequency occurs. Readjust secondary tuning until wave form of S-curve, as shown in drawing A, occurs.

7. If primary is mistuned and secondary is in tune, it is true that the center frequency will occur on the center line of the scope indicating zero output; a signal applied to the secondary is the same as the frequency to which it is tuned. However, if the primary is not tuned properly, the amplitude of the instantaneous signal on each side of the applied signal center frequency will not be the same; therefore, more energy will be coupled into the secondary when the signal deviates on one side of the center frequency as compared to the times that it deviates on the other. Thus, as shown in the drawing, nonlinearity will be present and the response curve will not be symmetrical on each side of center frequency, producing a distorted output. Readjust primary until normal pattern is obtained.

8. A double S-curve is very helpful in locating the center line of the scope. If a single S-curve is used, it is necessary to center the pattern properly and to mark the center line with a crayon or calibrated scale (by turning down vertical gain). With a double S-curve defects are more evident in the second row of response curves. The first curve is normal, showing the center frequency at the crossover of the two S-curves and a symmetrical resolution on each side of center frequency. In the second drawing, the secondary is mistuned. and it is evident that the crossover of the two S-curves does not occur at the center of their symmetrical regions and that the response pattern is unbalanced. Defect is very evident even though the center line would not be present. In the last line, a nonsymmetrical pattern is obtained, indicating the primary mistuned.

9. The bandwidth of the discriminator can be checked very conveniently by moving the marker over the linear part of the S-curve and checking the frequency limits with a signal generator.

The advantages of the sweep oscillator and oscilloscope visual alignment of the discriminator are evident. The response curve on the oscilloscope screen indicates amplitude of discriminator output, bandwidth of the discriminator, and linearity or fidelity of the discriminator output. The response curve, of course, appears while adjustments are being made, and the effect of any adjustments is immediately apparent on the screen.

159. *R-F Unit Alignment*

If the sound i-f system is aligned first with an accurate signal generator method, it can be used very effectively to assist in the alignment of the front end of the television receiver. This is the usual procedure in most television receivers.

There are two specific procedures involved: one is the alignment of the mixer and r-f amplifier for best gain at required bandwidth; a second in the proper setting of the local oscillator frequency on all television channels. Sequence of alignment of the r-f unit and the alignment of other sections of the receiver should be that recommended by the manufacturer of the receiver. Generally speaking, the r-f and mixer units of the receiver are aligned first and then the local oscillator is set. One important exception is in the case of an r-f alignment which requires removal of a signal at the video detector; in this case, the i-f system must first be properly aligned and, also, the local oscillator properly set on the channel to be checked.

LOCAL OSCILLATOR ALIGNMENT

There are a number of ways in which to align the local oscillator properly on each individual channel or, as in the case of continuous tuning, to have local oscillator track over the range of frequency spectrum receivable. The local oscillator can be simply and effectively aligned after the sound i-f system and discriminator have been precisely set on correct center frequency. A number of procedures for setting the local oscillator are indicated in Fig. 311. One method, as shown in drawing A, is to apply an unmodulated signal at the sound-carrier frequency for the channels to be aligned to the antenna input. A meter is then attached at the output of the discriminator or ratio detector (in some receivers, a convenient jack is provided for this attachment), and the fine tuning is set on mid-scale. The oscillator trimmer or slug is now tuned for minimum reading of meter. When the meter reading falls to zero, of course, it is an indication that the local oscillator is beating with the incoming signal to produce an i-f frequency at the center frequency to which the sound i-f system and discriminator is tuned.

Similar results can be obtained by applying a frequency-modulated signal

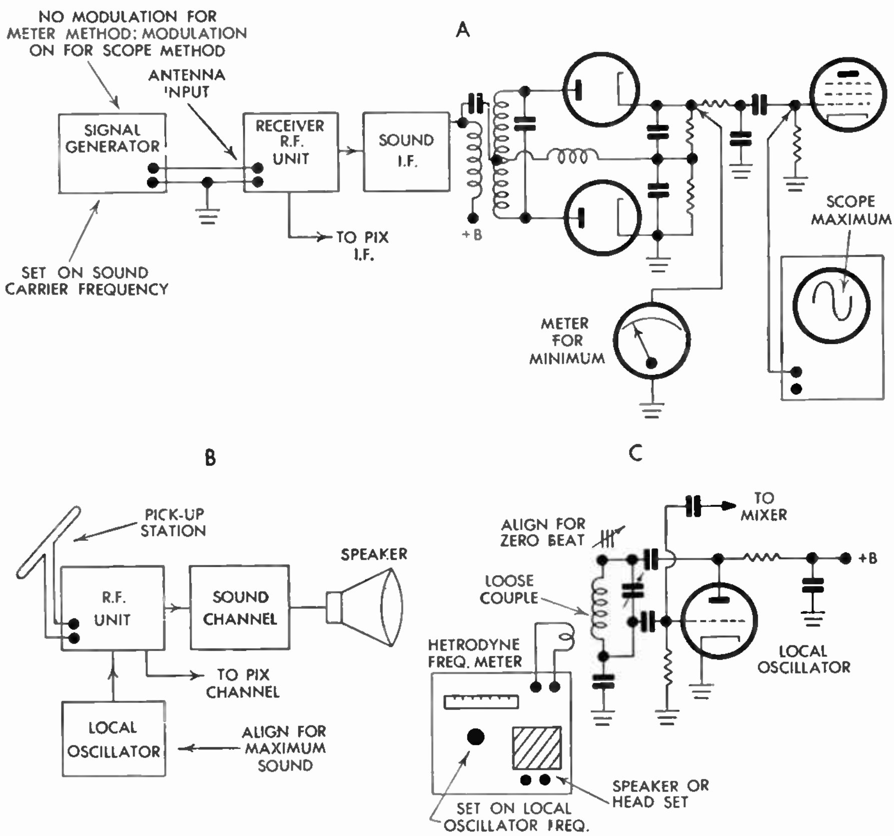


FIG. 311 Local Oscillator Alignment

to the antenna input and attaching an oscilloscope to the grid of the first sound audio amplifier. In this case, however, the local oscillator trimmer or slug is adjusted to maximum audio output or maximum vertical deflection of the oscilloscope.

One very precise method of adjusting the local oscillator when no accurate signal generator is available is to use the received sound signal from the various television stations in your area. After setting the fine-tuning control to mid-scale again, oscillator trimmer or slug is adjusted for best sound output. Another method of aligning the local oscillator without using the sound i-f section is with the assistance of a heterodyne frequency meter. The frequency meter is attached loosely to the local oscillator; a beat note can be heard when the frequency of the oscillator of the frequency meter is approximately that of the local oscillator. Although this method does not depend on the sound i-f section of the receiver to set the oscillator exactly, it is necessary that the heterodyne frequency meter be very accurate. In practice, the heterodyne frequency meter is set on the local oscillator frequency for the prescribed channel, then, the oscillator slug or trimmer is adjusted until a zero beat occurs at the output of the frequency meter. This indicates that meter and local oscillator are on the same frequency. It is very important to couple the frequency meter loosely to the local oscillator at a low-impedance point to prevent the presence of the meter from affecting the distributed capacity of the local oscillator stage.

In most front-end systems, a separate oscillator coil is used for each channel; therefore, tuning on each channel is a separate operation. However, in some receivers using continuous tuning or an incremental inductance method (as the RCA receiver), it is necessary to align the local oscillator in a certain specific sequence. For example, in the RCA receivers, local oscillator tuning starts at channel *13* and then proceeds in decreasing order, finally reaching channel *2*. This is necessary because in going from a high to a lower channel, the tuned circuit is simply extended with a new section, the old sections remaining a part of the tuned circuit for the now lower oscillator frequency.

ALIGNMENT OF R-F AMPLIFIER AND MIXER

A number of methods (Fig. 312) are again employed in the alignment of the r-f and mixer circuits. Manufacturers recommend a visual alignment pro-

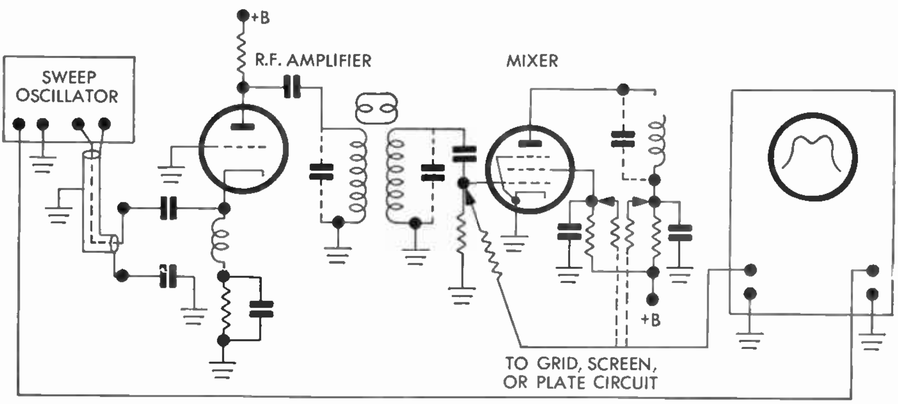


FIG. 312 R-F Amplifier and Mixer Alignment

cedure for the more or less critical adjustment of amplitudes and bandwidths of the r-f section. It is necessary to obtain the proper bandwidth to pass picture and sound and also to obtain a picture with high resolution. At the same time, it is important to obtain the utmost gain at proper bandwidth, of course, to keep the signal-to-noise ratio high. In the first method shown, the oscilloscope is attached to grid of mixer and a sweep signal is applied to the antenna input. With this method it is important that the sweep oscillator have a substantial output and the oscilloscope excellent sensitivity if a visual pattern is to be obtained on the screen after the sweep signal has been amplified by only a single r-f stage. The presentation on the oscilloscope screen and the signal applied to the vertical amplifier of the scope represent the variations in grid current as the signal rises and falls on the mixer grid, which, of course, of necessity for mixer action, is biased on the nonlinear portion of the transfer characteristic.

At times the oscilloscope is applied to the screen or plate circuit of the mixer, as recommended by the manufacturer, to obtain a visual pattern. This method takes advantage of some gain presented by the mixer. Method of take-off is satisfactory again because, in the mixer, the average component of plate or screen current varies with the amplitude variation of the applied signal. The amplitude variations, of course, are caused by the varying response of the r-f stage as the sweep frequency is swept over the channel spectrum.

A general alignment procedure for r-f and mixer alignment is as follows:

1. Apply sweep oscillator to antenna input circuit of receiver, using a balanced or unbalanced feed as required by receiver.

2. Attach oscilloscope to proper position in mixer circuit, again as recommended by manufacturer. Adjust sweep oscillator and oscilloscope until a pattern is obtained.

3. Inasmuch as in most r-f units some form of overcoupling is used (proximity of winding or a common coupling impedance), there are two necessary adjustments. One of these is tuning the circuit to correct resonant frequency; the second is adjusting the bandwidth of the overcoupled tuned circuit. Locate the respective elements in the r-f unit which control frequency and bandwidth.

4. Adjust resonant circuit tuning for peak amplitude pattern on the oscilloscope screen.

5. Adjust bandwidth control until the desired double-hump pattern is obtained, as recommended by manufacturer. Usually, if the circuits are brought into resonance, only a slight adjustment of bandwidth controls is necessary to obtain the proper spread. In some receivers, only a double-humped pattern is obtained on the low-frequency channels and the lower sections of the high-frequency channels. Because of the greater bandwidth of tuned circuits on higher frequency channels, only a single-hump pattern is often necessary to obtain the proper bandwidth.

6. In most receivers, a new set of tuned circuits is switched in for each channel. Therefore, each channel requires alignment. In some receivers the alignment must start at the highest frequency channel and progress to the lowest frequency one because of the addition of incremental inductances. In a receiver which employs a-f-c, r-f tuning and bandwidth adjustment must be made on only one channel. Oscillator alignment, of course, must be made on each channel.

160. *Alignment of Continuous-Tuning System*

In the alignment of an r-f unit which employs continuous tuning, there are three major adjustments—namely, tuning to resonance, bandwidth adjustments, and continuous tracking as the unit is tuned over the spectrum.

A typical example of alignment of a continuous-tuning system is the following procedure used to align the DuMont input tuner.

1. Attach oscilloscope to screen of mixer tube (Fig. 313). When attaching oscilloscope to an r-f point, be certain to use a shielded lead to prevent stray pickup.

2. Attach vacuum-tube voltmeter to output of discriminator for oscillator alignment.

3. Apply sweep signal generator to the antenna input of the receiver. For oscillator alignment, signal generator is attached at the same point.

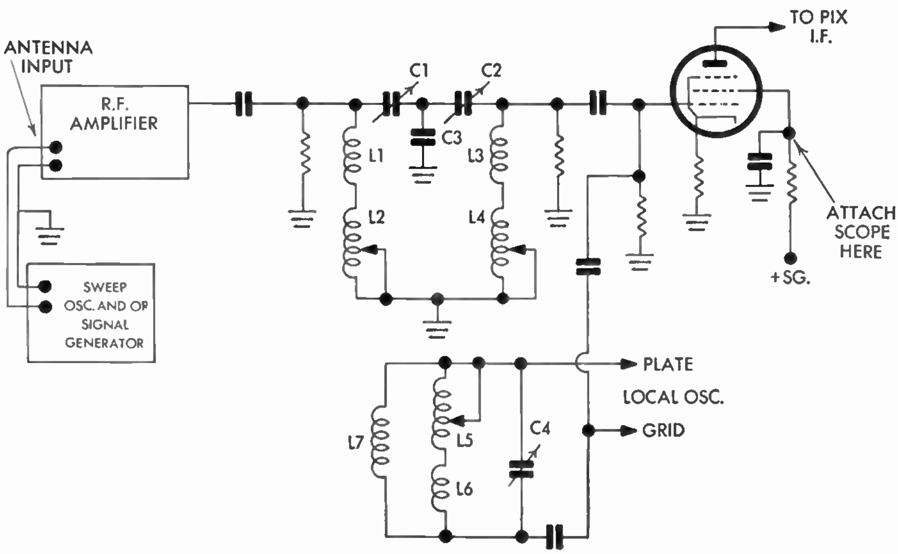


FIG. 313 Inductuner Alignment

4. Set tuning dial on frequency recommended by the manufacturer; in case of the DuMont tuner, on channel *4*. With the signal generator applied to the receiver input, rotate capacitor *C4* for minimum output on the voltmeter, again making certain that the true zero point is obtained.

5. Set dial to channel *13* and apply a signal generator frequency at the sound carrier of channel *13*. Adjust the end coil, *L6*, until minimum or zero reading is obtained on the voltmeter. This end coil is adjusted by squeezing or spreading the coil.

6. Check all channels to make certain that sound is received at correct dial setting. If assembly is shielded, be absolutely certain that the shield is on when adjustments are made.

7. Apply sweep oscillator to antenna input, and if marker system is not incorporated, loosely couple signal generator to act as marker.

8. With the sweep oscillator in operation, a double-humped pattern should be obtained on oscilloscope screen. This pattern should be first adjusted for

a 4.5-megacycle bandpass on channel *3*, and markers of the picture- and sound-carrier frequencies should appear on the peak of the double-humped characteristic. Low-frequency adjustments are made with capacitors *C1, C2,* and *C3;* capacitors *C1* and *C2* tune the resonant frequency of the primary and secondary tuned circuits of the bandpass coupling system, while capacitor *C3* has the most control over the bandwidth.

9. On the high-frequency channels, the end coils *L1* and *L3* control the bandwidth, which should be held between 4½ and 6 megacycles. Tuning is again affected at the higher frequencies by squeezing or spreading the end coils.

10. The bandpass should be observed on all channels. Just as in an AM receiver, adjustments made at either high end or low end affect slightly the adjustment at the opposite end of the spectrum, and it is often necessary to jockey controls slightly until optimum performance is obtained.

161. *Alignment of I-F Systems*

The major objectives in the alignment of the i-f systems are:

1. I-f system must be adjusted for proper amplitude gain to obtain the utmost sensitivity at the required bandwidth.

2. Bandwidth must be ample to amplify properly carrier and high-frequency sideband components of the television signal.

3. Various traps must be precisely adjusted to prevent serious interference or interference between the sound and picture signals in the receiver.

4. Picture carrier must be properly detuned to obtain proper phase and amplitude level to compensate for the partial sideband suppression at the transmitter.

PICTURE I-F TRAP ADJUSTMENT

The first task in the alignment of the i-f section is to adjust the various traps which trap out the associated sound from the picture i-f and also the adjacent channel sound and picture interference. These traps must be precisely tuned to correct frequency and, therefore, in their alignment, an accurate signal generator must be used. It is advisable to adjust the traps before going ahead with the alignment of the i-f transformers because their resonant characteristics affect the bandpass of the i-f system if they are tuned to a frequency within the desired spectrum to be passed by the picture i-f. Trap alignment procedure is as follows:

1. Connect a vacuum-tube voltmeter across the video detector (Fig. 314) load resistor and apply an unmodulated signal from the signal generator to the antenna input or, in some cases, to the mixer grid. An oscilloscope can also be used for trap adjustment if it is attached to the grid of the first video amplifier and a modulated signal from the signal generator applied to the receiver input.

2. Set the signal generator precisely on the trap frequency. Here, again, a simple crystal oscillator with an accurate crystal on the trap frequency would expedite precise alignment.

3. Align the adjacent channel traps in the picture i-f for minimum output (minimum reading on vacuum-tube voltmeter or minimum sine wave on oscilloscope screen).

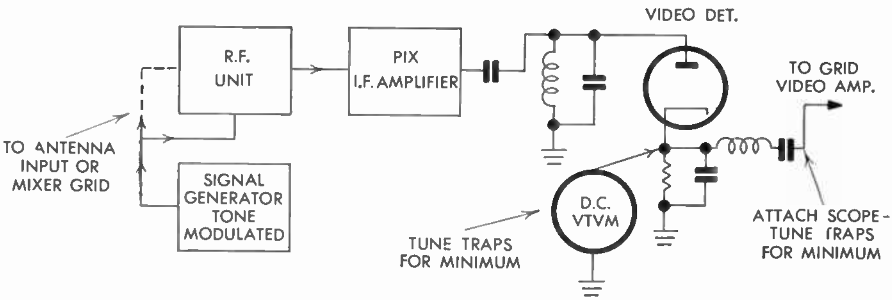


FIG. 314 Wavetrap Alignment

Occasionally, after the complete i-f alignment has been made, it is necessary to retouch the trap adjustment slightly. This can be done by observation of the visual response pattern on the screen of the oscilloscope, using the sweep oscillator set on the i-f range.

PICTURE I-F TUNED-CIRCUIT ALIGNMENT

A number of basic procedures can be used to align the picture i-f section, depending to a great extent on the type of i-f coupling system employed. When a simple stagger-tuned stage and a single-tuned transformer are used in the i-f system, a simple signal generator and output meter can be used to do the bulk of the alignment, and final touch-up can be done with a sweep oscillator and visual pattern. Although some work can be done with a signal generator and output meter in the alignment of an overcoupled system, a superior indication of the procedure and results obtained is evident when a visual method is used.

Generally the alignment proceeds from the stage ahead of the video detector back toward the first i-f stage. At times when oscilloscope used has a low sensitivity or the sweep oscillator a weak output, the sweep oscillator must be applied to an earlier stage in order to obtain sufficient gain to get a visible pattern on the scope. In this case, the resultant pattern on the screen is the composite response of two or more stages. This condition must be considered when aligning the last i-f stage.

Alignment of an i-f system of a television receiver should not be undertaken unless it is absolutely certain that the system is not properly aligned. As mentioned earlier, the response curve of the i-f system under check may be incorrect, but the chances are that the poor curve is the result of a defect and not a drift in alignment. If alignment is necessary, follow exactly the procedure recommended by the manufacturer. In lieu of this instruction, an i-f system can be aligned with a logical approach and a knowledge of the function of the component parts of the circuit to be aligned.

Alignment of a Stagger-Tuned System. A generalized procedure for alignment of a simple stagger-tuned system (Fig. 315) would be as follows:

1. Attach vacuum-tube voltmeter across video detector load resistor and apply unmodulated signal generator to the grid of the last i-f amplifier. An oscilloscope attached to the grid of the first video amplifier and a modulated signal generator could also be used.

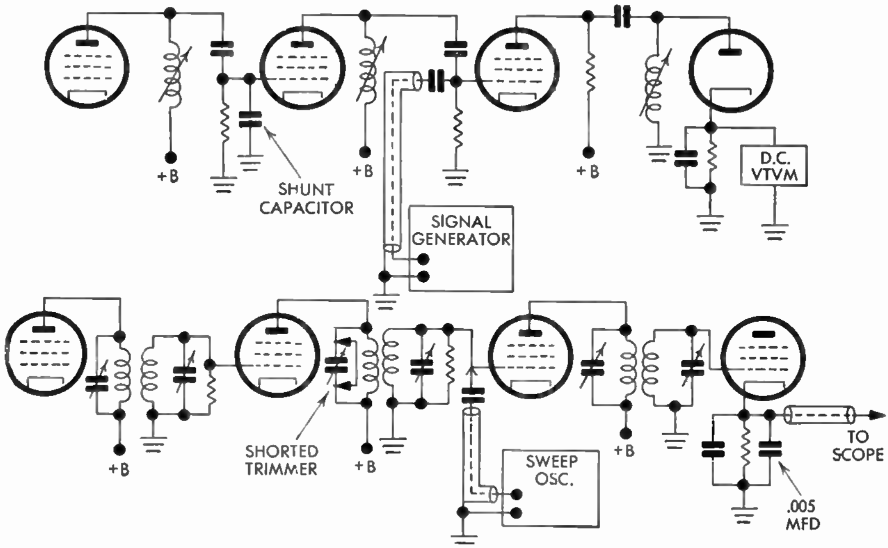


FIG. 315 Video I-F Alignment

2. To eliminate the effect of the preceding stages on the alignment of the last stage, it is preferable to shunt a capacitor of approximately 1,000 micromicrofarads from grid to ground of the earlier i-f stages. This will also reduce any oscillations that may be present in the i-f system which would cause erratic indications at the output.

3. Adjust a single-tuned transformer for a maximum output when the signal of frequency to which it must be tuned is applied to the grid of the preceding stage.

4. Move signal generator to the next earlier stage, removing shunt capacitor before signal is applied. Keep shunt capacitors attached to the earlier stages.

5. Adjust transformer between the last and the second-to-last i-f stages for maximum output when signal generator frequency is set on frequency to which it must be tuned.

6. Continue same process until all i-f transformers have been properly aligned.

7. Now attach sweep oscillator to antenna input or, if accessible, to the grid of the mixer. Attach oscilloscope across video detector load resistor. Adjust sweep oscillator and oscilloscope controls to obtain the response curve on the screen of the oscilloscope.

8. Observe carefully the response on the screen. See if it conforms with the desired response curve for a television i-f system.

9. Loosely couple an external marker generator (if not incorporated with the sweep oscillator) to check frequency distribution of the response curve, noticing the location of the picture i-f carrier and the extent of the bandpass. Also notice the positions of the trap frequencies and the so-called "overshoot" frequencies on the opposite side of the trap frequencies. Do not permit the marker to distort the shape of the curve.

10. If any portions of the response curve are low in amplitude, retouch the resonant circuits which dominantly control that part of the bandpass spectrum. If picture i-f carrier is not 50 to 60 per cent down, readjust those circuits which have frequencies near the picture i-f carrier frequencies. If any of the i-f traps are not precisely aligned as indicated by the marker, readjust those traps to their proper frequencies. If there is severe overshoot at frequencies outside the normal bandpass because of sharply resonant traps, adjust the resonant circuit which controls the overshoot until it is properly attenuated (Fig. 316).

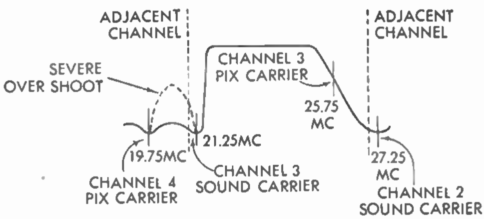


FIG. 316 Response Curve of Stagger-Tuned I-F

Under certain circumstances, it might be advantageous to permit alignment of an i-f stage singly, provided a sensitive scope is available along with a sweep oscillator with substantial output. Using this procedure, a crystal diode detector is first constructed (Fig. 317), which can be coupled into a plate or grid of an i-f tube to detect the signal present there. The 60- or 120-cycle detected signal, which represents the response characteristic, is then coupled through a filter to the input of the oscilloscope. Thus, any stage in the i-f system can be aligneo individually by applying the sweep oscillator to the grid of the tube preceding the transformer to be aligned and by applying the crystal probe and oscilloscope to the plate circuit of next tube. This system permits an individual check of any one stage in the i-f or, for that matter r-f section of the television receiver, provided, of course, that the sweep oscillator and oscilloscope meet the necessary signal strength requirements.

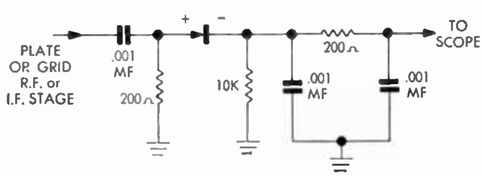


FIG. 317 Diode Probe Rectifier

Alignment of an Overcoupled System. Two basic approaches can be used in the alignment of an overcoupled system, which, although it produces more gain per tube than a simple stagger-tuned system, is somewhat more difficult to adjust and align. The visual method is the most common alignment procedure for this type of coupling (second drawing of Fig. 315). When a visual procedure is used, the sweep oscillator is first applied to the grid of the last i-f stage or the second to last, and oscilloscope is attached across the video detector load resistor, alignment starting at the stage immediately preceding the video detector. The sweep oscillator is then applied to successive stages, moving back toward the mixer.

General alignment and procedure for an overcoupled i-f system are as follows:

1. Attach oscilloscope to video detector load resistor, making certain that output circuit is shunted by large capacitor to remove spurious frequencies. Remove local oscillator and horizontal and vertical sweep generator tubes.

2. Apply sweep oscillator to grid of last video i-f tube. Remember to apply signal through a small isolating capacitor to prevent shunting of any applied external bias.

3. Adjust sweep oscillator and oscilloscope controls for a clean, discernible pattern. It is important to understand at this point that sweep oscillator output must be substantial and that the oscilloscope must be sensitive if response of a single stage is to be obtained.

4. Align interstage coupling system between last i-f stage and video detector for response curve recommended by manufacturer. Keep in mind that the amplitude of the curve is dependent on correct setting of the primary and secondary tuned circuits and that the bandwidth is under control of a mutual coupling element if such element is adjustable.

5. To prevent difficulty from oscillation when performing the alignment procedure, it is advisable to shunt the earlier grids of the i-f system with approximately 1,000-micromicrofarad mica capacitors. It is also advisable to shunt the primary tuned circuit immediately preceding the grid at which the sweep oscillator is attached to prevent it from acting as an absorption tuned circuit and removing signal at the frequency to which it is tuned. This expedient prevents a dip in the response curve of the stages under alignment due to reduction in amplitude of the sweep signal at this frequency because of absorption.

6. Response curve is calibrated, of course, with a marker signal to permit correct alignment. Always remember to supply just sufficient marker amplitude to provide a niche in the curve without distorting the true pattern.

7. Apply sweep oscillator to grid of preceding stage, making certain to remove shunting capacitor on grid and across primary tuned circuit.

8. Adjust coupling system between second to last and last i-f stage to produce desired response pattern recommended by manufacturer. The objective of the alignment procedure, again, is to obtain maximum amplitude of

the response curve at the proper bandwidth. When an overcoupled system is used, it is also important that the valley between the two humps be no less than 70 per cent of the peak to prevent low-frequency losses. The response curve shown on the oscilloscope, of course, is now a composite curve of the second to last and last i-f stages. We must realize that the individual curve of a single stage of the i-f system or of a group of stages is not ideal or even similar because each stage is designed with a certain characteristic and it is the over-all response of all i-f stages which produces the desired response. In many overcoupled systems, the tuning is slightly staggered to permit narrower bandwidth per stage, and therefore greater gain, while, at the same time, the over-all response is of the necessary broadness.

9. Move the sweep oscillator back one more grid and proceed as before. As the sweep oscillator is gradually moved toward the mixer, output amplitude must be correspondingly reduced to prevent overloading of the i-f system and to obtain a representative pattern at a rather weak signal level where the i-f system performance should be at its best. Follow the same procedure until all i-f stages are properly aligned and correct over-all response is obtained. At the conclusion it may be necessary to readjust the various traps to prevent overshoot or improper suppression of spurious signals.

In some i-f systems a special tuned circuit is incorporated to detune the picture-carrier frequency to the 50 or 60 per cent level. In other receivers each stage is individually detuned a slight amount at the picture-carrier frequency; therefore, the over-all picture carrier will be detuned the proper amount. In still other receivers, which employ stagger tuning, the individual frequencies to which the various tuned circuits resonate are chosen in a manner and with the correct stage gain to have the picture carrier detuned the correct amount.

Emergency Alignment. When the recommended equipment is not available, an overcoupled system can be approximately aligned with a vacuum-tube voltmeter and conventional signal generator by using the following procedure:

1. Signal generator is applied to grid of the last i-f stage, frequency of which is set at approximately the center of the i-f bandpass if specific frequency for which the last i-f transformer is to be tuned is unknown. If center of the bandpass of the last i-f stage is known, the signal generator is set on this frequency.

2. Shunt primary tuned circuits of last i-f transformer with a low-value resistor (approximately 500 ohms). Tune secondary for maximum reading on vacuum-tube voltmeter.

3. Now, remove shunt and attach it across secondary of last i-f transformer, tuning primary for maximum meter reading.

4. The bandpass of this stage can now be checked by removing the shunt and locating the frequency of the two peaks of the usual double-humped characteristic. Check relative amplitude of two humps and also their frequency spacing, judging performance by comparison with recommended response

curve of the manufacturer. If bandwidth is too great or too little and a bandwidth adjustment is incorporated in the coupling circuit, adjust it to obtain the proper separation between the two peaks.

5. Move the signal generator back to the next stage and use this same procedure to align the second to last i-f transformer. Again, make certain that the i-f system is never overloaded.

6. Follow the same procedure in the alignment of the remaining i-f stages—namely, shunt one tuned circuit of the overcoupled transformer and then the other to prevent overcoupling, tuning each tuned circuit individually to correct center frequency. When the shunt is removed, the overcoupled transformer will have the approximate overcoupled response desired.

7. After the complete alignment is performed, check various frequency points on the over-all response curve to determine if response is approximately the desired one for television i-f system. Check particularly the picture-carrier frequency, which must be 50 to 60 per cent down, and make necessary detuning adjustments if it is not.

162. *Alignment and Check of Special Circuits*

In the television receiver there are a number of special circuits which can be checked for performance very simply with conventional test equipment. There are also a number of checks which the capable technician can make on various parts of the receiver. These checks permit him to judge the performance and to locate defects in special circuits of the receiver. Many times it is not a defect but just a matter of improper alignment, and a slight readjustment of a tunable component will bring the receiver back to normalcy.

AUTOMATIC SYNC CONTROL

When an automatic sync system is used for horizontal sweep, it must be properly adjusted for best stability and least susceptibility to noise. In most receivers after a station is once tuned in, a sync-control system holds a picture locked in although horizontal hold control is varied over its entire range. Likewise, with horizontal hold control in most any position except near very extremities of its range, a sync-control system should be able to lock in the picture when a station is tuned in.

The two main adjustments in the horizontal sync-control systems are the frequency of the oscillator and the phase of the generated sine wave applied to the discriminator. The frequency of the horizontal oscillator must be set at a free-running point for which it will hold synchronism although the hold control itself is varied over its entire range. It is also important that the phase of the sine wave at discriminator is such that the generated sawtooth retrace period occurs at the same time a horizontal blanking pulse strikes the grid of the picture tube. If this is not the case, the retrace period of the sawtooth does not occur during the received horizontal blanking; therefore, the blanking occurs at some part of the trace of the sawtooth, producing a black vertical blanking bar on the screen.

A typical tune-up procedure for a horizontal sync-control system is as follows:

1. Tune in station and adjust fine tuning of receiver for best sound quality. Set contrast of picture slightly below normal.

2. Set frequency control of oscillator tuned circuit (primary of horizontal discriminator transformer) until a setting is found at which picture remains in synchronism over the entire range of the horizontal hold control.

3. Adjust phase control of discriminator tuned circuit (secondary of horizontal discriminator transformer) to a point where no blanking bar appears on the screen at either the extreme clockwise or counterclockwise setting of the horizontal hold control. This control actually determines the ratio of resistance and reactance reflected into the primary circuit of the oscillator and, therefore, phase of sine wave.

AUTOMATIC GAIN CONTROL (A-G-C)

One advantage of an a-g-c system is that it can be set to prevent overloading of the video section or loss of horizontal synchronism due to too weak or too strong a signal. In a locality in which signal strengths are not too divergent, receiver readjustments are not necessary when throwing the channel selector switch from one station to another. If the a-g-c system is specifically set for your locality, it means a simple tuning procedure and reduction of defects introduced by too strong or too weak received signals.

The a-g-c system is adjusted on the weakest signal. Thus, if in a certain locality the weakest signal is approximately 500 microvolts, the a-g-c system would be set to deliver the proper peak-to-peak signal to the grid of the picture tube when a 500-microvolt signal is applied to the antenna input circuit. In a stronger signal area, of course, the a-g-c system would have to be biased to a higher level to permit delivery of the same amount of signal to the grid of the picture tube when a correspondingly stronger peak signal is received. In some receivers there is no separate a-g-c adjustment, and oftentimes the contrast control is a part of the a-g-c system; therefore, its setting determines the action of the a-g-c on signals of widely divergent signal strength.

VIDEO AMPLIFIER GAIN AND RESPONSE

The gain of the video amplifier can be checked very readily with a simple audio signal generator and an a-c vacuum-tube voltmeter. Procedure is as follows:

1. Attach signal generator to grid of first video amplifier, being certain to insert a series isolating resistor if the signal generator has a low impedance output (Fig. 318). Apply a measurable low-amplitude signal, in the frequency range between 1,000 and 10,000 cycles, to the grid. Signal, of course, should be kept at a low amplitude (generally a 1-volt signal is satisfactory) to pre-

vent overloading of the video amplifier. If uncertain whether overload exists, simply attach oscilloscope to the plate circuit of the video output tube and check the fidelity of the sine wave.

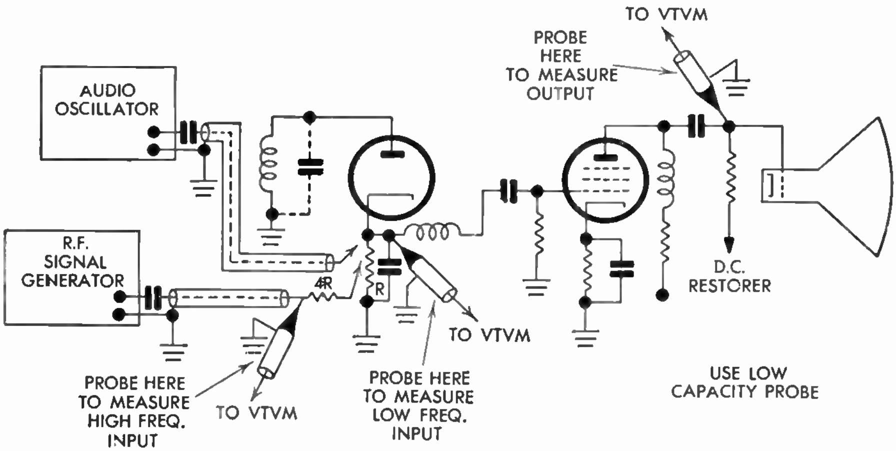


FIG. 318 Video Response Measurements

2. With the a-c vacuum-tube voltmeter, measure the amplitude of the signal applied to the grid of the first video amplifier tube. Then attach the a-c vacuum-tube voltmeter to the plate of the video output tube or to the grid of the picture tube, whichever is more convenient, and measure signal again.

3. The gain of the video amplifier, of course, is the quotient of output over input.

It is a bit more difficult to measure the frequency response of the video amplifier because of the number of precautions that must be taken in performing the check. An approximate check of video response can be obtained by simply applying an r-f signal generator to the grid of the first video amplifier tube and by observing the pattern on the fluorescent screen with no received signal, ascertaining to what frequency point a definite and discernible vertical bar pattern is apparent on screen. If a discernible pattern can be seen up to 2 or 3 megacycles, the video response is generally good.

A more reliable check of the video response can be obtained by making spot checks at various frequency points in the spectrum, approximately 30 cycles up to 3 or 4 megacycles. However, it is very important that the technician understand that the accuracy of such a response measurement depends on his understanding the instruments he is using to make the check, realizing their capabilities, and duly compensating for their loading effects. A video response can be made with a simple audio oscillator to cover the low-frequency ranges up to approximately 15,000 or 20,000 cycles and with an r-f signal generator which will pick up again at approximately 100 kilocycles and continue up to 3 or 4 megacycles. Range between 10 and 100 kilocycles is unimportant because in most video amplifiers this is an absolutely linear section.

The most logical method by which to make the check with the general run of test equipment (Fig. 318) is to attempt to plot gain against frequency. That is, the signal applied to the input of the video amplifier is measured at a specific frequency, and then output is measured with same meter. The gain, of course, is the quotient of the output over input readings. This method of attack eliminates the problem of keeping an absolutely constant applied input voltage or output voltage, and so long as the amplifier is not overdriven, a true response curve can be obtained by comparing the gain at the various frequencies. The process of using the same a-c vacuum-tube voltmeter, jumping the probe from input to output, eliminates balance discrepancies which the meter might otherwise introduce because of poor frequency response.

Another important factor to consider is the loading effect of the signal generator on the amplifier. If it has a low-impedance output, it must be properly isolated from the impedance of the circuit to which the instrument is attached. A series isolating resistor will do this job satisfactorily over the range of frequencies to be measured. So far as the audio signal generator is concerned, this resistor can be of any reasonable value, say 100,000 ohms, because the output of the audio signal generator is ample, and the a-c vacuum-tube voltmeter can be used to measure the voltage directly at the grid of the first video amplifier.

In the use of the r-f signal generator (which generates the higher frequencies we are mainly interested in), another problem arises because of the inherently low voltage output of an r-f signal generator. Thus, it is sometimes difficult to get a readable signal on the grid of the first video amplifier and at the same time have the signal generator properly isolated through the series resistor. This problem can be overcome, as shown in the figure, by accurately measuring the series isolating resistance and resistance from grid to ground of the first video amplifier. (Resistance from grid to ground cannot be readily read on a conventional ohmmeter unless the video detector is direct-coupled to the grid of the first video amplifier.) In this case, a-c resistance from grid to ground is approximately the value of the diode load resistor. If the circuit is capacitively coupled, the resistive component of the load can again be very closely approximated by reading the resistance of the diode resistor although an ohmmeter attached from grid to ground of the first video amplifier will not read this low value because of d-c isolation introduced by coupling capacitor. Of course, it must be realized that at video frequencies, coupling capacitor has an insignificant reactance and the load placed on the circuit is resistive, approximately equal to the value of the diode load resistor. The ratio of the isolating resistor and the resistive component of the grid load can be used to judge the amplitude of the grid voltage applied to video amplifier. Vacuum-tube voltmeter must be attached directly to output of the r-f signal generator in order to obtain a readable voltage. Thus, if the a-c vacuum-tube voltmeter reads 1 volt across the output of the r-f signal generator when attached directly across its output and ratio of isolating resistor to resistive load is 10:1, it means that only approximately 1/10 volt is being applied to the grid of the video amplifier. (This value would be so low the ordinary a-c vacuum-tube voltmeter could not give an indication.)

A general procedure for measuring the frequency response characteristic of a video amplifier is as follows:

1. Attach audio oscillator through an isolating resistor or directly to grid (if output impedance is high) of first video amplifier. It is also necessary to insert an isolating capacitor if any external bias is applied to grid of this amplifier.

2. Set the signal generator on approximately 1,000 cycles, and apply about 1-volt signal to the grid as measured on an a-c vacuum-tube voltmeter attached to the grid. It might be well at this point to apply an oscilloscope to the plate circuit of the video output tube to make certain that the video amplifier has no been overloaded by application of signal. If it has not been overloaded, be certain not to apply voltage much in excess of this value to make absolutely certain that amplifier tubes will not be driven into the nonlinear portion of their characteristics.

3. Disconnect oscilloscope and touch the a-c vacuum-tube voltmeter probe to the output circuit, measuring the voltage at this point. Divide output by input to determine the gain of the amplifier at 1,000 cycles. When using an a-c vacuum-tube voltmeter to measure a-c r-f frequencies, it is very important that the ground at the probe be used. If at all possible, do not attach any additional lead to the end of the probe, but try to reach the take-off point with the probe directly.

4. In like manner, measure the gain of the video amplifier at various other low-frequency points of sufficient number to get a general plan of the response characteristic. It is worth while observing that, if we use the gain method of plotting frequency response, the actual value of the applied signal voltage need not be held constant but can vary over wide limits so long as too much signal is not applied, thereby preventing overload of the amplifier.

5. After the low-frequency measurements have been made, attach the r-f signal generator through an isolating resistor and capacitor (high-value mica capacitor) to grid of the video amplifier tube. Measure the resistance of the isolating resistor and determine the resistive component from grid to ground of the video amplifier to permit calculation of applied signal.

6. Apply grid signal voltage.

7. To measure the input voltage, the a-c vacuum-tube voltmeter must be connected directly across output of the signal generator. True grid voltage signal is calculated by multiplying meter reading by the ratio of the grid resistance over the sum of grid resistance and isolating resistance.

8. Apply a 100-kilocycle signal to the grid of the first video amplifier tube, and calculate its amplitude by first measuring the total output of the signal generator with the a-c vacuum-tube voltmeter.

9. Attach the a-c vacuum-tube voltmeter to output of the video amplifier, and measure output voltage. Gain at 100 kilocycles is the quotient of the output voltage divided by the calculated grid voltage signal.

10. Follow the same procedure in measuring and calculating the gain of the video amplifier at sufficient high-frequency points to obtain a satisfactory plot of the high-frequency response.

11. With the a-c vacuum-tube voltmeter attached at the output, the r-f signal generator should be varied slowly over its frequency range of 100 kilocycles up to 4 megacycles. At any point where a decided dip or rise in the meter reading is apparent, the gain at that frequency should be checked in particular. Understand that the rise or dip may be caused by a characteristic of the r-f signal generator and not by the video amplifier. For that reason a measure of gain is necessary at this frequency point before any conclusions are drawn. If there is a decided drop or rise in gain at this particular frequency, there is a resonant defect somewhere in the video amplifier under test.

VIDEO RESPONSE CURVE ON OSCILLOSCOPE

A sweep oscillator and oscilloscope combination can also be used to check the frequency response characteristics of a video amplifier (Fig. 319). In this method, the sweep oscillator output must have a zero- to 10-megacycle

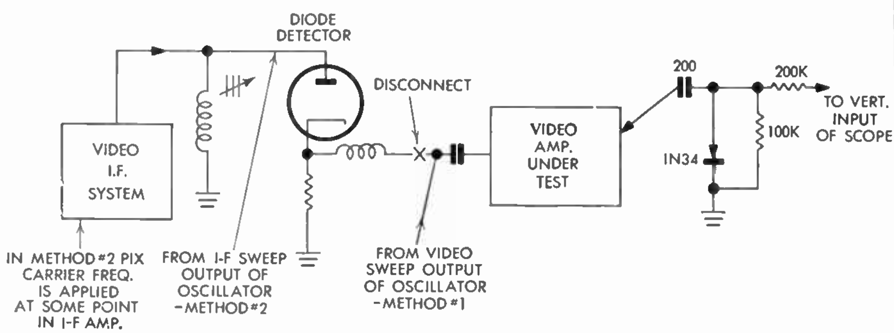


FIG. 319 Video Amplifier Response Using a Video Sweep

sweep, which is applied to the input of the video amplifier. Generally, the coupling capacitor between the picture second detector load resistor and the first video grid is disconnected, breaking the connection at the detector load. Sweep oscillator is attached to the free end of the coupling capacitor.

The vertical input of the oscilloscope is attached to the video amplifier output through the crystal rectifier circuit indicated. Thus, as the oscillator sweeps over the zero- to 10-megacycle range at a 60- or 120-cycle rate, the individual frequencies are rectified by the crystal and resultant waveform represents the response of the video amplifier. Again, the response curve can be obtained on the oscilloscope screen although the oscilloscope amplifier itself has only a limited response because it is only necessary that the vertical amplifier of the scope be capable of amplifying the detected 60- or 120-cycle component. Without the application of the crystal rectifier, it would be essential to have an oscilloscope with a linear response to 10 megacycles to obtain a response curve.

Another procedure for obtaining the curve of the video amplifier is to apply a conventional sweep oscillator signal (sweep oscillator without a video frequency sweep) to the video detector or last video i-f amplifier. At an earlier point in the i-f system, apply a single-frequency set on the picture i-f carrier frequency. The beat note, set up between the single-frequency picture-carrier frequency and the sweep frequency, the center frequency of which is also at the picture-carrier i-f frequency, produces a detected component of modulation at the output of the video detector that varies from zero to the extent of the sweep frequency range. It might extend outward 4 megacycles from the picture-carrier center frequency setting of the sweep oscillator. For example, if the single-frequency signal applied is set on 25¾ megacycles and the sweep frequency range is between 21¾ and 29¾ megacycles with 25¾, of course, the center frequency point, the swept frequencies from zero to 4 megacycles appear at the output of the video detector. This zero- to 4-megacycle sweep is occurring at the same rate as the frequency sweep of the sweep oscillator. Inasmuch as the frequencies are being applied at an essentially constant amplitude to the grid of the first video amplifier, each frequency will be amplified in the video amplifier in accordance with the gain of the amplifier at that particular frequency, producing a resultant wave form at the output of the last video amplifier which corresponds to the response of the video system. A crystal rectifier and load is again used to detect the low-frequency components which represent the response curve.

CHECK OF TRANSMISSION-LINE SURGE IMPEDANCE

The surge impedance of a transmission line can be conveniently ascertained with a sweep oscillator, oscilloscope, and a few known resistors. In

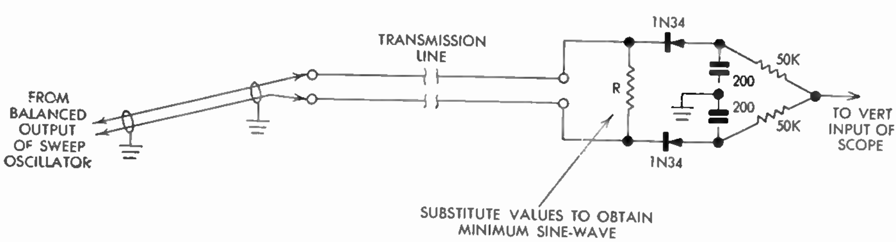


FIG. 320 Check of Transmission Line Surge Impedance

this check (Fig. 320), loosely couple sending end of a few quarter wavelengths of the transmission line under check (or entire length of line can be used if an extra section is not available to make the check) to sweep oscilator. The receiving end of the line is terminated with the resistor standard. Across these terminals is attached the diode crystal rectifier. Output of the detector is applied to vertical input of the oscilloscope. In operation, various values of known resistance are substituted at the termination until the very lowest amplitude sine wave appears on the oscilloscope screen. When the very minimum is obtained, the transmission line is properly terminated, and the value of the standard resistor represents the approximate surge impedance of the transmission line.

A sine wave appears at the output of the oscillator because of the changing characteristics of the section of transmission line to the various sweep frequencies applied. For example, because of its lengths, some frequencies applied during the frequency sweep (frequency sweep set near the frequency range which the transmission line is to convey) cause the transmission line to act as a parallel resonant circuit while to others of the swept frequencies, the section of line acts as a series resonant circuit. Frequencies for which the cable length is an odd or even number of quarter wavelengths occur during the frequency swing of the sweep oscillator. If it is an odd number of quarter wavelengths, a high impedance or parallel resonant condition is obtained, and a voltage maximum is found at the sending end of a line. If the line is an even number of electrical quarter wavelengths, end of line becomes a current maximum. Thus, the voltage across a terminating resistor rises and falls as impedance of the sending end of the line varies with the frequency applied. This variation is rectified and applied to the oscilloscope screen.

When the transmission line is properly terminated, however, there will be no standing wave on the line, and an essentially constant voltage will be present across the terminating resistor for the various swept frequencies applied. For, if a transmission line is properly terminated in its characteristic impedance, the transmission line presents a resistive load at its sending end to the various frequencies applied. It is apparent, therefore, that the sine-wave component will be at its minimum when the line is terminated by a resistor, the value of which is equivalent to the surge impedance of the line, developing an output component which is constant over the swept range.

LINEARITY CHECK

At the same time that the frequency response check is being made, it is also convenient to make a simultaneous check of linearity. This can be done simply by observing the bar patterns which appear on the screen of the picture tube. For example, when the audio signal generator is used to measure the low-frequency response, a series of horizontal bars appeared on the screen. At a very high frequency (10,000 cycles and approaching the line rate of 15,750), the horizontal lines are relatively close spaced. Linearity can be checked by comparing the separation between the lines at the top of the picture with separation at the center and lower section of the picture.

Inasmuch as the minima and maxima of the audio sine waves applied to video amplifier are of constant amplitude and also of constant time intervals, the bars which develop on screen of the picture tube when spaced linearly indicate that the vertical sawtooth sweep is linear as the beam scans from top to bottom of the screen. It is apparent that the crest of the sine waves arrives at the grid of the picture tube at a constant repetition rate; therefore, if the beam moves down the screen at a constant linear rate, the black bars (negative peak of sine wave on grid of picture tube) should appear with even spacing down the screen.

When the r-f signal generator is applied to the video amplifier to check the video response, a series of vertical bars will appear on the screen which can be used to check the horizontal linearity of the television system. (Rate of arrival of sine wave is now faster than the time required for the beam to trace from left to right across the screen; therefore, a series of black and white sections appear along each line and blend together from line to line to form vertical bars.) Again, the rate of arrival of the sine waves at the grid of the picture tube is a constant, and if the horizontal motion of the scanning beam is absolutely linear, the width and separation of the bars (caused by the negative sweep of the applied sine wave) should be uniform across the screen, indicating true horizontal linearity. If there is any nonlinearity, the vertical bars will appear crowded at some section of the screen in comparison to their spacing and width at other portions of the screen.

CROSSHATCH GENERATOR

A crosshatch generator generates a signal which, when applied to the video amplifier and onto the grid of the picture tube, forms a series of horizontal and vertical bars to set off a squared crosshatch pattern on the screen. This pattern is used to judge the linearity of the sweep system. Horizontal bars check the vertical linearity; vertical bars check the horizontal linearity. The crosshatch generator generates a repeating signal, which is higher in frequency than the line rate of the television system, and another signal, which is much lower than the line rate.

With the high-frequency signal, therefore, there are a number of variations appearing as an active line of the picture is being traced by the scanning beam. Consequently, along that line there will be a number of excursions into the black region. The low-frequency repeating signal is low in comparison to the rate at which the lines are scanned; consequently, the black region of the lower frequency signal will take up a few lines, and, likewise, the period between the black regions will occupy a number of lines. Consequently, as the scanning beam traces down the screen, it will meet groups of black signal which take up a few lines thereby tracing a series of horizontal lines on the screen.

A typical crosshatch generator, as manufactured by Philco for checking receiver linearity, is shown in block diagram form in Fig. 321. This generator consists of two oscillators—one operating at approximately 300 kilocycles and the second around 400 cycles, generating a series of sine waves which are first clipped and then applied to the video amplifier of the receiver under test. These sine-wave oscillators are synchronized by sync pulses derived from horizontal and vertical sweep systems of the television receiver, thereby producing stationary crosshatch lines on the television screen. The crosshatch signal is generally used when a signal is being received, and it superimposes the crosshatch pattern on top of the received signal. Received signal, therefore, contributes the blanking pulses, which black out the screen during the retrace period; therefore, during retrace periods, the crosshatch signal is also ineffective.

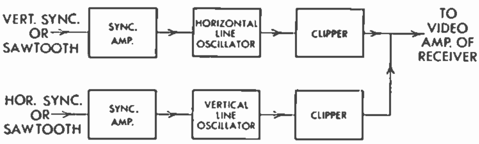


FIG. 321 Block Diagram of Crosshatch Generator

The 300-kilocycle signal from the crosshatch generator is approximately 20 times higher in frequency than the line repetition rate of the television system. Therefore, during the active trace of the horizontal sweep, approximately 15 to 16 variations occur as the beam sweeps left to right across the screen, producing 15 to 16 vertical lines. The 400-cycle components of signal space out approximately 12 horizontal lines as the scanning beam moves from top to bottom during the active trace. Thus, with the number of active lines equal to approximately 500, the 400-cycle signal blacks out a group of horizontal scanning lines every 38 active lines. The linearity of the television system is checked by observing the spacing between the two groups of lines. Horizontal linearity is satisfactory when the spacing between the vertical lines of crosshatch signal are uniform from left to right across the screen; vertical linearity is satisfactory when spacing between the horizontal crosshatch lines is uniform from top to bottom of the screen. Any crowding or spreading of spacing in comparison to other sections of the screen is an indication of poor linearity, and linearity control should be adjusted to obtain equal spacing throughout the entire scanning raster.

163. *Special Television Signal Generators*

In laboratory production plants or the more elaborate service centers, special television test signal generators are used to generate a composite television signal which can be piped throughout the area to supply a signal which can be used to check the performance of the receivers without having to rely on received signal. Two special tubes developed by RCA are at present applicable to this type of service. One is called a *flying spot scanner* and the other a *monoscope tube.* Each tube in conjunction with a satisfactory synchronizing system is used to generate a synthetic television signal which can be piped throughout the area on a coaxial line or used to modulate a low-powered high-frequency signal generator (Figs. 322 and 323). Flying spot cathode-ray tube 5WP15 is a modified picture tube producing a very brilliant scanning raster with an extremely short persistency. The picture-tube beam is electrostatically focused and is magnetically deflected in accordance with a regular scanning pattern. The elements of the screen fluoresce brightly under impact of the scanning beam and lose their illumination rapidly after the beam moves on.

A flying spot video signal generator, as shown in Fig. 322, consists of a horizontal and vertical sawtooth generator with associated magnetic sweep amplifiers to generate a fine spot scanning raster. To obtain sufficient bril-

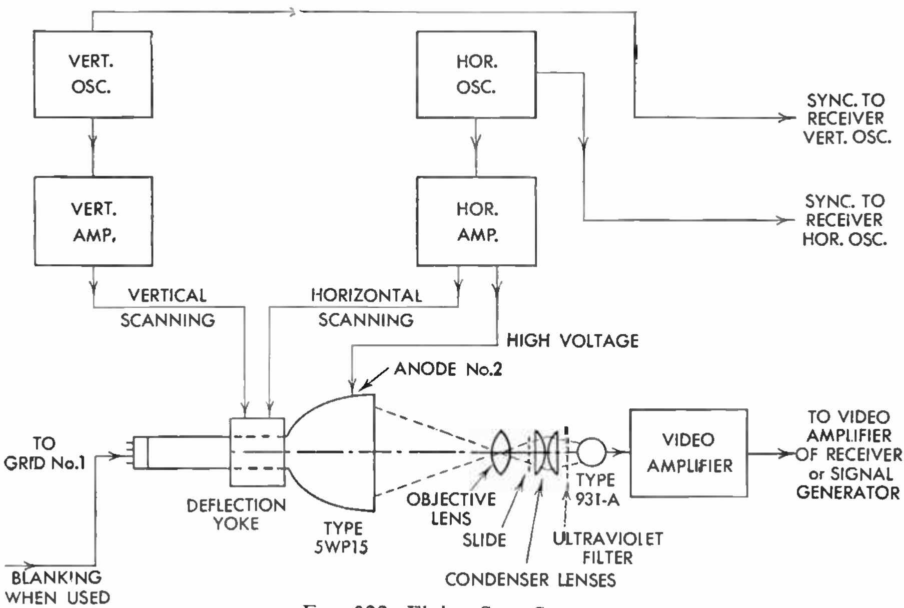


FIG. 322 Flying Spot Scanner

liancy, the 5-inch flying spot tube is operated with a 20,000-volt anode potential. A so-called "objective lens" focuses the moving light source on a slide transparency, thereby scanning the transparency in accordance with a standard pattern. The light emanating from the opposite side of the transparency varies in accordance with the picture information on the slide—no light being transmitted through the slide for a black section, and a maximum amount of light transfer for a section representing intense illumination. It must be understood that the light emanating from the back of the slide, of course, is being released in a sequential manner, element by element, line by line, as the front side is scanned progressively by the light.

This light leaving the exit side of the slide is progressively converged on a phototube by a condensing lens. Actual variation of light intensity being released progressively modulates the output of the phototube, which signal is passed on to a video amplifier to increase amplitude of video signal. This very same video signal can be applied to the grid of the first video amplifier in a television receiver, or it can be used to modulate an r-f signal generator, and the actual picture signal can be picked up by front end of television receiver.

To obtain a stationary pattern on the screen of the television receiver picture tube, it is necessary that the scanning generators of the receiver be synchronized with the scanning generators of the video signal generator. This can be done by removing a sync pulse from the vertical and horizontal oscillator of the video signal generator (usually across a small resistor inserted into the cathode circuit of the scanning oscillators) to synchronize the scanning generators of the television receiver. Actually, if the sync signals are taken off such a low-impedance output as the scanning generator cathode circuit, they can be piped about the area through a coaxial cable and various take-off points can be set up. The sync pulses are applied to the respective horizontal and vertical scanning generators, making certain that the sync pulses are applied to a low-impedance portion of the scanning oscillator or through a very large series isolating resistor to prevent impedance of the synchronizing source from affecting the basic frequency of the receiver sweep oscillators. The series isolating resistor should be approximately ten times the resistance measured from grid to ground of the scanning generator.

A flying spot video signal generator in conjunction with a transparency of the RETMA standard test chart, as discussed in Chap. 12, forms a complete system for checking the performance of television receivers without having to resort to reception of station signals. It is, of course, possible to generate a truly FCC standard television signal by employing a standard 525 interlaced sync generator, in which case a flying spot will trace a standard interlace pattern. Blanking pulses can also be applied to the grid of the flying spot tube to cut off the beam during the retrace and blanking intervals.

MONOSCOPE TUBE AND GENERATOR

A second method of generating a video signal is by means of a monoscope tube and associated generators. The disadvantage of the monoscope system as compared to the flying spot scanner is that only one type of signal can be transmitted—namely, information permanently imprinted on the signal electrode. With a flying spot system, any number of transparencies can be used just by sliding them in front of the object lens. The monoscope tube, however, operates at a lower anode potential and requires no external lens and phototube assembly.

The monoscope tube itself is very similar to an iconoscope but generates a stronger video signal. It contains an electron gun and a second-anode system, which acts as a collector for secondary electrons removed from the pattern electrode. The pattern electrode consists of an aluminized plate upon which the chart or other pattern information appears in printer's ink. When the pattern electrode is scanned by the magnetically deflected scanning beam, signal is released from the back of the electrode which is applied to a video amplifier to form a video signal. When the scanning beam is striking an

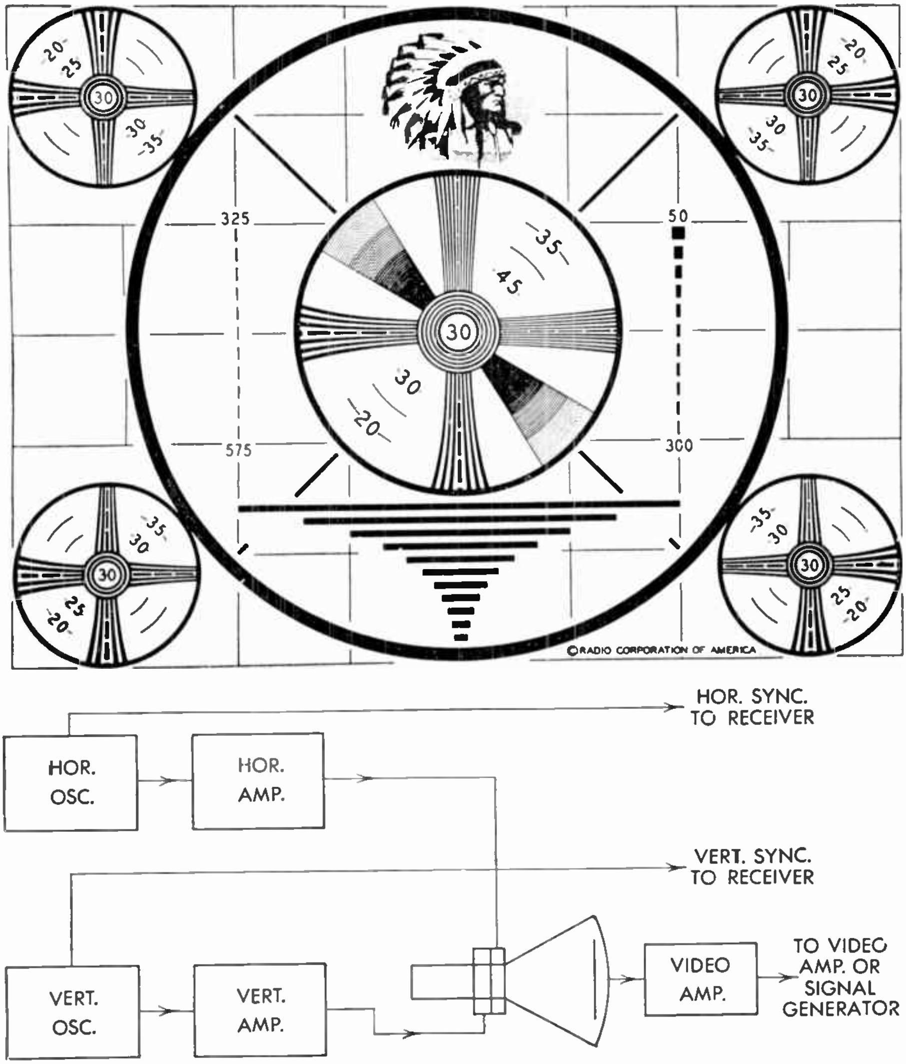


FIG. 323 Monoscope Video Signal Generator

aluminized section of the pattern, maximum secondary electrons are removed and the greatest signal current flows. When it strikes a dark section formed by the printer's ink, the secondary emission is very low; therefore, the output signal current is weak. In actual practice, and to generate a signal which has the same polarity as an iconoscope output signal (maximum signal for a black portion of the scene), a pattern electrode consists of a white on black or a reverse relation with respect to the generated signal. This arrangement is necessary to generate a signal for which the black section of the reproduced image generates the strongest signal output. Therefore, a monoscope tube can be substituted directly in a circuit in which an iconoscope has been in operation, and it will generate a signal with the same polarity arrangement as the original iconoscope signal (monoscope is not keystoned).

A regular, interlaced, standard sync generator can be used to generate the magnetic sweep voltages, or, for testing purposes, as shown in Fig. 323, a simple sync generator arrangement can be used to generate a video signal. In this system, a horizontal and vertical sawtooth oscillator generates the required sawtooth voltages, which are increased in amplitude by the associated magnetic deflection amplifiers. Thus, the pattern is scanned in regular sequence, element by element and line by line, generating a video signal at its output which is increased in amplitude and which can then be applied to the video amplifier of television receivers or used to modulate an r-f signal generator. The same synchronizing requirements must be met as in the flying spot system by removing sync pulses from the monoscope generator sweep oscillators and by conveying these sync pulses directly to the sweep oscillators of the television receivers, using the same distribution system mentioned previously. The same precaution must be observed to prevent loading of the receiver sweep oscillator by the synchronizing source.

The standard pattern, which appears on the pattern electrode of the RCA monoscope 2F21, shown in the second drawing of Fig. 323, contains information similar to the type of information which is conveyed by the standard RETMA test chart discussed previously. It is possible to check resolution, contrast, interlace, and frequency response as well as scanning linearity and aspect ratio. If a standard sync generator is used, picture signal generated is exactly the same as any chart transmitted by a television station and, therefore, gives an absolute indication of the performance of television receivers.

In a practical but noninterlaced, simple system, as shown in Fig. 323, frequency of the horizontal oscillator is set on 15,750 and the vertical oscillator on 60 cycles per second; a very close approximation of a standard signal can be generated. The only exception is that the pattern is not interlaced, this expedient very much reducing the complexity of the scanning generator. It means, however, that the vertical resolution of the reproduced picture will not be so good as that coming from an interlaced video signal generator or a standard received signal. However, this is unimportant in judging the performance of a television receiver because it is a known fact that the vertical resolution is largely dependent on the number of active scanning lines and the various beam spot sizes. So far as horizontal resolution is concerned, it is not impaired by using the noninterlaced system, and therefore, horizontal resolution can be checked accurately in the television receiver, horizontal resolution, of course, depending on the frequency response of the television receiver r-f, i-f, and video amplifiers. The horizontal resolution is unimpaired in this type of video generator because the scanning beam moves at the same velocity along the scanning line; therefore, the line information is being gathered at the same frequency it would be with an interlaced system. Likewise, to have the receiver reproduce an image, the same frequency response requirements must be set.

With a system for changing the centering and reducing the scanning width incorporated in the monoscope generator, it is possible to scan just the Indian (Fig. 323) at the top center of the standard pattern electrode. Scanning just this small area of the electrode generates a pattern which has approximately the same average brightness and contrast ratio as a studio pattern; it can be used to adjust the receiver for best performance, thereby indicating the performance of the receiver on the standard received television signal as it might be received during a studio show.

In summation, a flying spot scanning tube and monoscope tube can be used to generate a television signal which can be made exactly the same as the standard FCC signal or a close approximation of it. These video signal generators can be located in various points to generate signal continuously for purposes of alignment and checking performance of television receivers as well as transmitting equipment.

Another less costly but very effective composite signal generator is the Supreme Composite Video Generator. It forms a standard composite sync-blanking signal along with a video dot pattern (generated without a monoscope or scanner). It can be used to check receiver synchronization, linearity, and resolution.

164. *Complete Alignment of a Modern Intercarrier Receiver (UHF-VHF)*

As a conclusion to our discussion of alignment, the complete procedure of alignment is presented for a VHF-UHF receiver in the sequence recommended by the manufacturer. *To derive full benefit from this presentation, follow the procedure step by step referring to the schematic diagrams of the receiver.* Figs. 210b and 210c.

INSTALLATION INSTRUCTIONS The adjustments and checks described below are made prior to installation to determine if a receiver is functioning properly or requires actual alignment.

ION-TRAP MAGNET ADJUSTMENT Set the ion-trap magnet approximately in the position shown in Fig. 324. Starting from this position, immediately adjust the magnet by moving it forward or backward while at the same time rotating it slightly around the neck of the kinescope for the brightest raster on the screen.

DEFLECTION YOKE ADJUSTMENT If the lines of the raster are not horizontal or squared with the picture mask, rotate the deflection yoke until this condition is obtained. Tighten the yoke adjustment wing-screw.

PICTURE ADJUSTMENTS It will now be necessary to obtain a test-pattern

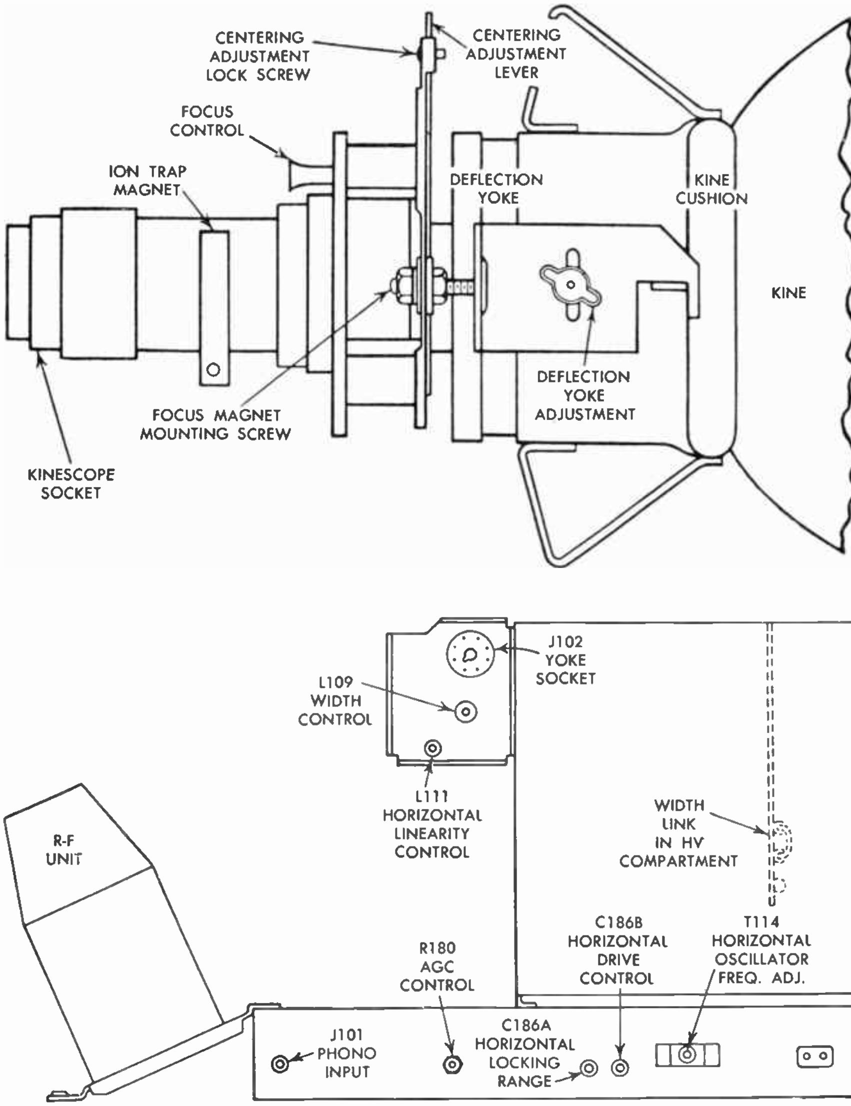


FIG. 324 Installation Adjustments

picture in order to make further adjustments. Connect the antenna transmission line to the receiver.

If the horizontal oscillator and a-g-c system are operating properly, it should be possible to sync the picture at this point. However, if the a-g-c control is misadjusted and the receiver is overloading, it may be impossible to sync the picture.

If the receiver is overloading, turn *R180* on the rear apron (see Fig. 324) counterclockwise until the set operates normally and the picture can be synchronized.

CHECK OF HORIZONTAL OSCILLATOR ALIGNMENT Turn the horizontal-hold control to the extreme counter-clockwise position. The picture should remain in horizontal sync. Momentarily remove the signal by switching off channel and then back to channel. Normally the picture will be out of sync. Slowly turn the control clockwise. The number of diagonal black bars will be gradually reduced; when only two or three bars sloping downward to the left are obtained, the picture will pull into sync upon slight additional clockwise rotation of the control. Pull-in should occur before the control has been turned 70 degrees from the extreme counterclockwise position. The picture should remain in sync for approximately 90 degrees of additional clockwise rotation of the control. At the extreme clockwise position, the picture should just begin to show a black bar in the picture on the left side.

If the receiver passes the above checks and if the picture is normal and stable, the horizontal oscillator is properly aligned. Skip "Alignment of Horizontal Oscillator," and proceed with "Centering Adjustment."

ALIGNMENT OF HORIZONTAL OSCILLATOR If in the above check the receiver failed to hold sync with the hold control either at the extreme counterclockwise position or over 90 degrees of clockwise rotation of the control from the pull-in point, it will be necessary to make the following adjustments.

HORIZONTAL FREQUENCY ADJUSTMENT Turn the horizontal-hold control to the extreme clockwise position. Tune in a television station, and adjust the *T114* horizontal frequency adjustment at the rear of the chassis until the picture is barely out of sync and the horizontal blanking appears as a vertical or diagonal black bar in the raster. Then turn the *T114* core until the bar is just visible at the extreme left side of the picture.

HORIZONTAL-LOCKING RANGE ADJUSTMENT Set the horizontal-hold control to the full counterclockwise position. Momentarily remove the signal by switching off and then back to channel. The picture may remain in sync. If so, turn the *T114* rear core slightly, and momentarily switch off channel. Repeat until the picture falls out of sync with the diagonal lines sloping to the left. Slowly turn the horizontal-hold control clockwise and note the least number of diagonal bars obtained just before the picture pulls into sync.

If more than two bars are present just before the picture pulls into sync, adjust the horizontal-locking range trimmer *C186A* slightly clockwise. If less than two bars are present, adjust *C186A* slightly counterclockwise. Turn the horizontal-hold control counterclockwise; momentarily remove the signal, and recheck the number of bars present at the pull-in point. Repeat this procedure until two or three bars are present.

Repeat the adjustments under "Horizontal Frequency Adjustment" and "Horizontal-Locking Range Adjustment" until the proper operating conditions specified under each are fulfilled. When the horizontal hold operates as outlined under "Check of Horizontal Oscillator Alignment" the oscillator is properly adjusted.

If it is impossible to sync the picture at this point and the a-g-c system is in proper adjustment, it will be necessary to adjust the horizontal oscillator by the method outlined in the alignment procedure.

FOCUS MAGNET ADJUSTMENTS The focus magnet should be adjusted to have approximately three-eighths of an inch of space between the rear cardboard shell of the yoke and the flat of the front-face of the focus magnet. This spacing gives the best average of focus over the face of the tube.

The axis of the hole through the magnet should be parallel with the axis of the kinescope neck, which extends through the middle of the hole.

CENTERING ADJUSTMENT No electrical centering controls are provided. Centering is accomplished by means of a separate plate on the focus magnet. The centering plate includes a locking screw which must be loosened before centering is attempted. Up-and-down adjustment of the plate moves the picture from side to side, and sidewise adjustment moves the picture up and down.

If a corner of the raster is shadowed, check the position of the ion-trap magnet. Reposition the magnet within the range of maximum raster brightness to eliminate the shadow, and recenter the picture by adjustment of the focus magnet plate. In no case should the ion-trap magnet be adjusted to cause any loss of brightness, since such operation may cause immediate or eventual damage to the tube. In some cases it may be necessary to shift the position of the focus magnet in order to eliminate a corner shadow.

WIDTH, DRIVE, AND HORIZONTAL-LINEARITY ADJUSTMENTS Adjustment of the horizontal drive control affects the high voltage applied to the kinescope. In order to obtain the highest possible voltage—hence the brightest and best-focused picture—adjust the horizontal-drive trimmer *C186B* for maximum drive (minimum capacity) consistent with a linear raster. Compression of the raster due to excessive drive can be seen as a white vertical bar or bars in the right half of the picture. Besides compression caused by excessive drive, another item to watch for is the change in linearity at the extreme left with changes of brightness-control setting. By proper adjustment of the linearity coil, the changes in linearity with changes in brightness can be made negligible. In general, to achieve this condition, the linearity coil should be set slightly on the high inductance side (core slightly clockwise) of the optimum position.

Pre-set the following adjustments as directed:

1. Place the width plug *P103* in the position of minimum width (top).
2. Set the width-control coil *L109* in approximately mid-position.
3. Set the linearity-control coil *L111* near minimum inductance (counterclockwise).
4. Set the drive capacitor *C186B* in the maximum drive position (counterclockwise).

If the raster is cramped or shows compression bars on the right half of the picture turn *C186B* clockwise until this condition is only barely eliminated.

Adjust the linearity-control coil *L111* clockwise until the best linearity and maximum deflection, or the best compromise, is obtained; then turn one-quarter turn clockwise from this position.

Retouch the drive trimmer *C186B* if necessary to obtain the best linearity and maximum width.

Check the horizontal linearity at various settings of the brightness control *R114A*. There should be no compression of the right half and no appreciable change of linearity, especially at the extreme left of the picture. If objectionable change does occur, turn the linearity coil *L111* slightly clockwise, and repeat the test.

Adjust the width control *L109* to fill the mask.

If the line voltage is low and it becomes impossible to fill the mask, move the width plug *P103* to the bottom position. The width coil *L109* is inoperative in this position.

HEIGHT AND VERTICAL-LINEARITY ADJUSTMENTS Adjust the height control (*R190* behind the front control panel) until the picture fills the mask vertically. Adjust vertical linearity (*R197* behind front control panel) until the test pattern is symmetrical from top to bottom. Adjustment of either control will require a readjustment of the other.

FOCUS Adjust the focus magnet for maximum definition in the test pattern's vertical "wedge" and for best focus in the white areas of the pattern.

Recheck the position of the ion-trap magnet to make sure that maximum brightness is obtained. If necessary readjust centering in order to align the picture with the mask.

CHECK OF RF OSCILLATOR Tune in all available VHF stations to determine if the receiver r-f oscillator is adjusted to the proper frequency on all channels. If adjustments are required, these should be made by the method outlined in the alignment procedure.

Note Some factory pre-aligned UHF inserts may require minor adjustment when installed in the tuner. This can be accomplished by using the UHF stations as a signal source.

CAUTION Observe, for reference, the initial positions of all cores before making any adjustments.

Set the fine-tuning control to the center of its range on each UHF channel which is to be adjusted. Adjust the oscillator core for each UHF channel in such manner as to obtain maximum audio output without distortion.

A-G-C THRESHOLD CONTROL The a-g-c threshold control *R180* is adjusted at the factory and normally should not require readjustment in the field.

To check the adjustment of the a-g-c threshold control, tune in a strong signal and sync the picture. Momentarily remove the signal by switching off and then back to channel. If the picture reappears immediately, the receiver is not overloading due to improper setting of *R180*. If the picture requires an appreciable portion of a second to reappear, or bends excessively, *R180* should be readjusted.

Turn *R180* fully counterclockwise. The raster may be bent slightly. This should be disregarded. Turn *R180* clockwise until there is a very, very slight bend or change of bend in the picture. Then turn *R180* counterclockwise just sufficiently to remove this bend or change of bend.

If the signal is weak, the above method may not work—it may be impossible to get the picture to bend. In this case, turn *R180* clockwise until the snow in the picture becomes more pronounced, then counterclockwise until the best signal-to-noise ratio is obtained.

The a-g-c control adjustment should be made on a strong signal, if possible. If the control is set too far clockwise on a weak signal, the receiver may overload when a strong signal is received.

FM-TRAP ADJUSTMENT In some instances interference may be encountered from a strong FM-station signal. A trap is provided to eliminate this type of interference. To adjust the trap, tune in the station on which the interference is observed and adjust the *L58* core on top of the antenna-matching transformer for minimum interference in the picture.

CAUTION In some receivers, the FM trap *L58* will tune down into channel *6* or even into channel *5*. Needless to say, such an adjustment will cause greatly reduced sensitivity on these channels. If channels *5* and *6* are to be received, check *L58* to make sure that it does not affect sensitivity on these two channels.

Replace the cabinet's back and connect the receiver-antenna leads to the cabinet's back. Make sure that the screws holding it are tight; otherwise, it may rattle or buzz when the receiver is operated at high volume.

165. *Step-by-Step Alignment Procedures*

TEST EQUIPMENT To service the television chassis of receiver properly, it is recommended that the following test equipment be available:

1. VHF sweep generator meeting the following requirements:
   a. Frequency ranges
      35 to 90 megacycles, 1- to 12-megacycle sweep width
      170 to 225 megacycles, 12-megacycle sweep width
   b. Output adjustable with at least 0.1-volt maximum
   c. Output constant on all ranges
   d. "Flat" output on all attenuator positions.
2. VHF signal generator to provide the following frequencies with crystal accuracy:
   a. Intermediate frequencies
      4.5, 39.25, 41.25, 43.5, 45.75, and 47.25 megacycles
   b. Radio frequencies

c. Output of these ranges should be adjustable and at least 0.1 volt maximum.

3. VHF heterodyne frequency meter with crystal calibrator if the signal generator is not crystal-controlled.

4. UHF sweep generator with a frequency range of from 470 to 890 megacycles.

5. UHF signal generator to provide the following frequencies with crystal accuracy:

Channel Number	*Picture-Carrier Frequency (In Megacycles)*	*Sound-Carrier Frequency (In Megacycles)*	*Receiver R-F Oscillator Frequency (In Megacycles)*
2	55.25	59.75	101
3	61.25	65.75	107
4	67.25	71.75	113
5	83.25	81.75	123
6	77.25	87.75	129
7	175.25	179.75	221
8	181.25	185.75	227
9	187.25	191.75	233
10	193.25	197.75	239
11	199.25	203.75	245
12	205.25	209.75	251
13	211.25	215.75	257
14	471.25	475.75	517
15	477.25	481.75	523
16	483.25	487.75	529
17	489.25	493.75	535
18	495.25	499.75	541
19	510.25	505.75	547
20	507.25	511.75	553
21	513.25	517.75	559
22	519.25	523.75	565
23	525.25	529.75	571
24	531.25	535.75	577
25	537.25	541.75	583
26	543.25	547.75	589
27	549.25	553.75	595
28	555.25	559.75	601
29	561.25	565.75	607
30	567.25	571.75	613
31	573.25	577.75	619
32	579.25	583.75	625
33	585.25	589.75	631
34	591.25	595.75	637
35	597.25	601.75	643
36	603.25	607.75	649
37	609.25	613.75	655
38	615.25	619.75	661
39	621.25	625.75	667
40	627.25	631.75	673
41	633.25	637.75	679

Channel Number	*Picture-Carrier Frequency (In Megacycles)*	*Sound-Carrier Frequency (In Megacycles)*	*Receiver R-F Oscillator Frequency (In Megacycles)*
42	639.25	643.75	685
43	645.25	649.75	691
44	651.25	655.75	697
45	657.25	661.75	703
46	663.25	667.75	709
47	669.25	673.75	715
48	675.25	679.75	721
49	681.25	685.75	727
50	687.25	691.75	733
51	693.25	697.75	739
52	699.25	703.75	745
53	705.25	709.75	751
54	711.25	715.75	757
55	717.25	721.75	763
56	723.25	727.75	769
57	729.25	733.75	775
58	735.25	739.75	781
59	741.25	745.75	787
60	747.25	751.75	793
61	753.25	757.75	799
62	759.25	763.75	805
63	765.25	769.75	811
64	771.25	775.75	817
65	777.25	781.75	823
66	783.25	787.75	829
67	789.25	793.75	835
68	795.25	799.75	841
69	801.25	805.75	847
70	807.25	811.75	853
71	813.25	817.75	859
72	819.25	823.75	865
73	825.25	829.75	871
74	831.25	835.75	877
75	837.25	841.75	883
76	843.25	847.75	889
77	849.25	853.75	895
78	855.25	859.75	901
79	861.25	865.75	907
80	867.25	871.75	913
81	873.25	877.75	919
82	879.25	883.75	925
83	885.25	889.75	931

6. Cathode ray oscilloscope: An oscilloscope with a sensitivity of 1 millivolt per inch is required. A suitable preamplifier may be employed with oscilloscopes of lesser sensitivity.

7. Electronic voltmeter: A voltmeter with a 1.5-volt d-c scale is required.

8. D-C milliammeter: A milliammeter with a range of from 0 to 50 milliamperes full scale.

9. Adapter socket: An adapter socket is required to meter the cathode cur-

rent of the 6S4 voltage control tube of the *KRK12* tuner. Wiring of the adapter is shown in Fig. 325.

PICTURE I-F TRAP ADJUSTMENT Connect the i-f signal generator across the link circuit on terminals *A* and *B* of *T104,* Fig. 210c.

Connect the VTVM to the junction of *R133* and *C133B.*

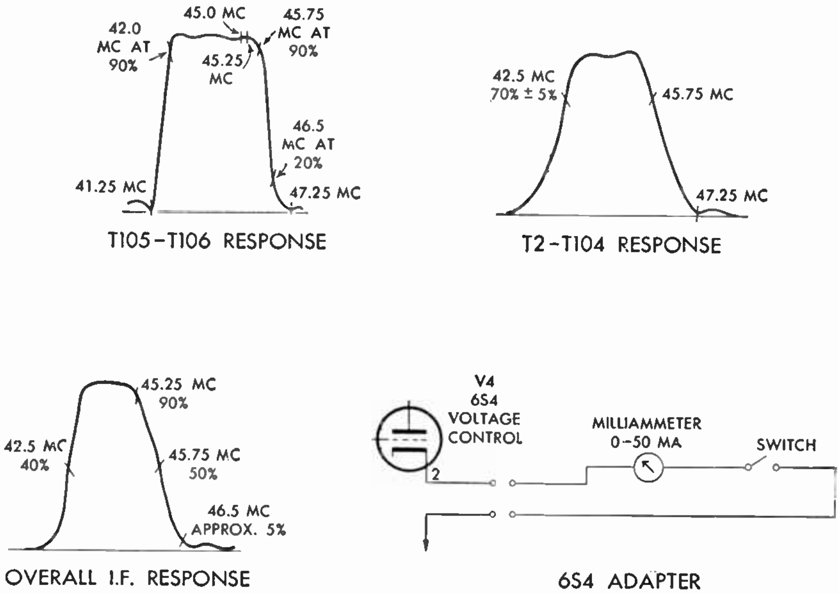


FIG. 325 Alignment Response Curves

Obtain two 7.5-volt batteries capable of withstanding appreciable current drain, and connect the ends of a 1000-ohm potentiometer across each. Connect the positive terminal of one battery to the chassis and the potentiometer arm to the junction of *R133* and *C133B.*

Set the bias to produce approximately −1.0 volt of bias at the junction of *R133* and *C133B.*

Connect the VTVM to pin *9* of *V110,* the *6CL6* video amplifier.

Set the signal generator to each of the following frequencies, and adjust the corresponding circuit for minimum d-c output at pin *9* of *V110.* Use sufficient signal input to produce 1.0 volt of d-c on the meter when the final adjustment is made.

39.25 megacycles	*T104* top core
41.25 megacycles	*T105* bottom core
47.25 megacycles	*T106* bottom core

PICTURE I-F TRANSFORMER ADJUSTMENTS Set the signal generator to each of the following frequencies, and peak the specified adjustment for maximum indication on the "VoltOhmyst." During alignment, reduce the input signal, if necessary, in order to produce 1.0 volt of d-c at pin *9* of *V110* with −1.0 volt of i-f bias at the junction of *R133* and *C133B*.

43.7 megacycles	*T109*
45.5 megacycles	*T108*
41.8 megacycles	*T107*

To align *T105* and *T106,* connect the sweep generator to the first picture i-f grid, pin *1* of *V106* through a 1000 micromicrofarad ceramic capacitor. Shunt *R137, R141,* and terminals *A* and *F* of *T109* with 330-ohm composition resistors. Set the i-f bias to −1.0 volt at the junction of *R133* and *C133B*.

Connect the oscilloscope to pin *9* of *V110,* the *6CL6* video amplifier.

Adjust *T105* and *T106* top cores for maximum gain and curve-shape as shown in Fig. 325. For final adjustment, set the output of the VHF sweep generator to produce 0.5 volt peak-to-peak at the oscilloscope terminals.

To align the crystal mixer, *T2,* and *T104,* connect the VHF sweep generator to the front terminal of the *1N82* crystal holder in series with a 1500-micromicrofarad ceramic capacitor. Use the shortest leads possible, grounding the sweep generator to the tuner case.

Set the channel selector to channel *5*.

Connect a 180-ohm composition resistor between terminal *B* of *T105* and the junction of *R131* and *C133A*.

Connect the oscilloscope diode probe to terminal *B* of *T105* and ground. Couple the signal generator loosely to the diode probe in order to obtain markers.

The shunt trimmer *C121* across terminals *A* and *B* of *T104* is variable and is provided as a bandwidth adjustment. Pre-set the shunt trimmer to minimum capacity. Adjust *T2* (top) and *T104* (bottom) for maximum gain at 43.5 megacycles and with 45.75 megacycles at 70 per cent of maximum response.

Adjust the shunt trimmer *C121* until the 42.5-megacycle point is at 70 per cent response with respect to the low-frequency shoulder. Adjust *T1* for maximum gain. Readjust *T2* and *T104* if necessary to obtain proper wave shape.

Disconnect the diode probe and the 180-ohm and the three 330-ohm resistors.

SWEEP ALIGNMENT OF PICTURE I-F Connect the oscilloscope to pin *9* of *V110*.

Adjust the bias potentiometer to obtain −6.0 volts of bias as measured by a VoltOhmyst at the junction of *R133* and *C133B*.

Leave the sweep generator connected to the front terminal of the *1N82* crystal holder on tuner. Use the shortest leads possible with not more than one inch of unshielded lead at the end of the sweep cable. If these precautions are not observed, the receiver may be unstable and the response curves obtained may be unreliable.

Adjust the output of the sweep generator to obtain 3.0 volts peak-to-peak on the oscilloscope.

Couple the signal generator loosely to the grid of the first pix i-f amplifier. Adjust the output of the signal generator to produce small markers on the response curve.

Retouch *T108* and *T109* to obtain the response shown in Fig. 325. Do not adjust *T107* unless it is absolutely necessary. If *T107* is adjusted to be too low in frequency, it will raise the level of the 41.25-megacycle sound i-f carrier and may create interference in the picture. It will also cause poor adjacent-channel picture rejection. If *T107* is tuned too high in frequency, the level of the 41.25-megacycle sound i-f carrier will be too low and may produce noisy sound in weak-signal areas.

Remove the oscilloscope sweep and signal generator connections.

Remove the bias-box employed to provide bias for alignment.

TUNER ALIGNMENT *Tuner VHF Alignment.* Remove the 6S4 voltage control tube from its socket and insert the adapter. Insert the 6S4 in the adapter.

Connect the 0- to 50-milliampere meter to the adapter socket leads, and turn on the adapter switch.

Remove the tuner cover shield.

Rotate the channel selector to a point midway between channels, disengaging the insert contacts, and observe the non-oscillating plate current. Some tubes may oscillate even with the tuned circuits disengaged. To be sure, the oscillator is in a non-oscillatory state. Short-circuit with a finger the spring contacts *12* and *13*, the two contacts nearest the tuner front.

Note: The contacts are at zero d-c potential. Should the plate current rise, keep a finger on the contacts while adjusting the oscillator plate current. Adjust *R6*, oscillator voltage control, for a 28-milliampere reading on the meter.

Replace the tuner cover shield.

Connect the VHF sweep generator to the antenna terminals.

Connect the VHF signal generator loosely to the antenna terminals.

Connect the oscilloscope, through the pre-amplifier if needed with the oscilloscope used, to test point *TP1*.

Ground the a-g-c bias at the tuner terminal board, using a clip-lead to insure that the bias will remain constant.

Turn off the adapter switch, removing the plate voltage from the oscillator. This is required because of r-f—i-f interaction when a crystal is used as a mixer.

Set the channel selector and the sweep generator to channel *2*.

Insert markers of the channel *2* picture carrier and sound carrier, 55.25 megacycles and 59.75 megacycles.

Adjust antenna *T6*, r-f amplifier plate *L29*, and mixer *L30* adjustments for a symmetrical curve with maximum gain at the center of the pass band. The curves will have a deep valley because of no crystal loading and nonlinear detector characteristics. The limits for the 100 per cent response points are shown in Fig. 326. If the bandwidth is out of tolerance, it can usually be corrected by redressing the coupling capacitor of the double-tuned circuit, *C40* on insert *A*. Maximum bandwidth occurs when the capacitor is centered in the insert chamber.

Repeat the above steps for all VHF channels, adjusting the appropriate antenna, r-f amplifier plate, and mixer slugs for a symmetrical curve with maximum gain at the center of the pass band.

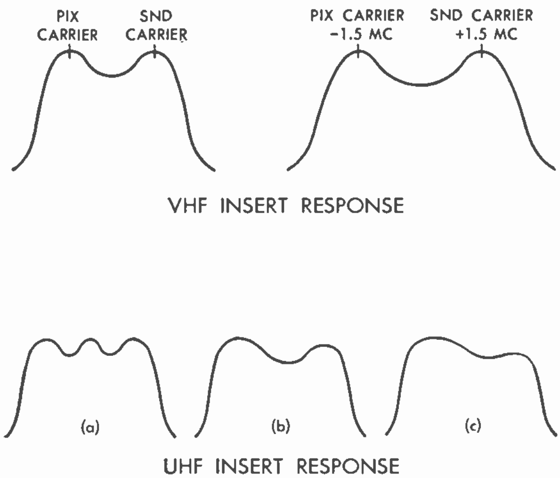


FIG. 326 Tuner Response Curves

Turn off the sweep generator.

Remove the oscilloscope, and pre-amplifier if used, from test point *TP1*.

Turn the a-g-c control fully clockwise, and remove the clip-lead grounding the a-g-c bias on the tuner terminal board.

Connect the potentiometer arm of one of the bias supplies to the a-g-c terminal on the tuner, and ground the battery positive terminal to the tuner case. Adjust the bias potentiometer to produce −3.5 volts of bias, as measured by the VoltOhmyst at the a-g-c terminal on the tuner.

Connect the potentiometer arm of the second-bias supply to the junction *R133* and *C133B*, and ground the positive battery terminal. Adjust the bias potentiometer to produce −5 volts of i-f bias as indicated on the VoltOhmyst at the junction point.

Connect the oscilloscope to pin *9* of *V110*. Use 3- to 5-volt peak-to-peak output on the oscilloscope.

Turn on the adapter switch to apply plate voltage to the oscillator.

Turn the channel selector to channel *13*.

Set the fine-tuning control to the center of its range.

Adjust the oscillator slug *L22* to proper frequency, 257 megacycles. This may be done in several ways. The easiest way—the way which will be recommended in this procedure—is to use the signal generator as a heterodyne frequency meter and to beat the oscillator against the signal generator. To do this, tune the signal generator to 257 megacycles with crystal accuracy. Insert one end of a piece of insulated wire into the tuner, through either of the two holes next to the oscillator tube on the right, front, top corner of the tuner. Be careful that the wire does not touch any of the tuned circuits, as it may cause the frequency of the oscillator to shift. Connect the other end of the wire to the "r-f in" terminal of the signal generator. Adjust the *L22* oscillator slug to obtain an audio beat with the signal generator.

Turn on the sweep generator and set to channel *13*. Adjust *T1* for maximum gain on the oscilloscope. Adjust mixer tank circuit *L21* for maximum gain and flat-topped curve. Recheck *T1* for maximum gain at the center of the band with the proper response. Maximum gain and flat-topped response should be obtained simultaneously.

Adjust the oscillator to frequency on all VHF channels by switching the receiver and signal generator to each VHF channel and adjusting the appropriate oscillator slug to obtain a beat with the signal generator. Adjust the appropriate mixer slug where necessary to obtain maximum gain and proper curve-shape as explained above.

Adjust the tunable i-f trap *C16-L7*. To do this, connect the signal generator to the fixed i-f trap *C2-L2* at the end opposite the antenna terminal plug. Set the signal generator to 43.5 megacycles, and adjust the output of the signal generator to obtain sufficient indication on the oscilloscope. Tune the i-f trap *C16-L7* for minimum marker indication on the oscilloscope.

Remove the signal generator and the oscilloscope.

Tuner UHF Alignment. To align the UHF inserts:

Turn off the adapter switch, removing plate voltage from the oscillator.

Ground the a-g-c bias at the tuner terminal board, using a clip-lead to insure that the bias will remain constant.

Connect the oscilloscope, through the preamplifier if needed with oscilloscope used, to test point *TP1*.

Connect the UHF sweep generator to the antenna terminals. Use a 10-decibel attenuator pad to assure proper alignment.

Connect the UHF signal generator loosely to the antenna terminals.

Set the channel selector to the desired position and the sweep generator to sweep the frequency of the insert being used.

Insert markers of the picture carrier and sound carrier for the desired channel.

Adjust the UHF antenna, link coupling, and mixer adjustments for a symmetrical curve, with maximum gain, centered about the pass band.

The responses are shown in Fig. 326. The curve-shape will usually vary from a to c, Fig. 326, going higher in frequency; however, any of these responses is acceptable.

Repeat the above steps for all UHF inserts used, adjusting the appropriate antenna, link coupling, and mixer slugs for a symmetrical curve, with maximum gain, centered about the pass band.

Remove the oscilloscope and preamplifier, if used, from test point *TP1*.

Remove the clip-lead grounding the a-g-c bias on the tuner terminal board.

Connect the potentiometer arm of one of the bias supplies to the a-g-c terminal on the tuner, and ground the battery positive terminal to the tuner case. Adjust the bias potentiometer to produce −3.5 volts of bias, as measured by the VoltOhmyst at the a-g-c terminal.

Connect the potentiometer arm of the second bias supply to the junction of *R133* and *C133B*, and ground the positive battery terminal. Adjust the bias potentiometer to produce −5 volts of i-f bias as indicated on the VoltOhmyst at the junction point.

Connect the oscilloscope to pin *9* of *V110*. Use 3- to 5-volt peak-to-peak output on the oscilloscope.

Turn on the adapter switch to apply plate voltage to the oscillator.

Turn the channel selector to the lowest UHF channel to be used, and set the fine-tuning control to the center of its range.

Adjust the oscillator core to proper frequency. To do this, connect the VHF signal generator to test point *TP1* with the shortest leads possible. Insert a 45.75-megacycle marker from the VHF generator.

Set the UHF sweep generator to sweep the desired channel, and observe the output on the oscilloscope. If the sweep generator is not sweeping the correct frequency range, it may be necessary to readjust the sweep in order to place the 45.75 marker on the over-all response curve.

Set the UHF marker generator to the picture carrier of the channel insert being adjusted and connect to test point *TP1*.

Adjust the oscillator core until the markers for 45.75 megacycles and the picture carrier coincide on the sweep pattern on the oscilloscope.

Adjust the mixer core for maximum gain with proper wave-shape.

Connect the VoltOhmyst to test point *TP1*, using a 1.5-volt d-c scale.

Set the oscillator injection adjustment to read 0.1 volt on the VoltOhmyst.

Repeat the above steps for all UHF inserts, adjusting the oscillator injection control only if the reading on the VoltOhmyst exceeds 0.3 volt. Adjust as necessary to read 0.3 volt or less at *TP1*.

RATIO DETECTOR ALIGNMENT In order to obtain good ratio detector alignment on AM-modulated signal, a generator that is exceptionally free from FM modulation must be employed. Set the signal generator at 4.5 megacycles, and connect it to the second, sound i-f grid, pin *1* of *V102*. Set the generator for 30-per cent 400-cycle modulation.

Connect the VoltOhmyst to the junction of *R111* and *C111*.

Connect the oscilloscope across the speaker voice coil, and turn the volume control for maximum output.

Tune the ratio detector primary *T102* top core for maximum d-c output on the VoltOhmyst. Adjust the signal level from the signal generator for −10 volts on the VoltOhmyst when finally peaked. This is approximately the operating level of the ratio detector for average signals.

Connect the VoltOhmyst to the junction of *R110* and *C110*.

Adjust the *T102* bottom core for zero d-c on the meter. Then, turn the core to the nearest minimum AM output on the oscilloscope.

Repeat adjustments of *T102* top for maximum d-c and *T102* bottom for minimum output on the oscilloscope, making final adjustment with the 4.5-megacycle input level adjusted to produce 10 volts d-c on the VoltOhmyst at the junction of *R111* and *C111*.

Connect the VoltOhmyst to the junction of *R110* and *C110*, and note the amount of d-c present. If this voltage exceeds ± 1.5 volts, adjust *R108* by turning it in until zero d-c is obtained. Readjust the *T102* bottom core for minimum output on the oscilloscope. Repeat adjustments of *R108* and *T102* bottom core until the voltage at *R110* and *C110* is less than ± 1.5 volts when *T102* bottom core is set for minimum output on the oscilloscope.

Connect the VoltOhmyst to the junction of *R111* and *C111*, and repeat *T102* top core for maximum d-c on the meter; again reset the generator, now having -10 volts on the meter.

Repeat the adjustments described in the two preceding paragraphs until the voltage at *R110* and *C110* is less than plus or minus 1.5 volts when the *T102* top core is set for maximum d-c at the junction of *R111* and *C111* and the *T102* bottom core is set for minimum indication on the oscilloscope.

SOUND I-F ALIGNMENT Connect the sweep generator to the first, sound i-f amplifier grid, pin *1* of *V101*. Adjust the generator for a sweep width of 1 megacycle at a center frequency of 4.5 megacycles.

Insert a 4.5-megacycle marker signal from the signal generator into the first, sound i-f grid.

Connect the oscilloscope in series with a 10,000-ohm resistor to terminal *A* of *T101*.

Adjust *T101* top and bottom cores for maximum gain and symmetry about the 4.5-megacycle marker on the i-f response. The pattern obtained should be similar to that shown in Fig. 327.

The output level from the sweep should be set to produce approximately 2.0 volts peak-to-peak at terminal *A* of *T101* when the final touches on the above adjustment are made. It is necessary that the sweep output voltage should not exceed the specified values; otherwise the response curve will be broadened, permitting slight misadjustment to pass unnoticed and possibly causing distortion on weak signals.

Connect the oscilloscope to the junction of *R110* and *C110*, and check the linearity of the response. The pattern obtained should be similar to that shown in Fig. 327.

SOUND TAKE-OFF ALIGNMENT Connect the 4.5-megacycle generator in series with a 1000-ohm resistor to terminal *C* of *T110*. The input signal should be approximately 0.5 volt

Short the fourth, pix i-f grid to ground, pin *1*, *V109*, to prevent any noise from masking the output indication.

Connect the crystal diode probe of a VoltOhmyst to the plate of the video amplifier, pin *6* of *V110*.

Adjust the core of *T110* for minimum output on the meter.

Remove the short from pin *1*, *V109*, to ground, if used.

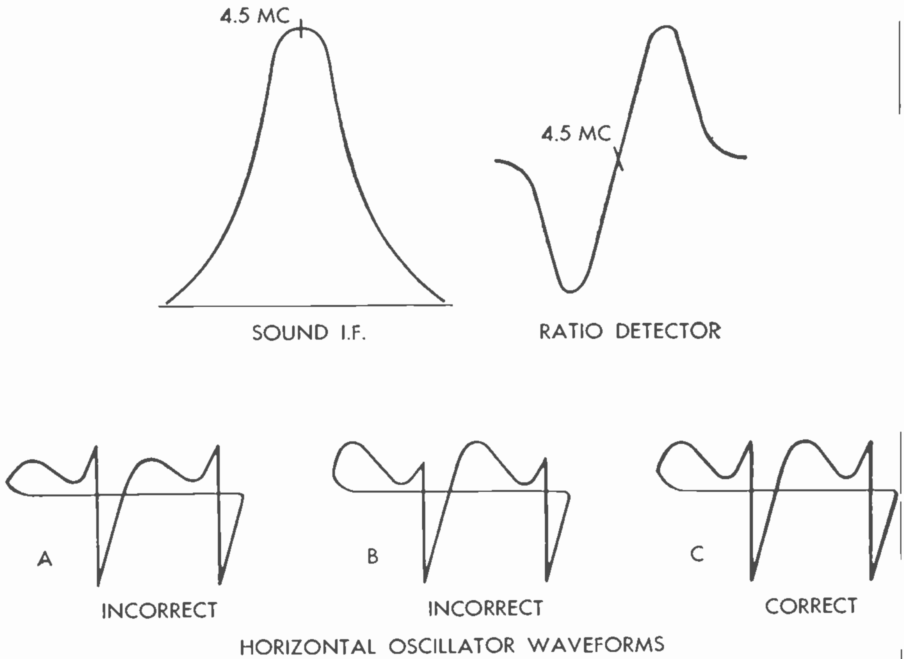


FIG. 327 Additional Alignment Curves

HORIZONTAL OSCILLATOR ADJUSTMENT Normally the adjustment of the horizontal oscillator is not considered to be a part of the alignment procedure, since the oscillator waveform adjustment may require the use of an oscilloscope. It cannot be done conveniently in the field. The waveform adjustment is made at the factory and normally should not require readjustment in the field. However, the waveform adjustment should be checked whenever the receiver is aligned or whenever the horizontal oscillator operation is improper.

HORIZONTAL FREQUENCY ADJUSTMENT Tune in a station and sync the picture. If the picture cannot be synchronized with the horizontal-hold control *R210*, then adjust the *T114* frequency core on the rear apron until the picture will synchronize. If the picture still will not sync, turn the *T114* waveform adjustment core (under the chassis) out of the coil several turns from its original position, and readjust the *T114* frequency core until the picture is synchronized.

Examine the width and linearity of the picture. If picture width or linearity is incorrect, adjust the horizontal drive control *C186B*, the width control *L109*, and the linearity control *L111* until the picture is correct.

HORIZONTAL OSCILLATOR WAVEFORM ADJUSTMENT The horizontal oscillator waveform may be adjusted by either of two methods. The method outlined in paragraph A below may be employed in the field when an oscilloscope is not available. The service shop method outlined in paragraph B below requires the use of an oscilloscope.

A. Turn the horizontal-hold control completely clockwise. Place adjustment tools on both cores of *T114*, and be prepared to make simultaneous adjustments while watching the picture on the screen. First, turn the *T114* frequency core (on the rear apron) until the picture falls out of sync and one diagonal black bar sloping down to the right appears on the screen. Then, turn the waveform adjustment core (under the chassis) into the coil while at the same time adjusting the frequency core to maintain one diagonal black bar on the screen. Continue this procedure until the oscillator begins to "motorboat"; then turn the waveform adjustment core out until the motorboating barely stops. As a check, turn the *T114* frequency core until the picture is synchronized; then reverse the direction of rotation of the core until the picture begins to fall out of sync with the diagonal bar sloping down to the right. Continue to turn the frequency core in the same direction. Additional bars should not appear on the screen. Instead, the horizontal oscillator should begin to motorboat. Retouch the adjustment of the *T114* waveform adjustment core, if necessary, until this condition is obtained.

B. Connect the low-capacity probe of an oscilloscope to terminal *C* of *T114*. Turn the horizontal-hold control one-quarter turn from the clockwise position to bring the picture in sync. The pattern on the oscilloscope should be as shown in Fig. 327. Vary the waveform adjustment core of *T114* until the two peaks are at the same height. During this adjustment, the picture must be kept in sync by readjusting the hold control if necessary.

This adjustment is very important for correct operation of the circuit. If the broad peak of the wave on the oscilloscope is lower than the sharp peak, the noise immunity becomes poorer, the stabilizing effect of the tuned circuit is reduced, and drift of the oscillator becomes more serious. On the other hand, if the broad peak is higher than the sharp peak, the oscillator is over-stabilized, the pull-in range becomes inadequate, and the broad peak can cause double triggering of the oscillator when the hold control approaches the clockwise position.

Remove the oscilloscope upon completion of this adjustment.

HORIZONTAL-LOCKING RANGE ADJUSTMENT Set the horizontal-hold control to the full counterclockwise position. Momentarily remove the signal by switching off and then back to channel. The picture may remain in sync. If so, turn the *T114* frequency core slightly, and momentarily switch off channel. Repeat until the picture falls out of sync with the diagonal lines sloping to the left. Slowly turn the horizontal-hold control clockwise, and note the least number of diagonal bars obtained just before the picture pulls into sync.

If no more than three bars are present just before the picture pulls into sync, adjust the horizontal-locking range trimmer *C186A* slightly clockwise. If less than two bars are present, adjust *C186A* slightly counterclockwise. Turn the horizontal-hold control counterclockwise; momentarily remove the signal, and recheck the number of bars present at the pull-in point. Repeat this procedure until two or three bars are present.

Turn the horizontal-hold control to the maximum clockwise position. Adjust the *T114* frequency core so that the diagonal bar sloping down to the right appears on the screen, and then reverse the direction of adjustment so that the bar barely moves off the screen, leaving the picture in synchronization.

SENSITIVITY CHECK A comparative sensitivity check can be made by operating the receiver on a weak signal from a television station and comparing the picture and sound obtained to that obtained on other receivers under the same conditions.

This weak signal can be obtained by connecting the shop antenna to the receiver through a ladder-type attenuator pad. The number of stages in the pad depends upon the signal strength available at the antenna. A sufficient number of stages should be inserted so that a somewhat less than normal contrast picture is obtained when the picture control is at the maximum clockwise position. Only carbon type resistors should be used to construct the pad.

RESPONSE CURVES The response curves shown and referred to throughout the alignment procedure were taken from a production set. Although these curves are typical, some variations can be expected.

The response curves are shown in the classical manner of presentation, that is, with "response up" and low frequency to the left. The manner in which they will be seen in a given test set-up will depend upon the characteristics of the oscilloscope and the sweep generator. The curves may be seen as inverted and/or switched from left to right, depending on the deflection polarity of the oscilloscope and the phasing of the sweep generator.

NOTE ON TUNER ALIGNMENT Because of the frequency spectrum involved and the nature of the device, many of the tuner leads and components are critical in some respects. Even the power-supply leads form loops which couple to the tuned circuits, and if resonant at any of the frequencies involved in the performance of the tuner, may cause serious departures from the desired characteristics. In the design of the receiver these undesirable resonant loops have been shifted far enough away in frequency to allow reasonable latitude in their components and physical arrangement without being troublesome. When the tuner is aligned in the receiver, no trouble from resonant loops should be experienced. However, if the tuner is aligned in a jig separate from the receiver, attention should be paid to insure that unwanted resonances, which might present a faulty representation of tuner alignment, do not exist.

166. *Trouble-Shooting Visually*

Despite the complexities of the television receiver, localization of most defects is not difficult if the technician has a thorough understanding of television circuit functions and some additional information on the special functioning of the specific model under test. Inasmuch as receiver troubles, except sound-channel failures, manifest themselves in some form of malfunction of the presentation on the picture-tube screen, many defects can be isolated by visual observation. The ability to interpret these malfunctions quickly and accurately comes with experience and a thorough knowledge of receiver techniques. In repairing TV receivers, a hit-or-miss technique for trouble shooting, without giving any thought to the nature of defect, will cause waste of many hours.

The angle of attack in locating troubles is first to localize the defect to a specific section of the receiver by visual observation of the scanning raster, and then to trouble-shoot defective section to find bad component. In general, troubles can be isolated readily to the following major sections of most receiver models:

POWER SUPPLY

A failure of the low-voltage power supply, of course, results in nonoperation of the entire receiver, picture, and sound sections. A decrease in supply voltage causes weak operation of various sections of the receiver. A loss of voltage will generally affect sweep and sync systems first, which require a higher voltage and are more critical to proper voltage. This results in unstable synchronism (inability to hold picture locked in) and inability to obtain proper raster height or width with satisfactory linearity. Hum on the low-voltage d-c causes a hum bar pattern to appear on the image and a sinusoidal hum pattern on sides or top and bottom of the scanning raster.

Failure of the high-voltage supply causes inoperation of the picture tube and no scanning raster although the sound will be normal. A reduction in high-voltage d-c results in a loss of brightness and, if substantial, a loss of resolution and inability to obtain sharp focus. In many modern receivers, the high voltage is obtained from a transient in the horizontal sweep amplifier; therefore, a reduction in low-voltage d-c also reduces high voltage and screen brightness. It is apparent that a failure of the horizontal sweep in this type of receiver also results in loss of high voltage and screen illumination.

R-F SECTION

An r-f section failure, because this section is as necessary in receiving the sound as it is in picking up the picture, causes both sound and picture to be lost if unit is entirely out. Scanning raster will be normal. A weak, insensitive r-f unit will produce a weak, faded-looking, and often noisy picture and a weak sound output.

If the local oscillator is not functioning, picture and sound will be absent. If mistuned, there is inability to tune for presentation of best picture and sound. (Improper alignment of i-f system causes same type of defect although in this case either one or the other can be tuned in satisfactorily but not together.)

If an r-f amplifier or mixer defect causes a loss in gain, picture is weak and noisy. If r-f bandwidth is affected, picture resolution degenerates.

SOUND I-F SECTION

A sound i-f failure will manifest itself as a normal picture with no sound. A defect which causes loss of gain in the i-f channel produces weak sound output. Often improper alignment of the sound system, especially the tuned circuits which remove the sound from the i-f, causes a weak or distorted sound output and substantial sound interference on the picture. It is important that the sound i-f be tuned precisely not only to boost the sound output but to prevent sound buzz.

PICTURE I-F SYSTEM

A dead picture i-f system manifests itself as a normal sound with absence of picture although scanning raster is normal. If defect causes a reduction in gain of the i-f system, synchronism will be poor and the picture weak. Contrast will be set near maximum to obtain even a weak picture. In most cases, a defect in the i-f system affects the bandpass of the i-f; therefore, resolution also deteriorates. If bandpass is affected, phase distortion is also likely to be present which will cause low-frequency streaking or high-frequency transients on the picture.

It is important to note at this point that a failure in the video amplifier causes a defective picture and leaves a normal or weak sound.

VIDEO AMPLIFIER

If there is complete failure of the video amplifier, the picture will not be present, and scanning raster will be normal. If defect causes a loss in gain of the video amplifier, signal is weak and contrast poor. Contrast will have to be set on maximum to obtain any picture at all. A loss of high-frequency components due to a video amplifier defect causes loss of picture resolution; a loss of lows causes low-frequency streaking. When high frequencies are overcompensated, transients appear on the picture.

It is important to note that the same types of defects are caused by video amplifier defects as picture i-f defects, and a few simple checks must occasionally be made to isolate trouble to one section. If receiver has a separate sync detector. a defect in the i-f system also affects synchronism, while synchronism is satisfactory if defect is in the video amplifier. This does not apply to those receivers which remove sync at output of last video amplifier. A sweep response check of the i-f or a gain check of the video amplifier assists in localizing the defect. In case of a dead picture system, a simple signal-tracing procedure can be used by applying a modulated signal generator at the i-f frequency to the grid of first i-f amplifier and by searching for the point where the signal drops out in i-f system or video amplifier.

SYNC SYSTEM

When there is a sync system failure, sound is normal and picture signal—although discernible on screen—is unstable. Type of sync defect is ascertained by observing if tear-out is horizontal or vertical. Improper setting of contrast control also causes unstable sync—too weak sync causing loss of synchronism, too high causing sync clipping in the video amplifier and also resulting in absence of sync in the sync stages.

VERTICAL SWEEP

Complete failure of the vertical sweep causes a thin horizontal line to appear on the screen instead of the scanning raster. Sound is normal. If defect causes a loss in sweep amplitude, it is not possible to obtain proper raster height with correct linearity. Other vertical sweep defects cause incorrect vertical linearity or inability to lock the pattern vertically.

HORIZONTAL SWEEP

Complete failure of the horizontal sweep in some receivers causes a thin, bright vertical line to appear on the screen. In receivers which use horizontal transient to generate picture-tube high voltage, a failure of the sweep also causes a loss of high voltage and no screen illumination. If defect causes loss in sweep amplitude, it is not possible to obtain full scanning raster width with correct linearity. With transient high-voltage system, screen illumination is also reduced. Other horizontal sweep defects cause poor linearity, foldover, or unstable synchronism.

It is obvious that by simply analyzing the behavior of the picture and scanning raster it is possible to isolate trouble to a receiver section with, at the most, four or five stages as the section in which the defect exists. The next task is to isolate the defective tube or component within that section. Here, again, as before, we cannot overstress the importance of understanding what each stage is to do and what approach is to be taken in checking its performance.

QUESTIONS

1. What precautions would you take in using an oscilloscope to check sweep waveforms?
2. List applications for an a-c VTVM in checking television receiver performance.
3. Describe basic systems for obtaining sweeping frequencies.
4. List every possible precaution to be taken in connecting equipment for a sweep oscillator alignment.
5. What are the characteristics of a satisfactory sweep oscillator?

6. What are the characteristics of a satisfactory oscilloscope?
7. Describe alignment of a typical r-f unit.
8. Describe alignment of a typical TV sound i-f section.
9. Describe alignment of a typical picture i-f section.
10. Describe a typical video signal generator.
11. Explain theory of crosshatch pattern.
12. Detail procedure for performing a video amplifier response measurement.

Chapter 14

PRACTICAL TELEVISION MATHEMATICS

167. *Formulas of the Vacuum Tube*

The three basic formulas of the vacuum tube—current increment, constant voltage, and constant current—also serve as base equations from which the specific formulas used in television application are developed. When a signal is applied to the grid of a vacuum tube, there occurs a change in plate current. This change in plate current also causes a change in plate voltage. Inasmuch as a change in plate voltage also has its effect on plate current, the actual plate current is a resultant of two plate-current increments or

$$i_p = i_1 + i_2$$

in which i_1 is the change in plate current caused by a change in grid voltage; i_2 a change in plate current caused by a change in plate voltage. The plate current change caused by a change in grid voltage for a given tube is dependent on the g_m or mutual conductance.

$$g_m = \frac{i_p}{e_s} \qquad \text{or} \qquad i_1 = g_m\, e_s$$

where E_p = constant

Likewise, the change in plate current contributed by the plate-voltage variation is dependent on the plate resistance (a-c)

$$r_p = \frac{e_p}{i_p} \qquad \text{or} \qquad i_2 = \frac{e_p}{r_p}$$

where E_g = constant

It follows that under conditions of a change in grid voltage and plate voltage the resultant plate current is

$$i_p = g_m\, e_s + \frac{e_p}{r_p} \tag{1}$$

In a vacuum-tube amplifier circuit (Fig. 328A) the plate-voltage change is a result of the plate-current variation in the load or

$$e_p = -i_p Z_L$$

and therefore

$$i_p = g_m e_s - \frac{i_p Z_L}{r_p}$$

$$i_p + \frac{i_p Z_L}{r_p} = g_m e_s$$

$$i_p = \frac{g_m e_s}{1 + \dfrac{Z_L}{r_p}} = (g_m e_s)\left(\frac{r_p}{r_p + Z_L}\right)$$

The voltage across the output is

$$e_o = -i_p Z_L = -g_m e_s \frac{r_p Z_L}{r_p + Z_L} \tag{2}$$

Observation of Eq. (2) indicates that r_p and Z_L act in parallel (product over sum), and an equivalent circuit is as shown in Fig. 328B, consisting of plate resistance and load in shunt driven by a constant current $-g_m e_s$. Equation (2) represents the vacuum tube as a constant-current device. This is par-

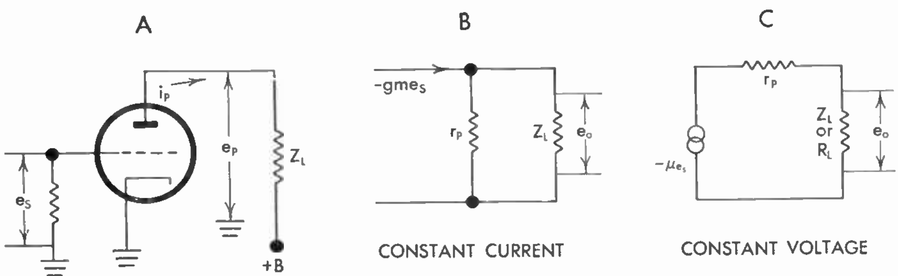


FIG. 328 Basic Vacuum Tube and Equivalents

ticularly helpful in calculating performance of a pentode because observation of pentode curves shows us the plate current is constant over a wide range of plate voltages. The bias curves on the $I_p - E_p$ characteristics of a pentode are nearly horizontal (linear portions) and it is difficult to accurately interpret changes in i_p. For triode $I_p - E_p$ characteristics the E_c curves are nearly vertical and the e_p parameter is more difficult to ascertain.

A third parameter of the vacuum tube is the μ, or amplification factor.

$$\mu = \frac{e_p}{e_s} = \frac{r_p i_p}{\dfrac{i_p}{g_m}} = g_m r_p$$

where I_p = constant

If we now substitute for g_m its equivalent μ/r_p, Eq. (2) becomes

$$e_o = -\frac{\mu e_s}{r_p}\left(\frac{r_p Z_L}{r_p + Z_L}\right)$$

$$e_o = -\mu e_s \frac{Z_L}{r_p + Z_L} \tag{3}$$

Now observation of Eq. (3) tells us a tube can also be considered a constant voltage source (Fig. 328C) with a generator voltage $-\mu e_s$ divided between internal plate resistance and output load. The fractional part of the generator voltage which appears across the load is $Z_L/r_p + Z_L$. A constant-voltage method can be employed conveniently to triodes because in the usual triode the plate voltage is essentially constant for a wide variation in plate current (bias curves nearly vertical).

168. *Video Amplifiers*

The equivalent circuit of a vacuum tube with a resistive load is shown in Fig. 328C. The voltage e_s represents the signal voltage applied to grid of a tube. Since μ is the tube amplification factor, the a-c voltage to be found in the plate circuit is μ times e_s or μe_s. However, the voltage μe_s is not the useful output voltage of the tube since a portion of this a-c variation is lost across r_p, the internal plate resistance of the tube. Thus the *stage gain* of the resistance-coupled amplifier is always less than the amplification factor of the tube. Simple application of Ohm's law permits us to determine the gain of the stage.

1. The current through the series combination of internal plate resistance, r_p, and total load resistance, R_L, is

$$i_p = \frac{\mu e_s}{r_p + R_L}$$

2. The voltage, e_o, across the output is the series current, i_p, times the resistance of the output load, R_L (constant voltage),

$$e_o = i_p R_L = \frac{\mu e_s R_L}{r_p + R_L}$$

3. The gain of the circuit is the ratio of output to input voltage, or

$$\text{Gain} = \frac{e_o}{e_s} = \frac{\dfrac{\mu e_s R_L}{r_p + R_L}}{e_s} = \frac{\mu R_L}{r_p + R_L}$$

4. Since it is convenient to express the gain of a pentode video stage in terms of mutual conductance instead of μ, and since pentodes are commonly used in video amplifiers, constant-current form is preferred. The μ can be removed from the gain formula by substituting its equivalent value $\mu = g_m r_p$. The formula for the constant-current generator then becomes

$$\text{Gain} = \frac{g_m r_p R_L}{r_p + R_L}$$

As an example, the gain of a resistance-coupled stage can be calculated as follows:

$$g_m = 9{,}000 \text{ micromhos with a screen voltage of } 150$$
$$r_p = 750{,}000 \text{ ohms}$$
$$R_L = 500{,}000 \text{ ohms}$$

$$\text{Gain} = \frac{g_m r_p R_L}{R_L + r_p}$$
$$= \frac{(9{,}000)(10^{-6})(0.75)(10^6)(0.5)(10^6)}{(0.5)(10^6) + (0.75)(10^6)}$$
$$= \frac{(9{,}000)(0.375)}{1.25} = 2{,}700$$

If the value of the load resistance is made less the gain will be decreased correspondingly. Thus, if the load resistance is 7,000 ohms the gain becomes

$$\text{Gain} = \frac{(9{,}000)(10^{-6})(0.75)(10^6)(0.7)(10^4)}{(0.7)(10^4) + (0.75)(10^6)}$$
$$= \frac{(90)(0.75)(0.7)}{0.757} = 62.4$$

Inspection of the formula shows that the tube plate resistance is effectively in shunt with the load resistor. Consequently, the gain of the stage for low values of plate load resistance is determined almost entirely by the value of the load resistance. If values of plate load resistance, R_L, which are low in comparison with the plate resistance, r_p, are used, the gain formula becomes

$$\text{Gain} = g_m R_L$$

This point is demonstrated by substituting the value of plate load resistance assumed in the previous example in the new formula and comparing the results.

$$\text{Gain} = g_m R_L$$
$$= (9{,}000)(10^{-6})(7{,}000) = 63$$

The reason for this is, of course, that the connection of a high resistance in parallel with a low resistance produces a resultant which differs only slightly from the value of the lower resistance.

It is evident if we use a low value of plate load resistance, as we do in a video amplifier, a reasonable gain requires the use of a tube with a high mutual conductance. It is also evident that greater gain is obtained with the highest permissible value of plate load resistance. Since the value of the plate resistance is inversely proportional to the total shunt capacity of the stage, a tube with low interelectrode capacities permits a higher value of load resistance and, therefore, a greater stage gain.

TOTAL LOAD RESISTANCE

Thus far we have considered R_L to be a single resistor. Actually, however, the value Z_L represents the total effective load impedance presented to the tube. As shown in Fig. 329, the effective value of plate load resistance at the middle range of frequencies consists of the plate load resistor, R_L, effectively in parallel with R_g, the grid-coupling resistor of the succeeding stage, if we consider the reactance of C_c to be negligible at these frequencies. However, as in the case of the tube's internal plate resistance, the resistance of the grid resistor may be neglected if it is large compared to that of the plate load resistor. With this qualification, the gain of the stage is $g_m R_L$ where R_L is the value of the plate load resistor only.

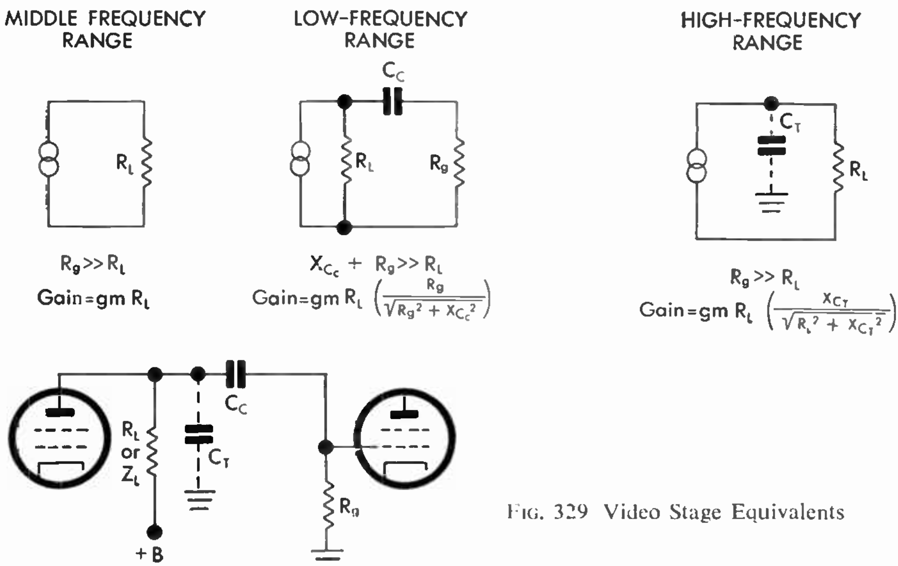


FIG. 329 Video Stage Equivalents

As the frequency is decreased, the reactance of the coupling capacitor, C_c, increases and a portion of the signal is filtered across the coupling capacitor, the percentage of loss increasing as the frequency decreases, gain of a resistance-coupled amplifier measured, of course, from grid to grid. In the case of the middle range of frequencies the signal variation at the plate appears essentially in its entirety at the grid of the succeeding tube, as explained previously. However, at low frequencies the coupling capacitor and grid resistor form a voltage-dividing network across the output of the tube. Thus the voltage at the plate of the first tube is still $\mu e_s\, g_m\, R_L$, but the voltage at the grid is dependent on the vector division of voltage by the coupling capacitor and grid resistor, R_g, the voltage across the grid resistor representing the useful output. The gain of the stage at low frequencies therefore becomes

$$\text{Gain} = g_m R_L \left(\frac{R_g}{\sqrt{R_g^2 + X_{C_c}^2}}\right)$$

The latter term of the equation represents the ratio of grid resistance to series impedance which is actually the fractional division in signal voltage resulting from the loss across the coupling capacitor. Thus, if it were necessary to have linear response down to 10 cycles, the numerical reactance of the coupling capacitor at 10 cycles must be 1/10 the resistance of the grid resistor (degeneration less than 0.5 per cent). In previous examples a grid-resistance value of 500,000 ohms was chosen. Using this value

$$X_{C_c} = \frac{R_g}{10} = \frac{500{,}000}{10} = 50{,}000 \text{ ohms}$$

$$C_c = \frac{1}{2\pi f X_{C_c}} = \frac{1}{(6.28)(10)(50{,}000)} = 0.32\ \mu f$$

LOW-FREQUENCY COMPENSATION

It is apparent that the low frequencies can be passed without serious loss by the interstage coupling system only if the RC combination is sufficiently large. A large time constant (RC), however, encourages relaxation oscillations and the larger physical dimensions of the parts increase stray capacities to ground, causing high-frequency degeneration. The proper time constant for most applications may not exceed 0.05 second. In the last example, the time constant, t, would be

$$t = RC = (500{,}000)(0.32)(10^{-6}) = 0.16 \text{ second}$$

Therefore, to prevent low-frequency degeneration and still not exceed the 0.05-second time constant another means of compensation is employed.

A low-frequency compensating circuit (R_fC_f) is shown in Fig. 330. In this circuit the time constant of R_gC_c is assumed to be 0.05 second and is equal to

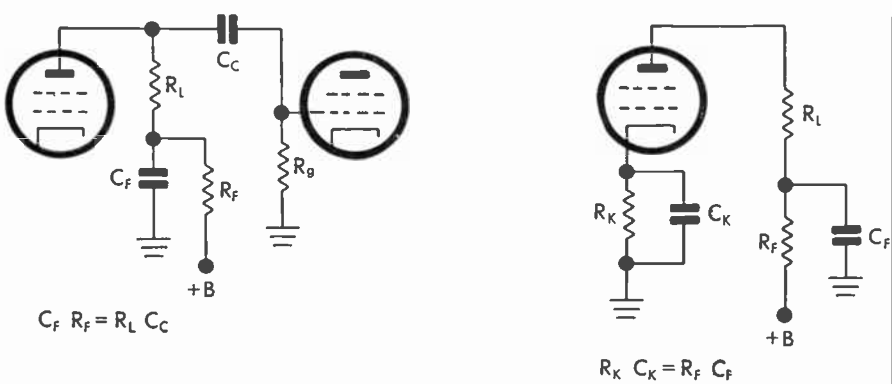


FIG. 330 Low-Frequency Compensation

the time constant of R_fC_f. The value of R_f is ten times the reactance of C_f at the lowest frequency to be passed. Thus, the R_fC_f combination is not only a low-frequency compensator but also a decoupling network which minimizes feed-back to the stage through the plate supply. At the middle and high ranges of frequency the reactance of C_f is negligible and the load presented to the tube is the value of R_L.

As the frequency decreases, the reactance of the capacitor C_f increases, effectively increasing the total plate load impedance and, in turn, increasing the gain of the stage at low frequencies. In order to equalize the response at the grid of the succeeding stage, therefore, the compensating network must increase the gain of the stage by the same percentage as that by which the coupling network decreases the gain. Let us assume that the voltage across R_g is 10 volts at 10 kilocycles, and drops to 7 volts at 10 cycles without compensation. When compensation is added, the plate signal at 10 cycles must be increased to a value which will develop 10 volts across R_g regardless of the degeneration caused by the coupling capacitor. Since, as stated previously, the gain of the tube is proportional to the plate load impedance, this required increase in voltage may be obtained by choosing compensating-network values which will increase the plate load impedance by a corresponding amount.

Without compensation the gain of our amplifier is, of course

$$\text{Gain} = g_m R_L \frac{R_g}{\sqrt{R_g^2 + X_{c_c}^2}}$$

With compensation the gain formula takes on a new factor and becomes:

$$\text{Gain} = g_m \left(\sqrt{R_L^2 + X_{c_f}^2}\right)\left(\frac{R_g}{\sqrt{R_g^2 + X_{c_c}^2}}\right)$$

To facilitate further calculations, the numerator and denominator are multiplied by R_L, or

$$\text{Gain} = g_m \sqrt{R_L^2 + X_{c_f}^2}\left(\frac{R_g}{\sqrt{R_g^2 + X_{c_c}^2}}\right)\left(\frac{R_L}{R_L}\right)$$

$$\text{Gain} = g_m R_L \left(\frac{\sqrt{R_L^2 + X_{c_f}^2}}{R_L}\right)\left(\frac{R_g}{\sqrt{R_g^2 + X_{c_c}^2}}\right)$$

Observation of the last two terms will show that each one is a ratio. The third term indicates the fractional division of voltage across the coupling capacitor, while the second term shows the ratio by which the gain of the amplifier is increased by the presence of the compensating capacitor. Since the gain of the amplifier at the middle range of frequencies is $g_m R_L$, the same gain must be maintained at the lows for a linear response. Thus it is evident the product of the last two terms must equal unity if linearity is to be retained.

$$\left(\frac{\sqrt{R_L^2 + X_{c_f}^2}}{R_L}\right)\left(\frac{R_g}{\sqrt{R_g^2 + X_{c_c}^2}}\right) = 1$$

$$\frac{R_g\sqrt{R_L^2 + X_{c_f}^2}}{R_L\sqrt{R_g^2 + X_{c_c}^2}} = 1$$

$$R_g^2(R_L^2 + X_{C_f}^2) = R_L^2(R_g^2 + X_{C_c}^2)$$
$$R_g^2R_L^2 + X_{C_f}^2R_g^2 = R_L^2R_g^2 + R_L^2X_{C_c}^2$$
$$X_{C_f}R_g = R_LX_{C_c}$$
$$\frac{R_g}{C_f} = \frac{R_L}{C_c}$$
$$R_gC_c = R_LC_f$$

Therefore, the low-frequency response can be held constant by maintaining the above relation. The resistance of the decoupling resistor, R_f, must be ten times the reactance of the coupling capacitor at the lowest frequency to be passed. However, there is a limit to the size of the resistor, because if it is too high it will drop the plate voltage to a low value. This, in turn, will lower the gain of the stage. Were it not for the limitations on the size of this resistor it would be possible to compensate the amplifier down to zero cycles.

COMPENSATION FOR CATHODE DEGENERATION

It is necessary also to compensate for the loss of gain caused by the increasing reactance of the cathode by-pass capacitor, C_k, at low frequencies. A similar type of compensation is employed as shown in Fig. 330.

The alternating potential developed across the cathode resistor and capacitor at low frequencies can be considered in series opposition to the applied grid signal. Thus, at low frequencies the effective decrease in grid signal arising from cathode degeneration must be compensated for by a capacitor and resistor combination in the plate circuit which increases the gain of the stage at low frequencies the proper amount to equalize response. As before, this is accomplished by having the time constant

$$C_kR_k = C_fR_f$$

However, in this case, definite relations must exist between the components, in consideration of the gain of the tube. For example, if at a certain frequency the signal drops off 1 volt at the grid, the added plate reactance or plate impedance presented by the compensating capacitor must not only restore the 1 volt to equalize the response, but it also may have to restore 10 or 15 volts, because all frequencies have been amplified by tube action. Thus, if we have a deficit of 1 volt at the grid and the gain of the stage is 15, we have a deficit of 15 volts in the plate circuit. Since approximately the same value of plate current flows in the cathode and plate circuits, the impedance added to the plate circuit must be equal to the gain of the tube times the cathode impedance at that frequency, or

$$R_f = (g_m R_L)R_k$$

Similarly

$$X_{C_f} = (g_m R_L)(X_{C_k})$$
$$\frac{1}{2\pi fC_f} = (g_m R_L)\left(\frac{1}{2\pi fC_k}\right)$$

$$\frac{2\pi f C_k}{2\pi f C_f} = g_m R_L = \frac{C_k}{C_f}$$

$$C_f = \frac{C_k}{g_m R_L}$$

It is important to note that in compensating for the grid-coupling and grid-resistor combination the time constant $R_L C_f$ was the important factor; in compensating for cathode degeneration, $R_f C_f$ combine. Actually, the compensating capacitor was all important in correcting the capacitive loss of C_c while the compensating resistor corrected the resistive degenerative-feedback loss.

It is not feasible to compensate for both losses in one compensating circuit because of difficulty in setting up proper relations. Fortunately, in receiver practice both types of low-frequency degeneration are not often necessary in the same video amplifier. When it is necessary to compensate for cathode degeneration it is not necessary to compensate for grid-coupling loss because of the high permissible value of the grid resistor (manufacturer's tube ratings permit high value of R_g when cathode bias is used). If cathode is grounded and external bias is used, permissible value of R_g is lower and compensation must be made for coupling combination. No cathode degeneration is present in this case.

HIGH-FREQUENCY RESPONSE

The response at high frequencies is limited by the distributed capacities to ground from the plate and grid circuits of the tubes. These distributed capacities consist of the input-and-output interelectrode capacities of the tubes, plus the capacities to ground of the wiring and parts. As the frequency increases or as the distributed capacity is increased, the reactance to ground is gradually reduced and more and more of the signal is shunted or by-passed. Thus, it is apparent that wiring and interelectrode capacities must be kept at a minimum to maintain high-frequency response.

The value of the plate load resistance also plays an important part in the high-frequency response. The lower the value of the plate resistance the higher the high-frequency range is extended. However, the lower the plate load resistance, the lower the gain of the stage since gain is equal to g_m times the load resistance. The plate impedance presented to the tube at high frequencies is the parallel combination of plate load resistor R_L and distributed shunt capacity C_t, as shown in Fig. 329. As in all vector combinations of a single resistor and capacitor, the frequency at which the capacitive reactance falls to a value equal to the resistance of the plate load resistor is the point where the response has fallen to 70.7 per cent of the middle-range amplitude.

The following examples demonstrate the various frequencies at which the response has fallen to 70.7 per cent for various operating conditions. In the first example the total shunt capacity C_t is assumed to be 60 micromicrofarads and the load resistance R_L, 100,000 ohms. The frequency at which the capacitive reactance equals the load resistance is

$$X_{c_t} = \frac{1}{2\pi f C_t} = R_L$$

$$100{,}000 = \frac{1}{(6.28)(f)(60)(10^{-12})}$$

and

$$f = 26{,}520 \text{ cycles}$$

Now if we lower the value of the load resistance to 10,000, we extend the high-frequency f_o limit at which the response is down 70.7 per cent, or

$$X_{c_t} = \frac{1}{2 f_o C_t}$$

$$10{,}000 = \frac{1}{(6.28)(f_o)(60)(10^{-12})}$$

and

$$f_o = 265{,}200 \text{ cycles}$$

Now if by proper choice of tubes and careful wiring we lower our distributed capacity to 30 micromicrofarads, the frequency limit is further extended.

$$X_{c_t} = \frac{1}{2 f_o C_t}$$

$$10{,}000 = \frac{1}{(6.28)(f_o)(30)(10^{-12})}$$

and

$$f_o = 530{,}400 \text{ cycles}$$

It is apparent, therefore, that the frequency at which the response is 30 per cent or 3 decibels down is dependent on circuit parameters. Likewise, if we properly choose these parameters, the 3-decibel downpoint can be set at a prescribed frequency. Let us call this frequency (f_o) our high-frequency limit.

As a practical example let us assume the distributed capacity of our stage is 20 micromicrofarads and we want the high-frequency limit to be 4 megacycles. What size load resistor is necessary?

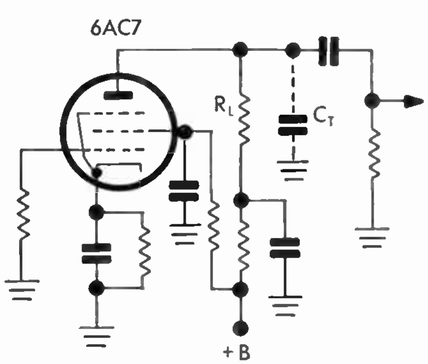


FIG. 331 Uncompensated Video Amplifier

$$R_L = \frac{1}{2\pi f_o C_t}$$

$$R_L = \frac{1}{(6.28)(4)(10^6)(20)(10^{-12})}$$

$$R_L = 1{,}990 \text{ ohms}$$

Again, using a 6AC7, Fig. 331, our gain over the middle range is

$$\text{Gain} = g_m R_L$$

$$\text{Gain} = 9{,}000\,(10^{-6})(1{,}990) = 17.9$$

At high frequencies our gain is less and at 4 megacycles it is specifically

$$\text{Gain} = g_m \frac{R_L X_{c_t}}{\sqrt{R_L^2 + X_{c_t}^2}}$$

$$\text{Gain} = 9{,}000 \left(10^{-6} \frac{(1{,}990)(1{,}990)}{\sqrt{(1{,}990)^2 + (1{,}990)^2}} \right) = 12.65$$

which proves the point that at the frequency at which $X_{c_t} = R_L$ gain is down 30 per cent from the middle-range gain. It is equally evident that gain of a video ampifier is very low and becomes lower still as the high-frequency limit is raised.

HIGH-FREQUENCY COMPENSATION

A high-frequency compensating circuit extends the linear frequency response of the stage, permitting a practical value of R_L and a reasonable gain. The simplest high-peaking system is the shunt peaking circuit of Fig. 332, in which

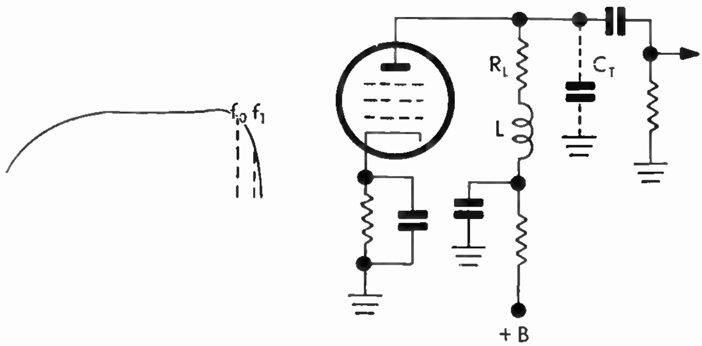


FIG. 332 Video Amplifier, Shunt Peaking

a small inductor, inserted in series with R_L, actually raises the plate load impedance of the stage at high frequencies. This increase in gain must be in proportion to loss normally introduced by C_t. Plate impedance of shunt-peaked stage is

$$Z_o = \frac{(X_{c_t})(\sqrt{R_L^2 + |X_L|^2})}{\sqrt{R_L^2 + (X_L - X_{c_t})^2}}$$

If the gain of the stage at f_o is to be the same as at the middle range Z_o must equal R_L at f_o, or

$$R_L = \frac{X_{c_t}\sqrt{R_L^2 + (X_L)^2}}{\sqrt{R_L^2 + (X_L - X_{c_t})^2}}$$

Likewise, we realize at f_o, X_{c_t} equals R_L and

$$R_L = \frac{R_L\sqrt{R_L^2 + (X_L)^2}}{\sqrt{R_L^2 + (X_L - R_L)^2}}$$

Solving

$$R_L = 2\,X_L$$
$$R_L = 2\,(2\pi f_o L)$$
$$L = \frac{R_L}{4\pi f_o}$$

The solution to this formula tells us the inductance necessary to compensate our stage up to a specific high-frequency limit. In our previous example size of shunt-peaking coil would be:

$$L = \frac{R_L}{4\pi f_o}$$
$$L = \frac{1{,}990}{(12.56)(4)(10^6)} = 39.0\ \mu\text{h}$$

If we now take the value of this inductor and find its resonant frequency with the distributed capacity, it is

$$f_r = \frac{1}{2\pi\sqrt{LC}}$$
$$f_r = \frac{1}{6.28\sqrt{39.5\,(10^{-12})(20)(10^{-12})}} = 5.6\text{ mc}$$

which is 1.4 times the high-frequency limit, f_o (in our example, 4 megacycles). Consequently, an alternative method of choosing L is to make it resonant with the distributed circuit capacity at a point 1.4 times higher than the chosen high-frequency limit.

In summation, shunt peaking added to our uncompensated video stage extends our linear response to the high-frequency limit (in uncompensated amplifier response is down 30 per cent at f_o). The resonant rise of the LC_t combination is responsible for this extension. Although LC_t is resonant beyond f_o the expected rise at the resonant frequency does not exist because of the poor Q of the circuit—large series resistance R_L.

A simple and practical design procedure for shunt-peaking a video amplifier if it is not possible to find C_t accurately is to put in a value of plate load resistor thought to be about right. Now measure the frequency response of the stage and find high-frequency limit at which point gain is down 30 per cent. Shunt capacity may now be calculated and correct value of load resistor inserted.

Now the peaking coil may also be inserted and the load resistor temporarily shorted. Set the input frequency to 1.4 f_o and adjust peaking coil for maximum output. After unshorting the load resistor the stage functions as a properly shunt-peaked amplifier. In applying above procedure it is important to keep signal generator output constant or use frequency-versus-gain data to find the 70 per cent point.

SERIES PEAKING

Series peaking, Fig. 333, permits a 50 per cent increase in gain over shunt peaking. The series-peaking coil is not the direct cause of this increase. Rather, the presence of the series coil permits use of a higher value load resistor per given bandwidth, which gives us our higher gain ($g_m R_L$). The series peaking system is actually a low-pass filter which conveys uniformly all frequencies

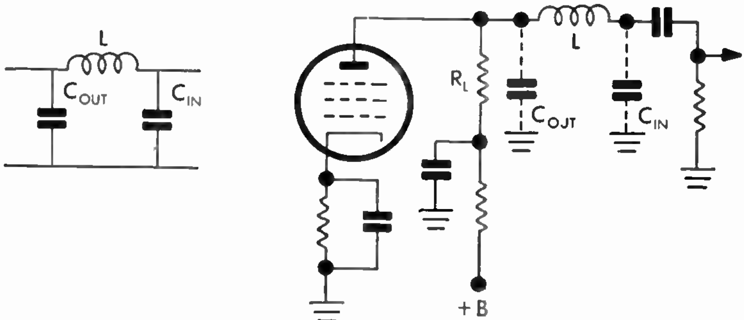


FIG. 333 Video Amplifier, Series Peaking

up to a finite high-frequency limit. Effectively, this series-peaking coil isolates the input and output capacities and reduces the extent the input capacity of the succeeding stage affects high-frequency response. The basic series peaking formulas are

$$R_L = \frac{1.5}{2\pi f_o C_t}$$

$$L = \tfrac{2}{3}\, C_t R_L^2$$

$$\frac{C_{\text{in}}}{C_{\text{out}}} = 2$$

The derivation of these relations is unduly complex and will not be treated. Important aspects are that R_L per given stage can be increased 50 per cent and the gain along with it. The peaking coil and C_{out} resonate at 1.4 f_o; peaking coil and C_{in}, at f_o. As in the case of shunt peaking for which L resonates with C_t at 1.4 f_o, L and C_{out} do the same, keeping the response flat to 4 megacycles.

While it first appears as though the series peaking coil would impede the transfer of high frequencies because of its rising reactance, remember that C_{in} and L form a series resonant circuit at f_o. The series current is higher as the frequency nears resonance and, consequently, keeps the output across C_{in} essentially constant despite the decreasing reactance of C_{out} and C_{in}. Above resonance (high-frequency limit f_o) the response drops off because of the decreasing reactance of C_{in} and C_{out} as well as the now decreasing series current.

One disadvantage of series peaking, as compared to shunt peaking and its simplicity, is the critical juggling of circuit capacities necessary to set up the ratio $C_{in}/C_{out} = 2$. This must be done by proper choice of tubes and grouping of circuit components on one side of L. Occasionally an actual capacitor is added to set up the relation. Series-peaking coils must often be loaded with shunt resistors to prevent overpeaking, which can be as harmful as underpeaking. When overpeaking exists the response curve has a resonant hump in the high-frequency region which introduces added phase shift.

The input capacity of a vacuum tube is not the capacity from grid to cathode alone but added to it grid-to-plate capacity multiplied by a factor which contains the stage-gain factor. This expression is

$$C_{in} = C_{gk} + C_{gp}\ (1 + \text{Gain})$$

In the case of a pentode video-amplifier stage both the stage gain and grid-to-plate capacity are low and the input capacity for most practical applications can be considered to be approximately equal to the grid-to-cathode capacity. When a triode video-amplifier is designed the true input capacity must be ascertained by substitution in the complete expression. Another loading condition is introduced by so-called "Miller effect" at the higher video frequencies reflecting a resistive component across the grid input. This resistive component is a function of the plate circuit reactance. Again, use of a pentode minimizes the resistive loading.

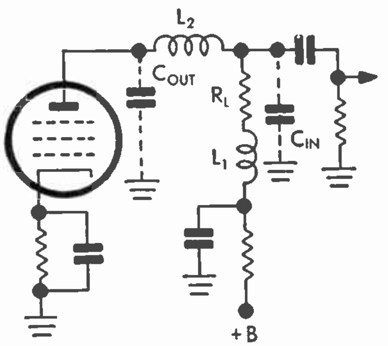


FIG. 334 Series-Shunt Peaking

It is important to realize that in any stage or stages (video or i-f) which have ajustable gain control facilities both the input capacity and resistive component contributed by "Miller effect" vary with gain setting. Thus, regulating the stage gain affects bandwidth, tuning, and response.

SERIES-SHUNT PEAKING

As might be expected, a further improvement in gain (80 per cent higher than shunt alone) is obtained with a combination of both types of peaking (Fig. 334); design formulas are

$$R_L = \frac{1.8}{2\pi f_o C_T}$$

$$L_1 = 0.12\ C_T R_L^2$$

$$L_2 = 0.52\ C_T R_L^2$$

$$C_{IN}/C_{OUT} = 2$$

169. *Phase Response and Time Delay*

An understanding of phase response and phase distortion with relation to video amplifiers hinges on an understanding of three factors—phase angle, time delay, and nonuniform time delay. Phase angle is the angular degrees through which a specific voltage or current leads or lags another. Time delay is the delay or lead in terms of time resulting from a certain angular phase shift. Nonlinear time delay is the difference between the delay times of two voltages of differing frequencies, which constitutes phase distortion. If all frequencies of a certain compound wave are delayed the same amount, no phase distortion is present. If the differing frequencies which make up the wave are delayed varying amounts, phase distortion and consequent distortion of the compound wave is prevalent.

Phase angle, using an interstage coupling *RC* as an example, is the angular relation between grid voltage and plate voltage of preceding stage (Fig. 335). As shown in the vector (which assumes the signal current i_s as a

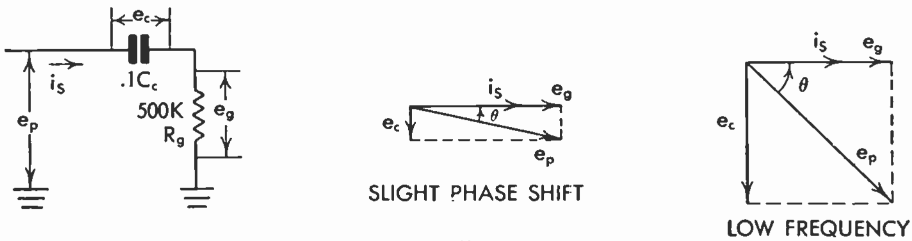


FIG. 335 Interstage Coupling and Vector Relations

zero point), the voltage e_g across the grid resistor is in phase with the signal current i_s. The voltage across the capacitor e_c lags i_s by 90 degrees. The location of the plate voltage e_p on the vector depends on the relative amplitudes of the e_c and e_g voltages (drawings B and C), which in turn are dependent on frequency and reactance-resistance ratio. At low frequencies it is evident that the phase angle is much greater because of the rising reactance of C_c and, therefore, grid voltage leads plate voltage by a greater angle.

The phase angle is equal to

$$\tan\theta = \frac{e_c}{e_g} = \frac{i_s x_c}{i_s R_g} = \frac{x_c}{R_g}$$

and

$$\theta = \tan^{-1}\frac{x_c}{R_g} = \tan^{-1}\frac{1}{2\pi f C_c R_g}$$

The phase angles for 100~ and 10-kilocycle frequencies, using schematic of Fig. 335 and the tangent tables in the appendix, are

$$\underset{(10\text{ kc.})}{\theta} = \tan^{-1}\frac{1}{3{,}140} \qquad\qquad \underset{(100\sim)}{\theta} = \tan^{-1}\frac{1}{31.4}$$

$$\underset{(10\text{ kc.})}{\theta} = 1\text{ min} \qquad\qquad \underset{(100\sim)}{\theta} = 1\text{ deg }49\text{ min}$$

The time delay is obtained by determining the actual time involved in the conveyance of that many degrees of the applied frequency. The period of a wave tells us how many seconds are required to form one sine wave of a certain frequency—one 360-degree period. If the wave is delayed a certain number of degrees, the time it is delayed is the fractional part of a total cycle it is delayed times the period of the wave, or

$$\text{Time delay} = \frac{\theta}{360^\circ} \times \text{period}$$

$$\text{Period} = \frac{1}{\text{frequency}}$$

In our example the time delays are

$$t_{10\,\text{kc}} = \frac{1'}{360^\circ} \times \frac{1}{10{,}000}$$

$$t_{10\,\text{kc}} = \frac{1'}{21{,}600'} \times \frac{1}{10{,}000}$$

$$t_{10\,\text{kc}} = 0.00046\ \mu\text{sec lead}$$

$$t_{100\sim} = \frac{1^\circ\,49'}{360^\circ} \times \frac{1}{100}$$

$$t_{100\sim} = \frac{109'}{21{,}600'} \times \frac{1}{100}$$

$$t_{100\sim} = 46\ \mu\text{sec lead}$$

The nonlinear time delay is the difference between the time delays of two frequencies compared, or

$$t = t_{100\sim} - t_{10\,\text{kc}}$$
$$t = 46 - 0.00046$$

If the time delay of both frequencies through the system were the same, there would be no phase distortion because they would arrive at the picture tube in exactly the same relation they were developed at the camera. However, if nonlinear time delay exists, one of the frequencies will be displaced with respect to the other and will arrive ahead of or behind its initial time relation to the other frequency. Consequently, the picture will not be reassembled in precisely the same manner in which it was released at the camera, producing picture distortion.

In the preceding example the 100~ as well as other low-frequency components would lead the middle-range components and low-frequency phase distortion would exist. Inasmuch as the beam moves down the screen relatively slowly this amount of nonlinear time delay would not produce excessive picture distortion. If there is appreciable phase distortion present at lows, there is a gradual change in shading top to bottom of the reproduced image and severe distortion causes tails to follow behind large letters or objects of the image.

In effect low-frequency phase distortion differentiates any long, continuous, uniform-light level causing sharp rises into black or saturated white spectrum and then a gradual return to even background.

HIGH-FREQUENCY PHASE SHIFT

High-frequency components of the television signal represent the light distribution along the individual lines of the picture. If there is any high-frequency phase shift the elements along the line will not be assembled in proper sequence. Inasmuch as the motion of the beam along a single line is very rapid, only a small amount of high-frequency phase distortion will cause a noticeable effect on the picture. The high-frequency components of the picture signal are delayed by the distributed circuit capacity which retards the build-up of a high-frequency voltage across the output circuit (Fig. 336).

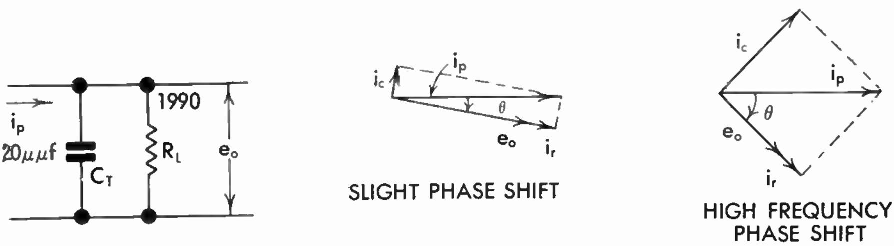


FIG. 336 High-Frequency Phase-Shift Vectors

As shown in the vectors the output voltage lags the plate current of the stage in accordance with the ratio between R_L and X_{c_t}—in the case of a shunt R and C, the larger the X_c the smaller the phase shift (an increase in X_c means a smaller i_c). The phase shift angle is

$$\theta = \tan^{-1}\frac{i_c}{i_r} = \tan^{-1}\frac{\dfrac{E}{X_{c_t}}}{\dfrac{E}{R_L}} = \tan^{-1}\frac{R_L}{X_{c_t}}$$

It is a fact with an uncompensated video amplifier that

$$R_L = \frac{1}{2\pi f_o C_t}$$

Substituting

$$\theta = \tan^{-1}\frac{R_L}{X_{c_t}} = \tan^{-1}\frac{1}{2\pi f_o C_T X_{c_t}}$$

$$\theta = \tan^{-1}\frac{2\pi f_1 C_t}{2\pi f_o C_t} = \tan^{-1}\frac{f_1}{f_o}$$

This latter expression, in the case of the uncompensated video amplifier only, gives us phase shift in terms of any frequency we might want to check with respect to our high-frequency limit. Using 4 megacycles again as f_o, let us check phase shift and time delay at 4 megacycles; ½ megacycle; and 10 kilocycles:

$$\theta = \tan^{-1}\frac{f_1}{f_o} \qquad \theta = \tan^{-1}\frac{f_1}{f_o} \qquad \theta = \tan^{-1}\frac{f_1}{f_o}$$

$$\theta = \tan^{-1}\frac{4(10^6)}{4(10^6)} \qquad \theta = \tan^{-1}\frac{0.5(10^6)}{4(10^6)} \qquad \theta = \tan^{-1}\frac{0.01(10^6)}{4(10^6)}$$

$$\theta = \tan^{-1} 1 \qquad \theta = \tan^{-1} 0.125 \qquad \theta = \tan^{-1} 0.0025$$

$$\theta_{4\text{ mc}} = 45\text{ deg} \qquad \theta_{1/2\text{ mc}} = 7\text{ deg } 7\tfrac{1}{2}\text{ min} \qquad \theta_{10\text{ kc}} = 8\tfrac{4}{7}\text{ min}$$

Time delay equals:

$$t_{4\text{ mc}} = \frac{45^\circ}{360^\circ} \times \frac{1}{4 \times 10^6} = \frac{1}{8} \times \frac{10^{-6}}{4} = 0.0312\ \mu\text{sec}$$

$$t_{1/2\text{ mc}} = \frac{427.5'}{21{,}600'} \times \frac{1}{0.5 \times 10^6} = 0.0396\ \mu\text{sec}$$

$$t_{10\text{ kc}} = \frac{8.57'}{21{,}600'} \times \frac{1}{0.01 \times 10^6} = 0.0396\ \mu\text{sec}$$

Note particularly that time delay between 10 kc and ½ mc is constant and does not become nonlinear until frequency is high. Nonlinear time delay is:

$$t = t_1 - t_2$$

$$t = 0.0396 - 0.0312 = 0.0084\ \mu\text{sec}$$

It has been customary practice to compare nonlinear time delay with the period of the high-frequency limit f_o. Thus, percentage phase distortion would be the period of our high-frequency limit (in our example 4 megacycles) divided by the nonlinear time delay, or

$$\% = \frac{\dfrac{1}{4(10^6)}}{0.0084(10^{-6})}(100) = 400\ (0.0084) = 3.36\%$$

In summation, remember that both poor frequency and poor phase response leave their effects on the image. Actually, it is the nonlinear phase shift that does the most harm to the picture. Fortunately, in receiver application we do not often consider phase shift because it is a fact that compensation for frequency response also makes the phase characteristics satisfactory. Since it is simpler to check frequency response than phase, it is the angle of attack used in receivers.

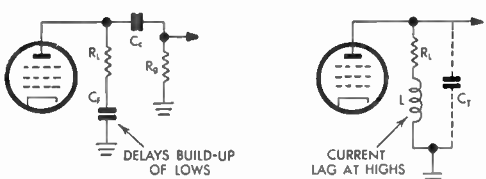


FIG. 337 Phase Correction with Frequency-Compensating Systems

Improvement of phase response by frequency-compensating circuits (Fig. 337) is not difficult to understand. In the case of the low-frequency compensating system the capacitive plate load at low frequencies delays the build-up of the lows in the plate circuit, correcting to an appreciable extent the low-frequency lead at the grid of the succeeding stage. Likewise, at high frequencies the presence of a compensating inductor improves high-frequency phase response because it introduces inductive lead into the plate load correcting for the capacitive delay at high frequencies.

PHASE SHIFT IN THE COMPENSATED STAGE

The development of the high-frequency phase-shift formula for a simple shunt-peaked stage follows. Effective plate impedance of the stage, Fig. 332, is:

$$Z_L = \frac{(R + jx_L)(-jx_c)}{R + j(x_L - x_c)}$$

$$Z_L = \frac{(-jRx_c + x_Lx_c)(R - j)(x_L - x_c)}{R^2 + (x_L - x_c)^2}$$

Isolation of real and reactive components in the form $a + jb$ will give us a series equivalent (right-angle vector which can be used to find expression for phase angle).

$$Z_L = \frac{Rx_c^2}{R^2 + (x_L - x_c)^2} + j\,\frac{x_Lx_c^2 - x_cx_L^2 - R^2x_c}{R^2 + (x_L - x_c)^2}$$

$$\text{and } \tan\theta = \frac{x_Lx_c - x_L^2 - R^2}{Rx_c}$$

Again with shunt peaking

$$R = \frac{1}{2\pi f_oC_t} \qquad x_L = 2\pi f_1L \qquad \text{and} \quad x_c = \frac{1}{2\pi f_1C_t}$$

Substitution in above equation results in

$$\tan\theta = \tfrac{1}{4}\left[\left(\frac{f_1}{f_o}\right)^3 + 2\left(\frac{f_1}{f_o}\right)\right]$$

$$\theta^\circ = \tan^{-1}\tfrac{1}{4}\left[\left(\frac{f_1}{f_o}\right)^3 + 2\left(\frac{f_1}{f_o}\right)\right]$$

If the same procedure is used to calculate nonlinear time delay for the shunt-peaked stage it will only be 2.3 per cent of the period of f_o, an improvement over the uncompensated stage.

A similar angle of attack can be used for the more involved peaking systems. The mathematics, however, is unduly complex and will not be treated. Series peaking offers the best phase response—only 1.1 per cent. Series-shunt peaking has the highest gain and a phase response of 1.5 per cent, just slightly higher than series alone.

170. *Cathode Follower*

The cathode follower (Fig. 338) is often used as the output tube of a video amplifier because it can transform a high-impedance signal to a low impedance without sacrificing frequency response. No physical transformer can pass the wide frequency spectrum contained in the television signal. Consequently, when it is necessary to match a low impedance, a vacuum tube connected as a cathode follower serves as an impedance transformer. This it does without frequency discrimination.

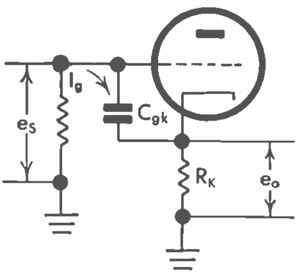


FIG. 338 Cathode-Follower Voltages

The gain of the cathode follower is always less than unity and, in most cases, the tube draws a substantial current from the supply. This latter characteristic is obvious when we consider that an appreciable a-c variation is necessary to produce even a small voltage variation across the small value cathode-resistor.

The basic formula of a vacuum-tube circuit, as developed at the opening of the chapter, is

$$i_p = g_m\, e_s + \frac{e_p}{r_p}$$

This basic form also applies to the cathode follower with the exception that e_s of the formula is equivalent to $e_s - e_o$ (Fig. 338), and $e_p = -e_o$. Our equation now takes the form

$$i_p = g_m\,(e_s - e_o) - \frac{e_o}{r_p}$$

$$i_p = g_m\, e_s - \frac{e_o}{\dfrac{1}{g_m}} - \frac{e_o}{r_p}$$

The latter equation represents the constant current form of the cathode follower, Fig. 339A showing the current available in the load is the total current minus the current required by the internal resistance and feedback. Output voltage, of course, is

$$e_o = i_p R_k$$

$$e_o = \left(g_m\, e_s - \frac{e_o}{\dfrac{1}{g_m}} - \frac{e_o}{r_p}\right) R_k$$

Substituting μ/r_p for g_m and solving for e_o will produce the constant-voltage version, or

$$e_o = \frac{\mu e_s R_k}{r_p + \mu R_k + R_k}$$

$$e_o = \frac{\mu e_s R_k}{r_p + R_k\,(\mu + 1)}$$

Dividing through by $\mu + 1$

$$e_o = \frac{\dfrac{\mu}{\mu + 1}\, e_s R_k}{\dfrac{r_p}{\mu + 1} + R_k}$$

Since

$$\frac{r_p}{\mu + 1} = \frac{r_p \dfrac{1}{g_m}}{r_p + \dfrac{1}{g_m}}$$

$$e_o = \frac{\mu}{\mu + 1}\, e_s \left(\frac{R_k}{\dfrac{r_p \times \dfrac{1}{g_m}}{r_p + \dfrac{1}{g_m}} + R_k} \right)$$

which tells the generator voltage of the cathode follower is $\dfrac{\mu}{\mu + 1}\, e_s$ and the output voltage is that fractional part of the generator voltage which appears across R_K of the total series impedance (Fig. 339B).

When a high-μ, high-plate-resistance tube is used, particularly with a high-μ pentode (screen must be bypassed to cathode), the cathode-follower for-

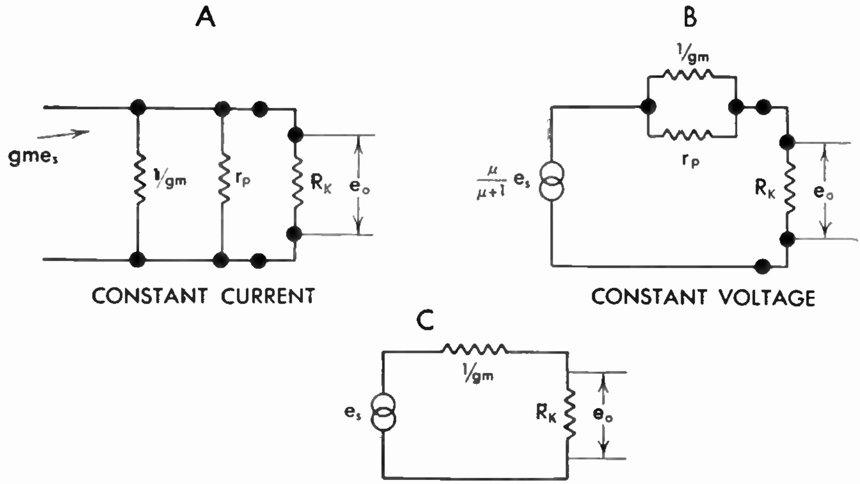


FIG. 339 Cathode-Follower Equivalents

mula can be greatly simplified. Now $\mu/(\mu + 1)$ is practically equal to one and the paralleled r_p and $1/g_m$ are simply $1/g_m$. Our equation becomes simply (Fig. 339C)

$$e_o = \frac{e_s R_k}{\frac{1}{g_m} + R_k}$$

$$\text{Gain} = \frac{e_o}{e_s} = \frac{R_k}{\frac{1}{g_m} + R_k}$$

From this expression it is evident that the gain is always less than unity and that the output impedance of the cathode follower is low because the factor $1/g_m$ is never more than a few hundred ohms. Output impedance, of course, becomes still lower if the value of R_k is reduced toward $1/g_m$. If R_k is made

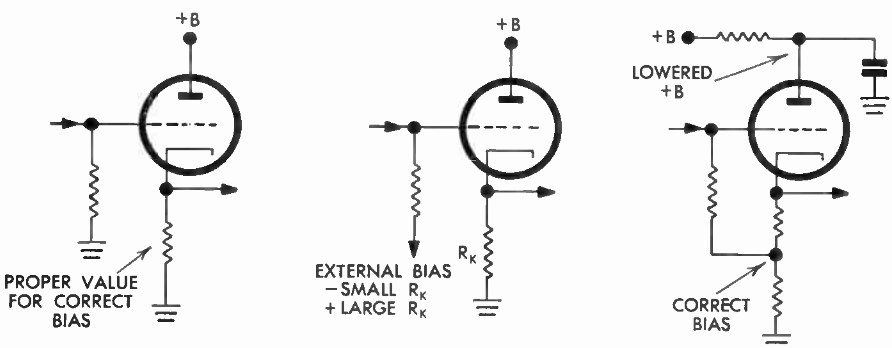


FIG. 340 Biasing the Cathode-Follower

higher than $1/g_m$ a gain of unity is approached. For television application it is generally necessary to have R_k under 200 ohms, in which case the gain of the cathode follower is considerably less than unity.

Two factors determine the size of R_k, the impedance match required and the proper bias on the cathode follower. If a specific value of R_k is necessary for good match, external grid bias can be used to prevent biasing the tube too near cutoff (high value R_k) or saturation (low value R_k), Fig. 340. Cathode load can also be tapped, or plate voltage reduced to a safer value if plate circuit is properly decoupled.

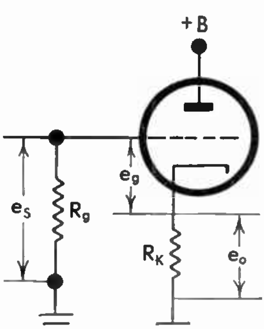


FIG. 341 Reduction in Input Capacity

REDUCTION OF INPUT CAPACITY

The cathode-follower connection also effectively reduces the input capacity of any tube so connected. This particularly applies to a pentode which has a low grid-plate capacity. The reduction in effective input capacity is demonstrated in the simple schematic of Fig. 341, showing the input capacity as C_{gk}. Reactive grid-signal current is

$$i_g = \frac{e_s - e_o}{\dfrac{1}{\omega C_{gk}}}$$

From our previous derivation

$$e_o = e_s \frac{R_K}{\dfrac{1}{g_m} + R_k}$$

and by substitution

$$i_g = \frac{e_s - e_s \left(\dfrac{R_k}{\dfrac{1}{g_m} + R_k} \right)}{\dfrac{1}{\omega C_{gk}}}$$

$$i_g = e_s \omega C_{gk} \left(1 - \frac{R_k}{\dfrac{1}{g_m} + R_k} \right)$$

$$i_g = e_s \omega C_{gk} \left(\frac{\dfrac{1}{g_m}}{\dfrac{1}{g_m} + R_k} \right)$$

in which the last term indicates the effective reduction of C_{gk}. Consequently, whenever a tube is connected as a cathode follower the effect of its input capacity on the preceding stage is reduced by a factor

$$\frac{\dfrac{1}{g_m}}{\dfrac{1}{g_m} + R_k}$$

171. *Time Constant*

If it were possible to apply a voltage E across a capacitor and there were no opposition to the displacement current necessary to charge the capacitor, the voltage E would appear instantly on the capacitor. However, there is resistance always present, if it is only the resistance of the conductor or internal resistance of the voltage source. Consequently, there is a finite time involved in the displacement of necessary electrons to establish the charge E—this time varies with the number of electrons which must be moved (capacity) and the opposition to the transfer (resistance). With proper choice of R and C this time can be made significant or insignificant in accordance with the circuit function and signal frequency.

In any *RC* combination (Fig. 342) the rise of the capacitor charge and fall of the charging current is not linear because as the charge builds up on *C* it is in opposition to the applied *E* and, consequently, the charging current falls, which in turn reduces the rate at which the capacitor charge builds up. Actually, the build-up of voltage and decrease of charging current is exponential—capacitor charge approaching applied *E* as a limit and charge current approaching zero as a limit. When the capacitor is charged to *E* no further current flows.

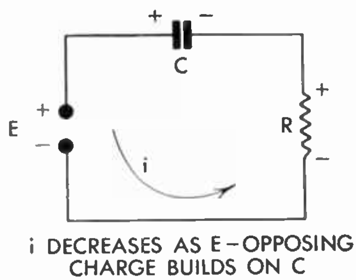


FIG. 342 Resistor-Capacitor Combination

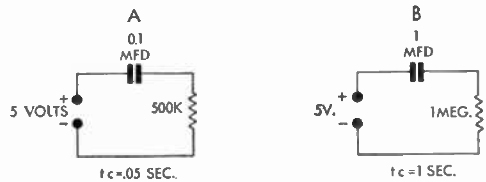


FIG. 343 *RC* Time Constants

The exponential equation for instantaneous resistor voltage e_r is

$e_r = E\varepsilon^{-t/T}$
e_r = resistor voltage at time t
E = applied voltage at time zero
ε = 2.718 (base of natural logs)
t = interval voltage is applied
T = a constant called the *time constant* and is equal to *R* times *C*, which we said affected the time required to build up the capacitor charge

The capacitor voltage is also exponential and equals at any instant

$$e_c = E - e_r = E - E\varepsilon^{-t/T}$$

As an example of the manipulation of this formula, find the charge on *C* and voltage across *R* when $t = T$, or at the instant of time one *time constant* after the application of the voltage. Assume applied *E* is 100 volts.

$$e_c = 100 - 100\varepsilon^{-1}$$

$$e_c = 100 - \frac{100}{\varepsilon^1} = 100 - \frac{100}{2.718}$$

$$e_c = 100 - 36.7 = 63.3 \text{ v}$$

The voltage across the resistor at the same instant is 36.7 volts.

Now find the charge on the capacitor at the end of five *time constants*, or $t = 5T$.

$$e_c = 100 - 100\varepsilon^{-5}$$

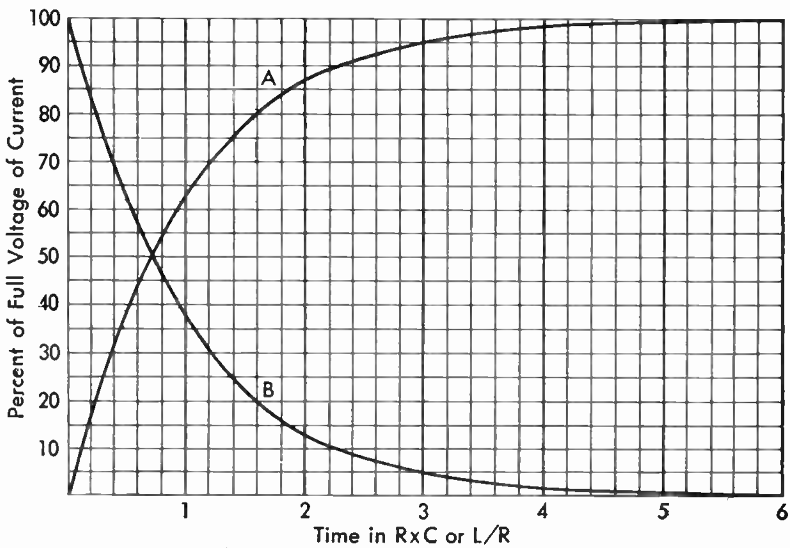
$$e_c = 100 - \frac{100}{\varepsilon^5} = 100 - 0.67 = 99.33$$

From the above examples it is established that with any RC combination and voltage applied, the capacitor will charge to 63.3 per cent of E in *one time constant* and almost to full value, 99.5 per cent, in *five time constants.* Consequently, if values of R and C are known the build-up of charge can be estimated quickly.

To demonstrate, the time constants of values shown in Fig. 343 are

$$T = R_{\text{ohms}}\,C_{\text{farads}}$$
$$T_A = 500{,}000\,(0.1)(10^{-6}) = 0.05 \text{ sec}$$
$$T_B = 1{,}000{,}000\,(1)(10^{-6}) = 1 \text{ sec}$$

Thus, in example A, Fig. 343 (with 5 volts applied), the capacitor will charge to 3.16 volts (63.3 per cent of 5) in 0.05 second; in example B, 1 second



Universal Exponential Curves for R-C and L-R Circuits

FIG. 344 Time-Constant Chart

would be required. To obtain almost full charge on C requires 0.25 second (5×0.05) in example A, 5 seconds in example B.

Very fast and essentially accurate calculations can be made with an RC chart (Fig. 344). In this chart curve A represents capacitor voltage on charge; curve B, on discharge. Curve B represents resistor voltage on charge and discharge of the capacitor. The vertical axis of the chart is calibrated in percentage voltage—percentage of applied E. Horizontal axis is calibrated in time constants or t/T (time over RC product).

To demonstrate application of chart, find charge on C and voltage across R at end of ½, 1, and 3 seconds for values of the example (Fig. 345). Time constant of RC is ½ second; therefore, at end of ½ second one time constant has elapsed and the charge on C is 63.3 per cent of 20 volts, and the voltage across R is 36.7 per cent of 20. One second elapsed time represents *two time constants* and voltage on C is 87 per cent; voltage across R is 13 per cent. Three seconds elapsed time is 6 RC, or *six time constants* putting full charge on C (20 volts) and no drop across R.

INDUCTOR-RESISTOR COMBINATION

It is also a fact that the rise of current in a coil is also exponential and the chart of Fig. 344 also applies, with the exception that curve A represents current instead of voltage.

In an RL combination the time constant equals L/R, the time constant decreasing with the size of R instead of increasing as with an RC combination. We studied, in connection with an RC combination, that the higher the resistance the slower the capacitor charged. In an RL combination the higher the resistance the faster the counter electromotive force of the coil is reduced and the quicker the current reaches its maximum (greater division of voltage across R as current begins to increase).

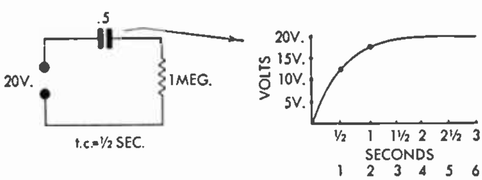


FIG. 345 Time Constant and Capacitor Charge

Upon first application of voltage to an RL combine, it appears in its entirety across L and gradually falls off (curve B, Fig. 344) as current rises (curve A). Voltage finally appears in its entirety across R. In an RC combine, voltage is first maximum across R and in time voltage shifts to maximum across C (full charge) and no voltage across R.

Discharge of voltage from capacitor and decay of current in a coil are also exponential (curve B, Fig. 344) and same procedure can be used to determine instantaneous values at any instant of time after start of decline.

172. *Time Constant and the Sine Wave*

The time constant is equally important when sine waves are applied to resistor, capacitor, and inductor combinations. In the case of a sine wave we cannot consider the voltage as being applied instantly. Instead, the applied voltage builds up gradually toward peak amplitude. The action of the combination on the amplitude of the sine wave is a factor of the sine-wave frequency and time constant of the network. When the time constant is long in comparison to the period of the sine wave applied to a coupling-capacitor grid-resistor combination (Fig. 346) the capacitor does not charge and discharge with the sine wave because the sine-wave variations are much faster than time required to charge or discharge C_c any appreciable amount. Consequently the a-c variations appear in their entirety across R_g and not across C_c. At low frequencies where the sine-wave period is longer than the time constant, the coupling capacitor will charge and discharge an appreciable amount during the sine-wave period, full variation no longer appearing across the grid resistor.

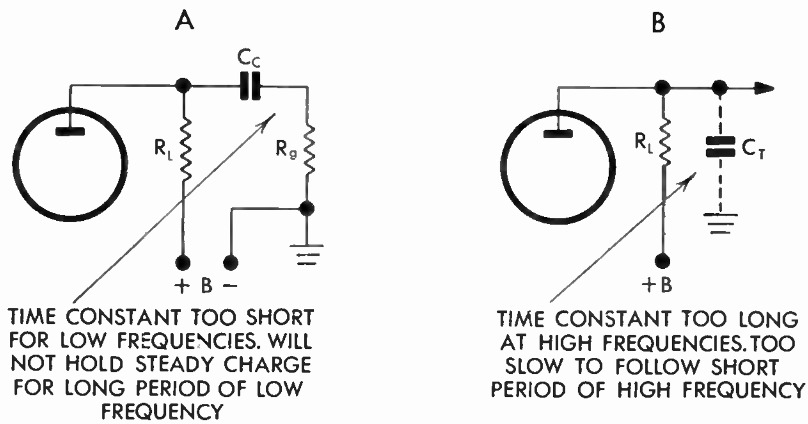


FIG. 346 Time Constant and Sine-Wave Period

From previous calculations we know that our gain is down 30 per cent for low frequency f_1 at which $X_{C_c} = R_g$.

$$f_1 = \frac{1}{2\pi X_{C_c} C_c}$$

and since $X_{C_c} = R_g$ at 30 per cent point

$$f_1 = \frac{1}{2\pi R_g C_c}$$

$$\frac{1}{p} = \frac{1}{2\pi R_g C_c} \qquad \text{because } \frac{1}{\text{period}} = f_1$$

$$R_g C_c = \frac{p}{2\pi}$$

Inasmuch as $t = R_g C_c$ 30 per cent down, time constant $= p/2\pi$, or $1/2\pi f_1$. In other words, when the time constant of $R_g C_c$ is equal to the period of a wave divided by 2π, response is down 30 per cent. To keep loss to less than 2 per cent, time constant at lowest frequency to be passed should be

$$t = R_g C_c = \frac{5}{2\pi f_1}$$

This is the condition satisfied when $R_g = 5X_{C_c}$ at lowest frequency to be passed. If $R_g = 10X_{C_c}$, loss is less than 0.5 per cent. Thus, if we make our time constant sufficiently long, no a-c variation appears across C_c—it charges to d-c plate-voltage level and remains there. Full a-c variation appears across R_g.

In some circuits the presence of capacity filters a desired variation. Such is the case with distributed capacity across the plate load (Fig. 346*B*). At high frequencies the period of the wave is short in comparison to the time constant of $R_L C_t$ and consequently the capacitor reduces the a-c variation because it does not have time to charge and discharge and, therefore, cannot follow the variation. Thus the high frequencies, because of their short periods, are filtered.

To reduce this filtering, C_t or R_L must be decreased in value, producing a shorter time constant which can follow the variations. Again a 30 per cent loss occurs at high frequency f_o where

$$\underset{(R_L C_t)}{\text{Time constant}} = \frac{1}{2\pi f_o}$$

At this frequency $R_L = \dfrac{1}{2\pi f_o C_t}$. To prevent more than a 2 per cent loss, the time constant should be less than the period of highest frequency wave to be passed.

$$\text{Time constant} = \frac{1}{10\pi f_o}$$

This condition is satisfied when $R_L = 1/5\ X_{c_t}$ at highest frequency. Degeneration is less than 0.5 per cent when $R_L = 1/10\ X_{c_t}$.

173. *RC Filtering*

Truthfully speaking, all our decoupling and so-called "bypass points" are filters and function in conjunction with some resistance or impedance to form a long time constant to filter the variation. Various supply voltage points are indicated in Fig. 347. At each point, to prevent degeneration or regeneration, it is necessary to charge the capacitor to some d-c voltage level which must be held constant in spite of the variations in the associated circuits. To prevent any variation at the supply voltage points it is necessary that the time constant of the *RC* combinations be long at the lowest frequency (has longest period), preventing charge and discharge of capacitor. Current displacement in and out of the capacitor branch is sufficient to hold the current constant through the resistive branch and, therefore, the voltage constant.

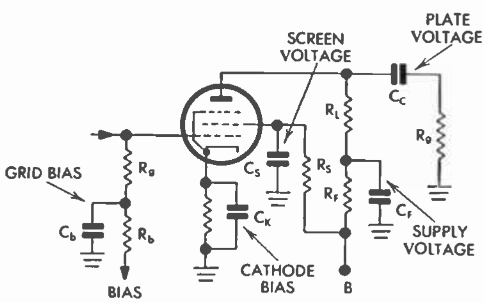


FIG. 347 Supply-Voltage Points *RC* Filtering

If the current is low it is possible to use an *RC* combination as a high-voltage filter. If the current drawn is small the resistive component is high, and a long time-constant *RC* can be obtained with a practical value of capacity. In a typical TV high-voltage supply (Fig. 348) the capacitor is charged quickly to peak value by the alternation of the applied sine wave which causes conduction of the rectifier. The capacitor is charged quickly because of the low resistance of the conducting rectifier (short time constant). Between

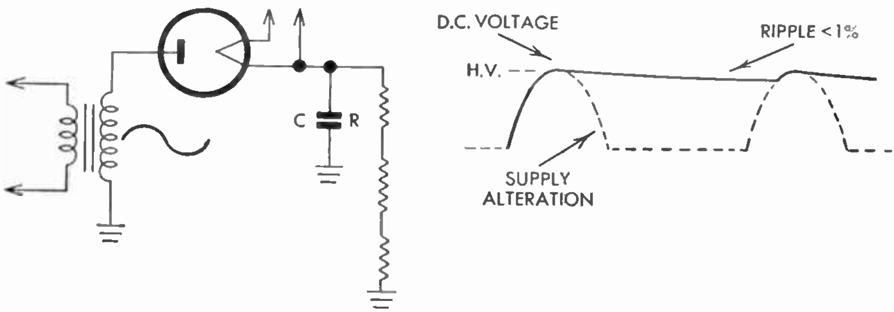


FIG. 348 High-Voltage *RC* Filtering

alternations the capacitor discharges in accordance with the current required. If current required is small and time-constant *RC* long, very little charge comes off the capacitor. The discharge of capacitor is limited to 1 per cent of peak voltage to prevent ripple interference. Discharge formula is

$$e_c = e_p \varepsilon^{-t/RC}$$

If discharge is limited to 1 per cent, formula becomes

$$e_p - 0.01e_p = e_p \varepsilon^{-t/RC}$$

$$\frac{e_p - 0.01e_p}{e_p} = \varepsilon^{-t/RC}$$

Inasmuch as the period between charges of the capacitor (60~ supply) is $t = 1/f = 1/60\sim = 0.0166$ second, the *RC* required is

$$0.99 = \varepsilon^{-0.0166/RC}$$

Solving for *RC*

$$\log_\varepsilon 0.99 = -\frac{0.0166}{RC} \log_\varepsilon \varepsilon$$

$$9.99 - 10 = -\frac{0.0166}{RC} \quad (\log_\varepsilon \varepsilon = 1)$$

$$-0.01\,RC = -0.0166$$

$$RC = 1.66 \text{ seconds}$$

Thus, to obtain a high-voltage potential which has less than a 1 per cent ripple the *RC* time constant should be 1.66 seconds (100 times period of wave). Choice of *R* and *C* is dependent on current required—*R* itself representing the actual load on the supply. It is evident that the lighter the load (high-value *R*) the smaller *C* can be for a given time constant.

If a high-frequency high-voltage supply is used the time-constant requirement for a given ripple per cent is much lower. For example, if the oscillator is on 100 kilocycles, time constant is as follows:

Period of 100-kilocycle wave is $1/10^5$ second or 10 microseconds

$$e_c = e_P \varepsilon^{-t/RC}$$
$$0.99 = \varepsilon^{-10/RC}$$
$$\log_\varepsilon \varepsilon\, 0.99 = -\frac{10}{RC} \log_\varepsilon \varepsilon$$
$$-0.01\, RC = 10$$
$$RC = 1{,}000\ \mu\text{sec}$$

If a double-section filter is used, smaller value capacitors can be used for the same ripple percentage. The first filter capacitor charges to peak value and its discharge is set by the two series of resistors. Remainder of the ripple is further filtered by the second capacitor and resistive load.

174. *Differentiation and Integration*

A complex wave such as a triangular, squared or sawtooth wave contains a fundamental and a group of harmonically related frequencies of various amplitudes. The shape of the complex wave is a function of the number of harmonic waves present and their relative amplitudes. For example, in drawing A of Fig. 348A it is seen that the addition of a fundamental sine wave and a third harmonic wave of a certain amplitude produces an essentially triangular wave.

Development of a squared wave is depicted in drawing B and is a result of the algebraic addition of a fundamental, a third, and a fifth harmonic wave. If a squared wave of straight vertical sides and level flat-top is to be formed, many harmonic waves need be present, in fact, an absolutely straight side and perfectly level flat-top would require an infinite number of harmonic waves. A usable and apparently squared wave is produced when odd harmonics up to the fifteenth are present.

A rectangular pulse also results from the addition of a fundamental and a group of harmonic waves, drawing C. A very crude rectangular wave results from the addition of a fundamental and second harmonic wave and as more harmonics are added in correct amplitude the wave has straighter sides and a more level flat-top. For example, a rectangular wave with a fundamental frequency of 1,000 cycles and a duration of 1 microsecond could very readily consist of a fundamental frequency and a few thousand harmonic waves. Actually the fundamental frequency and low-order harmonics, because they represent such a small part of the pulse area, could very well be entirely absent without changing the appearance of the short-duration pulse.

If the above-mentioned pulse is applied to a circuit which removes or compresses the amplitude of some of the harmonic components of the pulse, the actual shape of the pulse can be altered. If the high-order harmonics are removed or attenuated the pulse is integrated and its steep sides and level top are lost. To do so requires a low pass filter, and, of course, the longer the time constant the more effective is the removal of the higher-order harmonics. When the low-order harmonics are removed the flat top of the pulse is lost while the steep sides are retained and the wave is differentiated. To do so

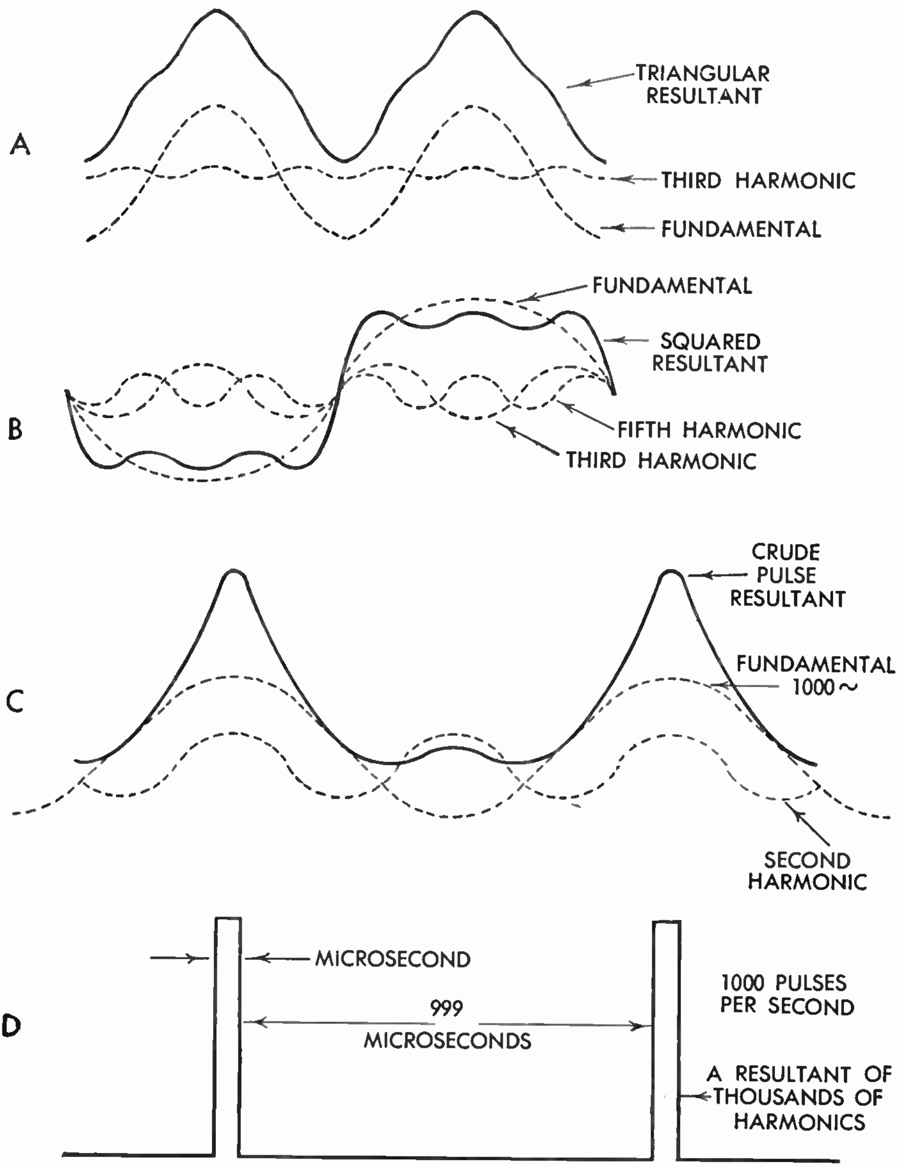


FIG. 348A Components of Complex Waves

requires a high pass filter, and the shorter the time constant the more effective is the removal of the low-order harmonics.

We have considered the action of an *RC* combination upon application of d-c or sine-wave voltages. Still another important consideration as far as television is concerned is the behavior of an *RC* combination upon application of pulses of various durations. In effect, the action of a pulse is similar to the application of a d-c voltage which is switched on and off at a predetermined repetition rate. When the pulse is applied the voltage of the flat top is impressed on the *RC* combination for the duration of the pulse (switch closed)—then is removed (switch open)—then reapplied in a definite time sequence.

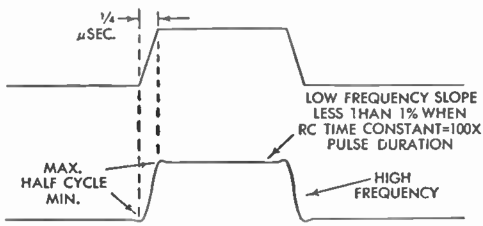


Fig. 349 Frequency Components of Pulse

The three synchronizing pulses of the composite television signal are horizontal sync, equalizing sync block, and vertical sync block. Leading edges of all sync pulses rise in 0.254 microsecond. Actually, the leading and trailing edges of these pulses represent high-frequency components. For example, Fig. 349 shows the leading edge of a pulse compared to one-half cycle of a sine wave (minimum to maximum). This sine wave rises from minimum to maximum in ¼ microsecond approximately and, therefore, has a period of ½ microsecond. A ½-microsecond period represents a base frequency of 2 megacycles. It is true that this is not continuous transmission of a 2-megacycle frequency. Nevertheles, if it is transmitted for a lengthy period or only a portion of a period, the same frequency-response requirements are imposed on the amplifier or circuit through which the leading edge is to be passed. Thus, if the leading edge is to be passed the time constant of the *RC* combination can be as little as:

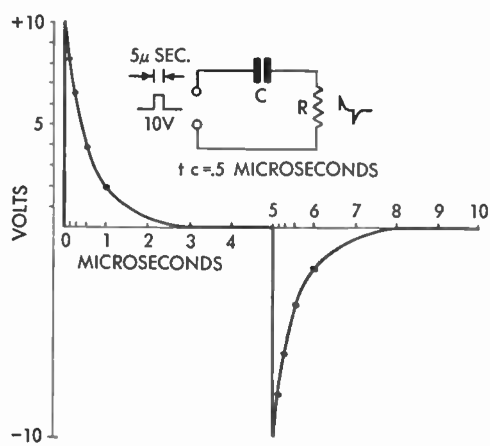


Fig. 350 Voltage across Differentiating Resistor

$$\text{Time constant} = \frac{5}{2\pi f} = \frac{5}{6.28\ (2)(10^6)} = 0.4\ \mu\text{sec}$$

Most differentiating circuits in television receivers have time constants which vary from 0.2 to 10 microseconds. This means the high-frequency components of the pulse pass readily; low-frequency components are rejected. This circuit is used to segregate leading edge from remainder of pulse.

It is also necessary that the fidelity of the pulse leading edges be preserved and shunt time constants of the plate load resistors and distributed circuit capacities must be kept low to prevent integration and rounding-off of the leading edge. Consequently, time constants of load circuits must be kept preferably lower than the period of a 2-megacycle wave.

In an integrator circuit the output voltage appears across the capacitor. Inasmuch as it is necessary for the capacitor to charge during the vertical sync block, the time constant is long in comparison to the approximately 200-microsecond duration of the vertical sync block. Thus, the horizontal differentiating circuit is a high-pass filter which conveys the leading edges of the pulses; vertical integrating circuit, a low-pass filter which functions in accordance with pulse duration. The time constants of the commercial integrating circuits vary from 500 to 2,000 microseconds.

Differentiation and integration by simple *RC* circuits is demonstrated in the following charts. In the first chart (Fig. 350), a plot of the output across the differentiating resistor has been made when a 5-microsecond pulse (comparable to horizontal sync pulse duration) is applied to a 0.5-microsecond *RC* combination. If the pulse amplitude is 10 volts, 10 volts appear across the resistor at the instant the pulse is applied ($t = 0$). One-tenth microsecond later the resistor voltage has dropped to 8.2 (capacitor charge at this instant is 1.8 volts) as indicated on the chart of Fig. 344 for 0.1 miscrosecond, or 1/5 time constant. Using the chart, various other points at time intervals of 0.2, 0.5, 1, 2, and 3 miscroseconds have been plotted. At the end of 3 microseconds (six time constants) the voltage across *R* is zero and the capacitor is fully charged. This static condition continues until the end of the pulse at 5 microseconds. At this point the pulse is removed and the capacitor storage is no longer sustained—the algebraic sum of voltage drops across *R* and *C* must now equal zero. Thus, at the instant the pulse is removed a *negative* 10 volts appear across *R*, which falls toward zero as the original positive 10 volts on *C* discharge through *R*. Using the chart again, points have been plotted at 5.1, 5.2, 5.5, 6, 7, 8, and 10 miscroseconds. Observation of the plotted waveform shows the spiked differentiated wave used to synchronize the horizontal oscillator.

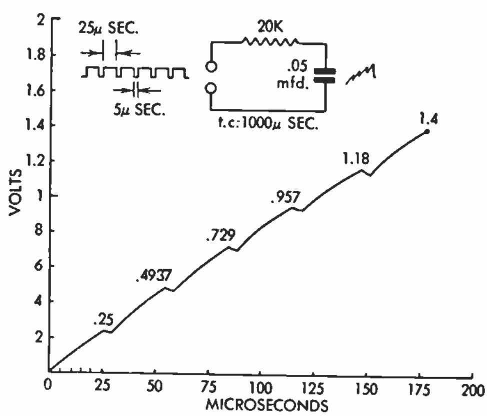


FIG. 351 Voltage across Integrating Capacitor

The action of a simple integrating *RC* combination can be demonstrated in like manner. The chart of Fig. 351 shows the voltage building up on the integrating capacitor in a series of steps in accordance with the duration and spacing of the pulses of the vertical sync block. In the example, six 25-microsecond pulses with 5-microsecond spacing have been chosen (175 microseconds of block). Again, if 10-volt pulses are applied, capacitor *C* will charge for 25 microseconds and discharge for 5, slowly charging toward 10 volts—actually, at end of last pulse, charge is still under 2 volts with 1,000-microsecond time constant used.

The capacitor charge 10 microseconds after start of the first 25-microsecond pulse is as follows:

$$e_c = E - E\varepsilon^{-t/RC}$$
$$e_c = 10 - 10\,(2.718^{-10/1{,}000}) = 0.1 \text{ volt}$$

At the end of the first 25-microsecond pulse the capacitor charge is

$$e_c = E - E\varepsilon^{-t/RC}$$
$$e_c = 10 - 10\,(2.718^{-25/1{,}000}) = 0.25 \text{ volt}$$

Now the 10 volts are removed for 5 microseconds and the capacitor discharges 0.00125 volt.

$$e_c = 0.25 - 0.25\,(2.718^{-5/1{,}000}) = 0.00125 \text{ volt}$$

The charge on capacitor at start of second 25-microsecond pulse is therefore

$$e_c = 0.25 - 0.00125 = 0.24875 \text{ volt}$$

From the above it is evident that with a 1,000-microsecond time constant the capacitor will alternately charge to 2½ per cent of applied voltage and discharge only 0.5 per cent of its charge between pulses.

At the start of the second pulse the residual charge on *C* is 0.24875 volt and, consequently, the applied voltage is 10 − 0.24875 or 9.75125 volts. The charge at the end of the second pulse is now 2½ per cent of 9.75125 volts added to the residual charge of 0.25 volt from the first pulse.

$$e_c = 0.025\,(9.75125 + 0.24875) = 0.4925$$

During space between the second and third pulse the capacitor loses 0.5 per cent of 0.4925, or charge on *C* at start of third pulse is

$$0.4925 - (0.005 \times 0.4925) = 0.49$$

A continuation of this procedure places a voltage of 1.4 on the capacitor at the end of the sixth pulse.

SQUARE-WAVE RESPONSE

Likewise, the importance of time constant and the fidelity of a pulse which is to be conveyed is demonstrated in Fig. 352. In this chart a 200-microsecond pulse (approximately vertical sync block duration) is applied to a 2,000-microsecond time-constant RC combination. Resistor voltage is plotted. Again, at start of the pulse full 10 volts appear across R and capacitor begins its charge. At the end of 50 microseconds the voltage across R is

$$e_r = E\varepsilon^{-t/RC}$$

$$e_r = 10\ (2.718^{-50/2{,}000}) = 9.75 \text{ volts}$$

At the end of 200 microseconds the voltage across R has fallen to

$$e_r = 10\ (2.718^{-200/2{,}000}) = 9.05 \text{ volts}$$

Thus, the flat top of our pulse has lost approximately 1 volt, and this slope or drop-off distorts it somewhat with respect to the original. If absolute fidelity is to be sustained time constant should be fifty to one hundred times pulse duration, permitting only an extremely tiny charge to build up on C (also proven in our discussion of RC filtering).

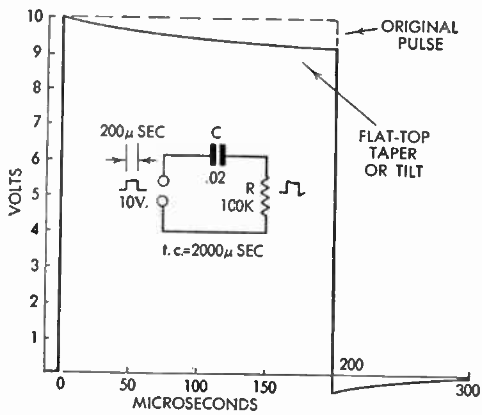


FIG. 352 Conveying a Pulse through RC Coupling Network

Whenever t/RC has a value smaller than 0.1 the formulas for e_c and e_r simplify. Normally

$$e_c = E - E\varepsilon^{-t/RC}$$

$$e_r = E\varepsilon^{-t/RC}$$

Inasmuch as e^{-x} is simply $1 - x$ whenever x is less than 0.1 (refer to table of exponentials) the following relations can be used to simplify calculations:

$$e_c = E - E\left(1 - \frac{t}{RC}\right)$$

$$e_c = E - E + E\left(\frac{t}{RC}\right)$$

$$e_c = E\left(\frac{t}{RC}\right)$$

also

$$e_r = E\left(1 - \frac{t}{RC}\right)$$

$$e_r = E - E\left(\frac{t}{RC}\right)$$

showing that $1 - (t/RC)$ can be substituted for $\varepsilon^{-t/RC}$ in those cases for which the time constant is very long in comparison to pulse time.

For a pulse of a given duration it is possible to calculate drop-off of flat top very simply if the time constant of the coupling RC is known. This decrease represents the charge on C (charge on C equivalent to drop in voltage on R) divided by peak amplitude of the pulse.

$$\text{Percentage decline} = \frac{e_c}{E} \times 100 = \frac{E\dfrac{t}{RC}}{E} \times 100 = \frac{t}{RC} \times 100$$

A more exact solution for drop-off in excess of 10 per cent uses the complete formula

$$\text{Percentage decline} = \frac{E - E\varepsilon^{-t/RC}}{E}\,(100) = 1 - \varepsilon^{-t/RC}\,(100)$$

The low-frequency limit (3 decibels down) of same RC combination can be found using the simple relation

$$R = \frac{1}{2\pi f_L C} \qquad \text{and} \qquad f_L = \frac{1}{2\pi RC}$$

For example, if a 20-microsecond pulse is applied to an interstage combination with a time constant of 1,000 microseconds the drop-off of the flat top of the pulse can be expected to be

$$\text{Percentage decline} = \frac{t\,(100)}{RC} = \frac{20\,(100)}{1{,}000} = 2\%$$

The response of a parallel RC combination (plate load resistor and distributed circuit capacity) to rise time (t) of a pulse can be anticipated. Rise of voltage on capacitor is again

$$e_c = E - E\varepsilon^{-t/RC}$$

Fractional build-up of voltage on the capacitor as compared to peak pulse voltage is

$$\frac{e_c}{E} = \frac{E - E\varepsilon^{-t/RC}}{E}$$

$$\frac{e_c}{E} = 1 - \varepsilon^{-t/RC}$$

Solving for RC (time constant for given rise time)

$$RC = \frac{t}{\log_\varepsilon \left(1 - \dfrac{e_c}{E}\right)^{-1}}$$

For example, if rise time of the circuit is to be 0.25 microsecond to reach 90 per cent peak, time constant of any $R_L C_t$ may be no more than

$$RC = \frac{0.25\,(10^{-6})}{\log_e\,(1 - 0.9)^{-1}} = 0.109\ \mu\text{sec}$$

To meet this requirement high-frequency limit of a single stage must be

$$f_o = \frac{1}{2\pi C_t R_L}$$

$$f_o = \frac{10^6}{6.28\ (.109)(10^{-6})} = 1.44\ \text{mc}$$

175. *Sawtooth Generation*

The exponential rise of voltage on a capacitor is also used to generate a sawtooth voltage. In the generation of such a sawtooth only the early portion of the charge, which is essentially linear, is used. A vacuum tube (Fig. 353), which is turned on and off at a prescribed rate, generates the sawtooth voltage. When the tube is cut off the capacitor C charges through resistor R; when the

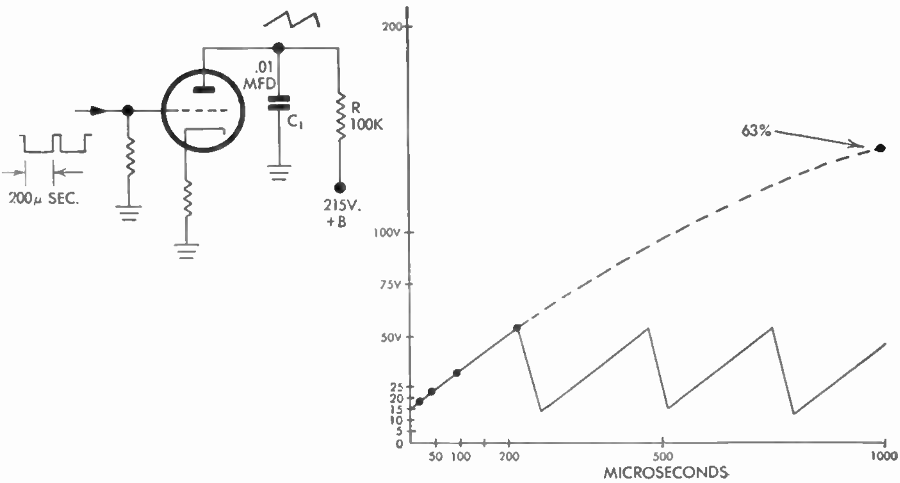


FIG. 353 Generation of a Sawtooth

tube conducts, the capacitor is discharged by the low resistance of the conducting tube. As shown, the capacitor begins charging from a low plate-voltage point (15 volts) at the instant the tube is cut off (low plate-voltage point is the plate voltage of the tube when conducting) toward the supply voltage for the time during which the tube is cut off.

If the tube is cut off only 1/5 time constant the capacitor will only charge over the linear portion of its curve. For example, if the tube is cut off by a 200-microsecond pulse the capacitor is charged 5 volts in 25 microseconds.

$$e_c = E - E\varepsilon^{-t/RC}$$

$$e_c = 200 - 200\ (2.718^{-25/1,000})$$

$$e_c = 5\ \text{v}$$

At the end of 50, 100, and 200 microseconds, the capacitor charges are respectively 10, 19, and 36.2 volts. After 200 microseconds the curve begins to round off and the sawtooth becomes nonlinear. At this point the tube again conducts and the capacitor discharges. The discharge is very fast because its time constant is short and equal to R_pC_t (R_p is the tube's plate resistance, which is low).

This very same method is used to generate the sawtooth sweep voltages of a television receiver. In fact, the method just discussed has the same sequence of operation is a discharge tube driven by a blocking oscillator (Fig. 354). Likewise, the plate circuit of the second section of the multivibrator generator functions in the same manner—the capacitor charging when the second section is cut off, discharging when the tube conducts.

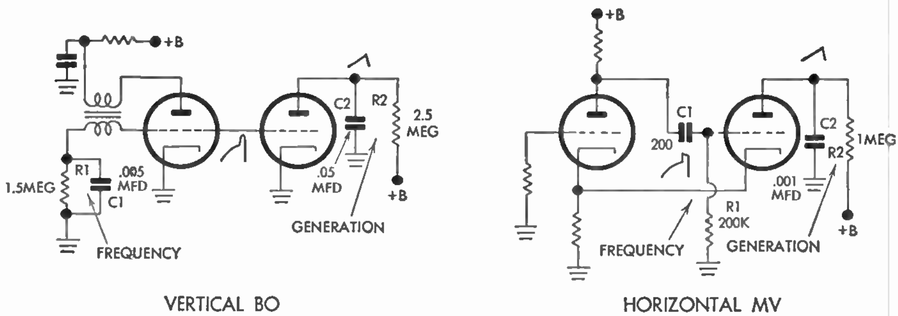


FIG. 354 Blocking Tube and MV Sawtooth Generator

The frequency of the blocking oscillator or the multivibrator sawtooth generator is a function of the grid-circuit time constants, or how long it takes the grid capacitor to discharge to the conduction point. For example, if the grid of the vertical blocking oscillator is driven to −200 volts by feedback and the conduction point is −20 volts, the approximate frequency of the oscillator can be calculated. Assume RC time constant is 7,500 microseconds. The capacitor discharges toward zero but reaches only the conduction level.

$$e_r = E\varepsilon^{-t/RC}$$

$$20 = 200\varepsilon^{-t/7,500}$$

$$\log_\epsilon 0.1 = -\frac{t}{7,500}\log_\epsilon \varepsilon$$

$$10 - 7.697 = -\frac{t}{7,500} \text{ (from natural log tables)}$$

$$-2.303 = -\frac{t}{7,500}$$

$$t = 17,300 \ \mu\text{sec}$$

This tells us it requires 17,300 microseconds to discharge from 200 to 20 volts. If another 700 microseconds is added for the conduction period of the tube, the period of the relaxation cycles is 18,000 microseconds, and one over this quantity gives us the base frequency of

$$f = \frac{1}{\text{period}} = \frac{1}{18{,}000} = 55\tfrac{1}{2} \text{ cycles per second}$$

176. *D-C Restoration*

The operation of a d-c restorer is also dependent on the time constant of its *RC* combination (Fig. 355). In this circuit negative-going composite signal drives the restorer diode into conduction. The fast time constant during diode conduction charges the capacitor to peak value of the applied signal. Inasmuch as the sync tips are the most negative parts of the composite signal, it is the

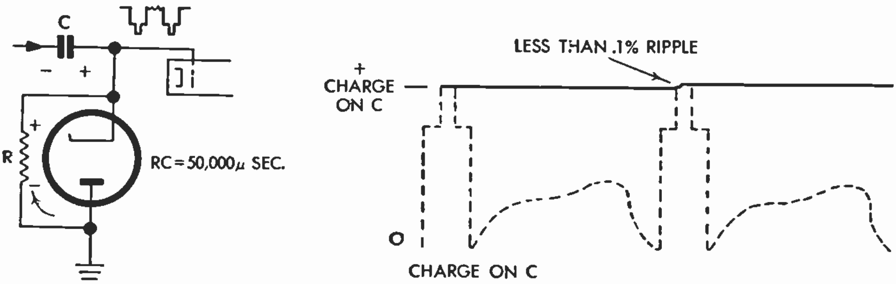


FIG. 355 D-C Restorer Time Constant

sync tips which set the peak charge of the capacitor. As soon as the composite signal drops below the sync tip the diode stops conducting because of the charge placed on the capacitor by the sync tip. The diode will not conduct again until the next sync tip (only part of the signal great enough in amplitude to overcome the capacitor charge). The diode is held cut off between sync tips (58.5 microseconds) by the negative capacitor charge which is sustained by the long time constant of *R* and *C*.

A typical restorer time constant is 50,000 microseconds, which would permit a discharge of only 0.1 per cent in 63.5 microseconds.

$$e_c = E - E\varepsilon^{-t/RC}$$

$$e_c = 1 - 1\ (2.718^{-58.5/50{,}000})$$

$$e_c = 0.001 \text{ part in } 1, \text{ or } 0.1\%$$

Thus, most restorer capacitors lose less than 0.1 per cent of their charge between sync tips, and the bias they contribute remains essentially constant. This bias, of course, changes with average brightness because the peak amplitude of the sync tips follows the relatively slow-changing scene background brightness. A change in level of the sync tips, of course, alters the capacitor charge and the bias that the restorer contributes.

177. *Wide-Band Amplification*

To pass linearly a broad band of frequencies with band-pass-tuned circuits, it is necessary to load the resonant circuits with resistance. Such loading reduces tuned circuit Q and impedance but widens the response. The actual load resistance (Fig. 356) is, in most cases (particularly for wide-band i-f amplifiers) dominantly a physical resistor of correct value shunted across the tuned circuit. The tube plate resistance and a shunt resistive component contributed by the inductor are also factors which figure in the total R value. In a properly designed system, however, the dominant resistance is the physical resistor, and its value can be used to approximate the stage gain.

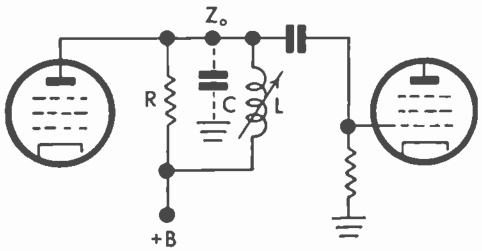


FIG. 356 Wideband Amplifier Output Circuit

$$\text{Gain} = g_m \frac{RZ_o}{R + Z_o} \quad (Z_o = \text{impedance of resonant circuit})$$

Again, in wide amplification $Z_o > R$ and gain becomes simply

$$\text{Gain} = g_m R$$

At frequencies off resonance, however, the reactive component decreases toward the value of the loading resistance R. Off resonance the gain of the stage becomes

$$\text{Gain} = g_m \frac{Rx}{\sqrt{R^2 + x^2}}$$

Inasmuch as $x = R$ at *some frequency,* the gain of the stage at this particular frequency is

$$\text{Gain} = \frac{g_m Rx}{\sqrt{R^2 + x^2}}$$

$$\text{Gain} = \frac{g_m R^2}{\sqrt{2R^2}} = \frac{g_m R}{1.414} = 0.707\, g_m R$$

Thus, at the frequency points (one on each side of resonance) at which the reactive component of the tuned circuit equals the value of the loading resistor, the gain of the amplifier is at the 70.7 per cent level of peak gain. The frequency range between the two 70.7 per cent levels, one on each side of the resonant frequency, is generally considered to be the band width of the tuned circuit.

It is apparent that the gain of the stage is directly proportional to the value of the load resistor. However, the size of the load resistor is limited by the

x_L and x_c at resonance and the band width of the stage, as proven by the following equations.

The inductive reactance of the tuned circuit at some point off resonance is

$$x_L = 2\pi f_r L \pm 2\pi f_\Delta L$$

in which f_r equals resonant frequency and f_Δ change in frequency (band width) on one side of resonant frequency. The ratio by which the inductive reactance has been changed is

$$\text{Ratio} = \frac{2\pi f_r L \pm 2\pi f_\Delta L}{2\pi f_r L} = \frac{f_r \pm f_\Delta}{f_r}$$

Inductive reactance, therefore, at any frequency off resonance is simply x_L at resonance times the ratio by which the frequency has been raised or lowered.

$$x_{L_1} = \underset{(\text{resonance})}{x_L} \left(\frac{f_r \pm f_\Delta}{f_r}\right)$$

Now let us derive the equation which shows us the capacitive reactance at some frequency off resonance in the same manner.

$$x_c = \frac{1}{2\pi f_r C \pm 2\pi f_\Delta C}$$

$$\text{Ratio} = \frac{\dfrac{1}{2\pi f_r C \pm 2\pi f_\Delta C}}{\dfrac{1}{2\pi f_r C}} = \frac{f_r}{f_r \pm f_\Delta}$$

$$x_{c_1} = \underset{(\text{resonance})}{x_c} \text{ times } \frac{f_r}{f_r \pm f_\Delta}$$

The combined reactances or reactive component of tuned circuit off resonance is

$$x = \frac{\underset{(\text{resonance})}{x_L}\left(\dfrac{f_r \pm f_\Delta}{f_r}\right) \underset{(\text{resonance})}{x_c}\left(\dfrac{f_r}{f_r \pm f_\Delta}\right)}{\underset{(\text{resonance})}{x_L}\left(\dfrac{f_r \pm f_\Delta}{f_r}\right) - \underset{(\text{resonance})}{x_c}\left(\dfrac{f_r}{f_r \pm f_\Delta}\right)}$$

At the resonant frequency $x_L = x_c$. Let us term this the reactance x_r at resonance and simplify our equation

$$x = \frac{x_r^2}{x_r\left(\dfrac{f_r \pm f_\Delta}{f_r}\right) - \left(\dfrac{f_r}{f_r \pm f_\Delta}\right)}$$

Also, at the 70.7 per cent level, $x = R$, or

$$R = \frac{x_r f_r\,(f_r \mp f_\Delta)}{f_r^2 \mp 2f_r f_\Delta \mp f_\Delta^2 - f_r^2}$$

$$R = x_r \frac{f_r\,(f_r \pm f_\Delta)}{f_\Delta\,(2f_r \pm f_\Delta)}$$

Inasmuch as $f_\Delta < f_r$ for television wide-band amplification:

$$R = \frac{x_r f_r^2}{2f_\Delta f_r} = \frac{x_r f_r}{2f_\Delta} = \frac{f_r}{2\pi f_r C \cdot 2f_\Delta}$$

These, then, are the basic formulas for loading a tuned circuit with a resistor. Thus, for a given L-to-C ratio, resonant frequency, and bandwidth, the size of the loading resistor to obtain this bandwidth (between 70.7 per cent levels) can be calculated.

Inasmuch as value of R_L determines stage gain ($g_m R_L$) we can see that the stage gain is less as the bandwidth requirements are increased. Likewise, the stage gain is increased with a higher L-to-C ratio, because the size of the loading resistor per given bandwidth is higher (x_r in the formula is high when the inductance is high and the capacity low).

Solving for $2\Delta f$, the bandwidth of a single tuned circuit is:

$$B = \frac{f_r}{2\pi f_r RC} = \frac{1}{2\pi RC}\left(x_r = \frac{1}{2\pi f_r C}\right)$$

When a single tuned circuit is used, gain of the stage is, of course, $g_m R$. The product of gain and bandwidth or figure of merit is:

$$GB = g_m R \times \frac{1}{2\pi RC} = \frac{g_m R}{2\pi RC} = \frac{g_m}{2\pi C}$$

This gain-bandwidth factor is a constant and from it the effect of change in bandwidth on gain or vice versa is immediately evident. For example the GB of a 6AC7 (g_m of 9,000 micromhos and C_t of 25 μμf) is approximately 60. Thus the 6AC7 with a single-tuned load and a bandwidth of 4 mcs. would have a gain of 15 (4 times 15) or with a bandwidth of 6 mcs. a gain of 10 (6 times 10 equals 60 again).

If a number of similar stages follow consecutively, tuned to the same frequency and with the same bandwidth, the over-all bandwidth drops off with each new stage added. Over-all bandwidth equals:

$$\text{Bandwidth} = \frac{\text{Bandwidth of single stage}}{1.2\sqrt{\text{Number of stages}}}$$

To avoid this reduction in bandwidth the stages can be stagger-tuned. In the above system the extremes of the bandpass were 30% down for a single stage and inasmuch as the drop-off accumulates, at the second stage the same two frequency extremes are 50% down. When stagger-tuning is used, however, each stage has a different center frequency (most common practice in TV receiver i-f systems) and thus the gain per stage can be kept high because the bandwidth per stage can be less than the desired over-all bandwidth.

The preceding wide-band computations apply particularly to the single-tuned stage. If a double-tuned transformer is used the capacities are isolated and a higher L-to-C ratio results for each tuned circuit. Both resonant circuits are tuned to the same frequency and have almost identical Q's. When the tuned circuits are overcoupled two peaks appear above and below the resonant frequency. Separation of these peaks, as determined by the degree of coupling and resistive loading, sets the bandwidth. Stage gain is higher because of improved L-to-C ratios and less severe loading for a given bandwidth.

When a double-tuned transformer is used additional stage gain can be obtained with somewhat better phase and overshoot characteristics. Likewise each or one tuned circuit is shunted by a loading resistor to obtain desired bandwidth. However, less loading is necessary because of the additional bandwidth obtained with overcoupling. Gain of stage using a double-tuned transformer is:

$$\text{Gain} = \frac{g_m\sqrt{R_pR_s}}{2}$$

for the equal Q case which is generally used in low Q wide-band double-tuned stages. For the unequal Q tuned circuits:

$$\text{Gain} = \sqrt{2}\, g_m\, R_s \sqrt{\frac{C_s}{C_p}}$$

(In which Q_p is very much higher than Q_s)

Bandwidth for the equal Q case is:

$$\text{Bandwidth} = \frac{\sqrt{2}}{2\pi RC}$$

in which $R = \sqrt{R_pR_s}$ and $C = \sqrt{C_pC_s}$.

Bandwidth for unequal Q type is:

$$\text{Bandwidth} = \frac{1}{\sqrt{2}\, 2\pi R_sC_s}$$

(In which Q_p is very much higher than Q_s)

QUESTIONS

1. Calculate values of the component parts for a three-stage video amplifier using two 6AC7 video amplifiers and your choice of a cathode-follower output tube. Frequency response is to be linear from 10 cycles to 4 megacycles.
2. If a 10-microsecond pulse is applied to a series-coupling R-C combination with a time constant of 15 microseconds, what is the percentage of drop-off?
3. Design a network to differentiate a 5-microsecond 30-volt pulse. Plot E applied, e_c, and e_r voltages on graph paper.
4. Plot the integrated wave when the FCC standard vertical sync block is applied to a 500-microsecond integrating network.
5. Plot the output of the integrator as in the previous example, during the interval that six equalizing pulses are applied. Assume a residual charge on the capacitor at the start of the first equalizing pulse of 1 volt.

6. Discuss constant-voltage and constant-current conceptions of vacuum-tube operation.
7. Of what significance is time constant of a decoupling *RC* combination?
8. Differentiate between phase shift, time delay, and nonlinear time delay.
9. Discuss low-frequency and high-frequency phase shift.
10. Discuss important considerations and computations involved in the design of a wide-band i-f or r-f amplifier.

Chapter 15

UHF TELEVISION

178. *Propagation at UHF*

Propagation characteristics at UHF are not nearly as favorable as at VHF frequencies. In the propagation of a VHF-UHF wave, the signal tends to travel in almost as straight a line as does a beam of light. However, in traveling through our immediate atmosphere or troposphere it encounters some bending or refraction. This refraction is largely the result of the declining density of the atmosphere with height above ground—it causes the top of the propagated wave to speed up slightly with respect to the lower section, and the propagated energy gradually bends over. It bends and follows the curvature of the earth for a limited distance.

In VHF-UHF propagation a new term for the path of travel is assigned and called "signal line-of-sight." This is to be differentiated from the term "optical line-of-sight." We realize "optical line-of-sight" refers to the distance we can see to the horizon before our vision is interrupted by the curvature of the earth. This is a straight-line distance. As far as the propagated wave is concerned it encounters some bending due to refraction, and signals travel beyond the horizon, Fig. 357. In this illustration the optical line-of-sight path is shown with a dashed line, the signal line-of-sight by a curved, continuous line.

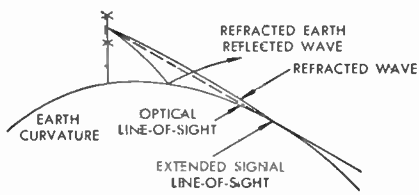


Fig. 357 VHF-UHF Propagation

At the low-frequency end of the VHF-UHF spectrum, signal line-of-sight is a substantial distance beyond the horizon, separation of the signal and optical lines-of-sight decreasing as we go higher and higher in frequency until they essentially overlap at some very high microwave frequency. This fact causes our high-band stations to have, in general, less range and coverage than our low-band stations. UHF channels have less range. In fact, many, many times the effective radiated power of our present VHF stations is necessary to obtain only 70 per cent of the coverage of our present VHF stations. Such a limitation means that the UHF channels can be used best for small-city allocations or for strictly local coverage in metropolitan districts, unless very high antenna heights and power are used.

The UHF wave is not as easily pulled from its path as the VHF wave. This is a disadvantage because it minimizes signal levels in back of obstacles. The VHF wave will fill in to a certain extent at the rear of an obstacle, such as a hill or structure, while the UHF wave dips but slightly, Fig. 358. Thus UHF signal levels are far weaker in back of an obstacle, and there are definite signal empty pockets in UHF coverage.

Channel frequencies and element wavelengths in inches are shown below in chart form.

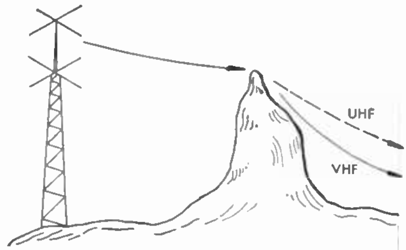


FIG. 358 VHF-UHF Coverage at Rear of Hill

NEW UHF TV CHANNELS

Channel	*Wavelength in Inches at Channel Center*	*Frequency in Megacycles*	*Channel*	*Wavelength in Inches at Channel Center*	*Frequency in Megacycles*
14	24.7	470–476	49	17.1	680–686
15	24.4	476–482	50	16.95	686–692
16	24.1	481–488	51	16.8	692–698
17	23.8	488–494	52	16.65	698–704
18	23.5	494–500	53	16.5	704–710
19	23.2	500–506	54	16.35	710–716
20	22.9	506–512	55	16.2	716–722
21	22.6	512–518	56	16.05	722–728
22	22.3	518–524	57	15.9	728–734
23	22	524–530	58	15.75	734–740
24	21.8	530–536	59	15.65	740–746
25	21.6	536–542	60	15.55	746–752
26	21.4	542–548	61	15.45	752–758
27	21.2	548–554	62	15.35	758–764
28	21	554–560	63	15.25	764–770
29	20.8	560–566	64	15.1	770–776
30	20.6	566–572	65	15.0	776–782
31	20.4	572–578	66	14.9	782–788
32	20.2	578–584	67	14.8	788–794
33	20	584–590	68	14.7	794–800
34	19.8	590–596	69	14.6	800–806
35	19.6	596–602	70	14.45	806–812
36	19.4	602–608	71	14.35	812–818
37	19.2	608–614	72	14.25	818–824
38	19	614–620	73	14.15	824–830
39	18.8	620–626	74	14.05	830–836
40	18.6	626–632	75	13.95	836–842
41	18.4	632–638	76	13.85	842–848
42	18.2	638–644	77	13.75	848–854
43	18.0	644–650	78	13.65	854–860
44	17.85	650–656	79	13.55	860–866
45	17.7	656–662	80	13.45	866–872
46	17.55	662–668	81	13.35	872–878
47	17.4	668–674	82	13.25	878–884
48	17.25	674–680	83	13.2	884–890

INFLUENCE OF WEATHER

Weather has a decided influence on VHF-UHF propagation. In fact, fringe-area signals are very much a function of the weather. As a general statement, it can be said that coverage is better in hot, humid weather than in cold, dry weather and by night than by day. Even within 10 miles of a station there is some signal climb at night. In addition, signal levels vary from hour to hour, day to day, and month to month as atmospheric conditions change. These variations are most pronounced in fringe areas.

Weather gradients determine how much the VHF-UHF wave is refracted. Therefore, the signal line-of-sight shifts to and fro with respect to the fixed position of the optical line-of-sight. In cases of decided subnormal refraction, the signal line-of-sight comes in very near to the position of optical line-of-sight, and station range is poor. At other times when longer-range conditions are good (abnormal refraction), the signal line-of-sight is miles beyond the optical line-of-sight, and the station is placing the signal into a vast area. As far as UHF range is concerned, these decided signal-level shifts occur nearer to the station because the average signal line-of-sight is much nearer to the optical line-of-sight.

SOME UHF-PERFORMANCE NOTES

When one is able to spend days and days in many sections of a UHF area instead of only a few well-chosen spots, the idiosyncrasies of the field are revealed. Per given power and antenna height, the range is not as great nor as consistent as that of VHF. It is true that there are isolated areas that seemingly disprove this statement—but we must base our reasoning on majority, and not minority, results. Ignition and man-made noises are pleasingly absent on the UHF picture, but the influence of scattered ground reflections and the activity on the ground are more decided. Now this does not imply that good UHF results cannot be obtained—but we recognize that for many installations the technician must be skilled, careful, and must take the required time to do thorough work if the receiver is to derive full benefits from a received signal. Some of the conditions encountered are as follows:

1. Use of present VHF antennas now mounted and peaked for VHF reception are satisfactory in only a few locations (a matter of chance location where one of their peak lobes is in the direction of a station and no serious reflections are present). A UHF signal delivered to the converter from this type is often weak because of improper orientation and high line losses. With the UHF-inferior line used with this type of installation, signal levels vary greatly with weather, as a result of moisture absorption and moisture-laden dirt or salt deposit on the line. At other locations, the signal may be strong but the picture may be afflicted with smear or reflections. This type of antenna has multiple lobes and must be oriented critically if smear, caused by the scattered near-

reflections or the multiple images of a spaced reflection, is to be minimized.

2. Where UHF signal levels are high, the picture resolution is often a problem new to the heretofore strictly fringe location. Here is an opportunity to enjoy a crisp, sharp picture. Thus careful installation and orientation of antenna have more than one major objective: not only reception of a strong signal but a clear, well-defined one as well. This very requirement can make choice of antenna-type and mounting position as critical as choice and installation of an antenna system for long-distance reception. A poor installation, though it may receive a strong signal, can introduce smear, transients, or reflections. This condition, to an extent, is less critical with the use of a UHF-band cut antenna (rather than a modified VHF-UHF type) because of the fewer minor lobes and the broader major lobe.

3. The UHF reflections are sharp and have a decided influence on the signal level at the antenna. The scattered, near reflections that arrive at almost the same angle as that of a direct signal can add or subtract from the direct signal, having a substantial influence on the apparent signal strength at the antenna position. This condition makes the UHF antenna-mounting position critical, both horizontally and vertically, where strong reflections exist.

This reflection sharpness also causes picture-level breathing as the antenna sways in the wind. The same signal-level breathing and flutter over a wide amplitude range can be caused by aircraft or moving vehicles, thus imposing strict requirements on a-g-c systems. In one test location the motion of highway traffic cutting perpendicularly across the angle of signal arrival had an influence on signal levels. In fact, a large tractor-trailer, when it passed along a critical segment of the highway, did cause a 10-decibel swing of signal level.

4. In city districts the indoor antenna has not been satisfactory because of a weak signal at low levels (particularly when closed-in by buildings) and the annoying signal-strength variations caused by traffic motion or room activity.

5. At the moment, the corner reflector, bow-tie reflector and Yagi-type antennas are preferred for many areas—high gain, small size, and good pattern. The VHF-UHF end-fire Vees and the trombones are good UHF antennas (except for multiple lobes) on UHF position, but the average 2- to 5-decibel loss that must be taken on the VHF band, as compared to VHF types now existing in some areas, is too much for a VHF fringe area. Furthermore, many new UHF areas are old VHF fringe areas (not a brand new television area), and most VHF antennas are in position and operating. UHF is merely an addition to existing service.

179. UHF *Antennas and Transmission Lines*

The same basic types of antennas can be used in the assigned UHF band (470 to 890 megacycles) as are used for our present VHF allocations. Of course, antennas are much shorter and compact because a wavelength in this UHF range is only about one to two feet in length. From this it is apparent that high-gain multi-element systems can be used without undue physical proportions. Some representative types used and checked for UHF work are shown in Fig. 359. The simplest antenna type is the dipole. For UHF operation and good broadband response, the dipole elements can be flat metal triangles. Driven-element and reflector-director combinations can also be used—even a Yagi, if utmost gain is desired. Stacked collinear elements or end-fire broadside groups can be used. End-fire groups function quite well because of their broad unidirectional pattern.

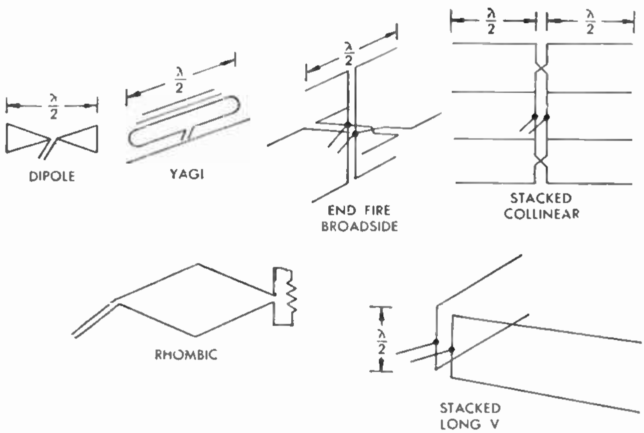
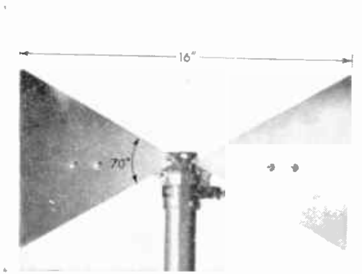
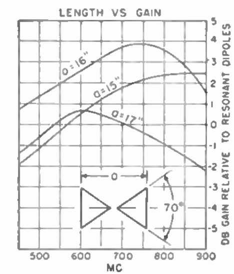
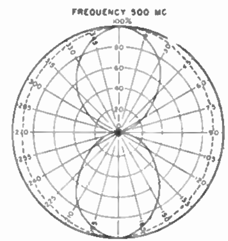
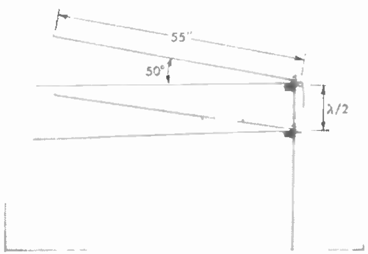
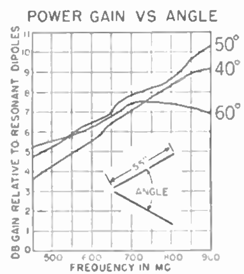
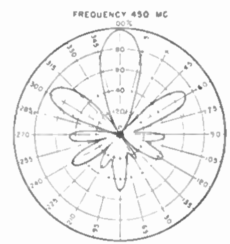


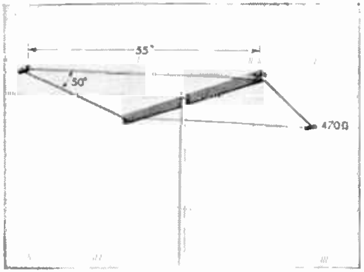
Fig. 359 UHF Antennas

Rhombic and long *V* antennas also function and are not massive at this high frequency. One successful and inexpensive type tested was that of two stacked *V*'s. The *V* elements are made a few wavelengths long in order to obtain high gain and are stacked to obtain a good vertical pattern. A sharpened vertical pattern means less pick-up from ground levels. Half-wavelength spacing of stacked *V*'s permits center feed and ideal matching. This same antenna also functions on the VHF band.

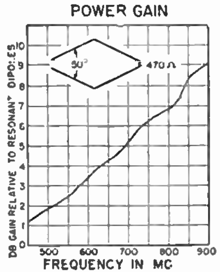
A most important consideration in UHF work is the proper positioning of antenna. The UHF wave is more compact, and reflections are sharper—thus positions of wave addition (space loop) and subtraction are quite sharp and decided. Check for space loops (where the preponderance of arriving waves is additive) near the tentative mounting position, and try to position the UHF antenna where it will deliver a peak signal to the receiver input. The influence of reflections has been brought out distinctly in some of the field checks. It was found that in some few sites no more signal was delivered to the receiver input by a high-gain antenna than by a simple dipole. At other sites relatively free of multiple reflections the advantages of a high-gain antenna were evident. These extreme cases do bear out the importance of probing a site before positioning of a UHF antenna.

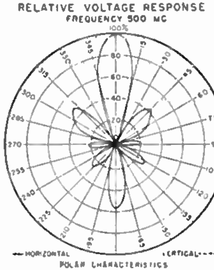
UHF ANTENNA PERFORMANCE CHART

TYPE AND DIMENSIONS	GAIN	PATTERN	CHARACTERISTICS
 TRIANGLE DIPOLE (BOW-TIE)			The triangular dipole has a bi-directional pattern and a rising gain with frequency. It provides an optimum match over the UHF band and represents a simple, small, and effective antenna for UHF strong signal areas.
 STACKED V			The stacked V is a longer-element and higher-gain antenna for UHF reception. It is inexpensive to construct and permits longer-range reception. It does have a multiple-lobe pattern and cannot be used in areas with reflection problems or severe interference.

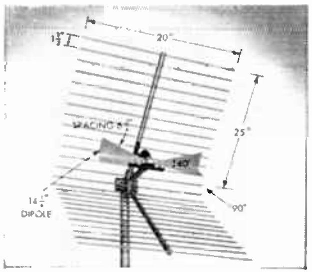


RHOMBIC

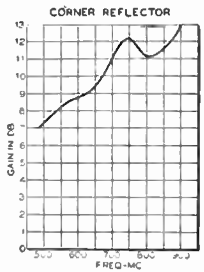


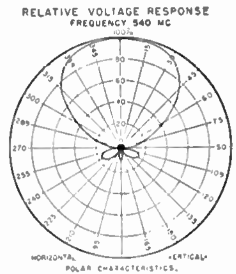


The rhombic antenna is a practical type for UHF reception because proper rhombic length at these frequencies comes down to some 50 inches. It has a high gain and a sharp directional pattern and must be oriented carefully. Fringe UHF reception is possible with this type. Although some minor lobes exist, reasonably good rejection of reflections and interference is provided.

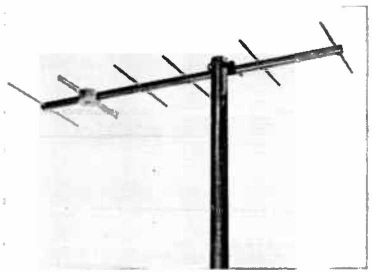


CORNER REFLECTOR

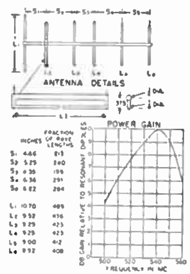


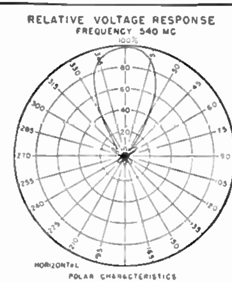


The corner reflector is a more elaborate UHF antenna, but it has a high gain and an excellent response pattern when dimensioned properly. It can be used for UHF fringe reception and, despite its high gain, will also have wide application in stronger signal areas because of its excellent rejection of reflections and interference. It will solve many propagation problems which other types cannot.



YAGI



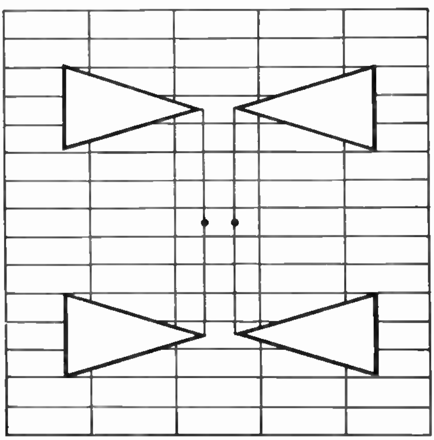


The Yagi is also an excellent UHF antenna but is restricted to operation on a limited number of adjacent channels. It has a high gain and a good directional pattern. Its rather narrow-band sensitivity means it has good noise and interference rejection (rejects image frequencies and many types of beat interference).

Data courtesy of Johnson and Kolar of RCA

Another interesting disclosure was that a broad unidirectional pattern is at times preferable at a site troubled by multiple reflections. It seems that reflections arriving approximately at the same angle as that of the direct signal reinforce that signal, while those arriving from the rear and at sharper angles reduce the receiver signal level. Thus, the single broad lobe obtained by a 90-degree UHF end-fire group helps, because of its rather wide-angle acceptance range in the forward direction and its effective rejection of signal coming in from the back and other sharp angles. It is true that combination VHF-UHF antenna types can be used, reducing the need for a dual-antenna installation in less troublesome locations.

A lengthy series of experiments was conducted at Bridgeport by Johnson and Kolar of the Advanced Development Section of RCA. The results of



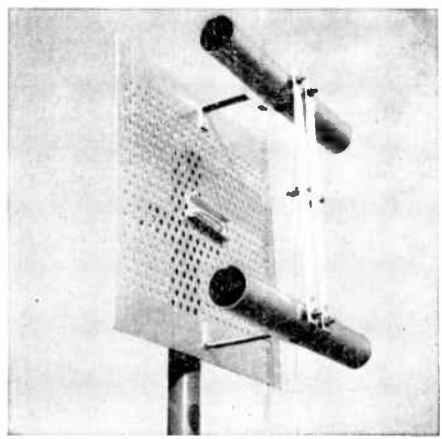


Fig. 360 Stacked UHF Antennas

these extensive tests are presented below in chart form. Information indicates the gain, pattern, and general characteristics of the various types checked and presents actual dimensions for operation in the UHF band. The wide-band types have been dimensioned to give an optimum broad-band response over the UHF band.

It is possible to obtain additional gain from any basic UHF-antenna type by proper stacking. As in VHF operation, an optimum stacking dimension would be between 0.5 and 0.8 wavelength, and of course, this dimension on the UHF band is substantially less than on the VHF band. Typical dimensions are a vertical spacing of some 10 to 12 inches, which is a half-wavelength at the low-frequency end of the UHF band and permits suitable stacking over the entire UHF band. Stacking increases gain by an average of 2.5 to 4 decibels over the band, while a 4-stack arrangement increases gain by a factor of 5 to 6.5 decibels over a single version of the same basic type.

Some stacked types, as illustrated in Fig. 360 are the bow-tie and large-diameter dipoles with screen reflectors. It should be mentioned that stacking not only raises gain (two antennas instead of one) but also that the bandwidth of the system is improved—response skirts are extended, and dips in the over-all response flatten out. Taco supplies a 4-stacked solid-wire bow-tie, either with or without screen reflector, Fig. 361.

The Yagi-type antenna is a popular one for UHF operation because its bandwidth limitation is not restricting on higher frequencies. For example, a 20-per cent bandwidth antenna near 600 megacycles represents a span of frequencies that could contain 20 television channels, while the same basic design near 60 megacycles would cover only a two-channel span. A Snyder wide-band Yagi, Fig. 362, covers approximately a 35-channel span; it is available in three models which respectively cover the lower half of the UHF band, upper half of the band, and a span of frequencies about the middle of the band.

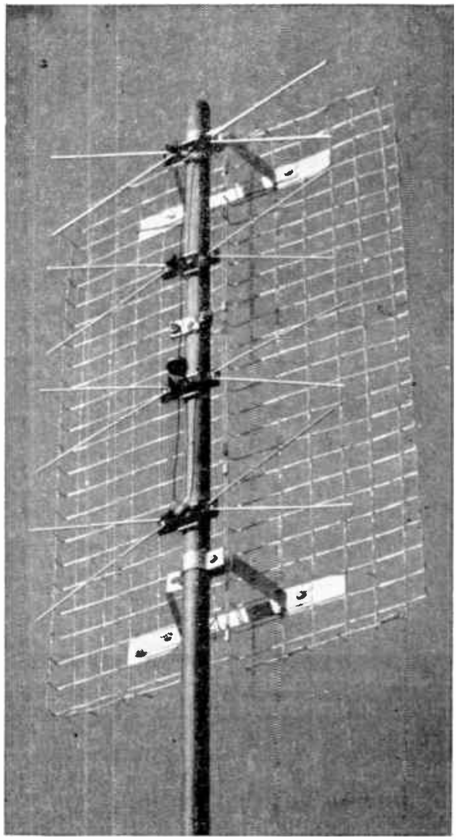
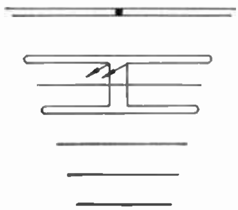


FIG. 361 Taco 4-Stacked Bow-Tie with Reflector

The plan of the antenna is to use a double-folded dipole driven element, each dipole favoring one end of the spectrum to be passed. A double reflector helps to hold up the gain over the span but is more instrumental in preserving a good front-to-back ratio. Directors favor the high end of the band and keep the pattern unidirectional over the entire span. A special parasitic element between dipoles acts at the high end as a reflector with relation to the short driven element and at the low end as a director with regard to the longer dipole, thus retaining gain and pattern uniformity. Stacked and four-stacked Yagi versions permit a high-gain and pure-pattern UHF antenna.

The combination VHF-UHF antennas evolve from the long-wire *V* principle, namely, that whenever an antenna is made longer than a half-wavelength its elements must be tilted forward in such a way as to align and form a major and sensitive forward lobe. For example, 44-inch elements form a half-wave antenna on channel *3*, Fig. 363. Elements of the same length function as a 3/2-wavelength antenna on channel 9, but to align the major lobe forward a

120-degree tilt is required. To operate the same elements on the UHF band would require a tilt angle of 65 degrees. It must be recognized that with the elements tilted so sharply, there is a decline in gain at the low-frequency end of the spectrum (channel *3* range).



	FREQ. RANGE	STACKING DIMENSION
A.	14–48	$12\frac{1}{2}$
B.	27–72	$10\frac{3}{4}$
C.	45–83	$8\frac{3}{4}$

FIG. 362 Snyder UHF Yagi

Three combination VHF-UHF types are illustrated in Fig. 364. It is to be noted that the tilt angle can be adjusted to favor either the VHF or UHF band. The gain curves demonstrate how the gain favors the UHF band for sharp tilt and the VHF band for the least tilt. The angle must be set according to the needs of the area. Antennas of this type also have many minor lobes and, consequently, are more sensitive to scattered reflections and ghosts and are more difficult to orient. The trombone version, because of its double-driven sections and partial stacking, retains a better optimum gain over the television spectrum.

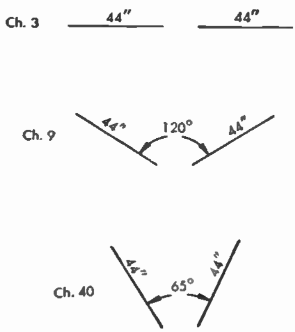


FIG. 363 Antenna Element Tilt Angle

A small- but still a high-gain and wide-band antenna can be evolved by dimensioning the antenna into the high band. We realize our present VHF antennas, though cut for the low band, have their higher gains on the high band, functioning as $3/2$-wavelength antennas. Driven elements must, of course, be tilted forward properly. It is significant that the same third-harmonic relation exists between the VHF high band and the UHF band and that an antenna cut for the high band will function as a basic half-wave on the high band and a higher-gain $3/2$-wavelength type on the UHF band. This channel *7*- to *83*-operation can be attained with a small antenna and without an excessive number of minor lobes present on the UHF band pattern.

An antenna of this plan could use screen-type elements—half-wave on the VHF high band and a tilt of 20 to 30 degrees to permit operation on the UHF band, Fig. 365. Stacked combinations further increase sensitivity, forming a high-gain UHF band antenna.

A second advantage is the fact that the same antenna functions with limited gain for high-band VHF operation. There are numerous small cities with

UHF and VHF high-band allocations only where this type of antenna is ideal, as only one antenna installation need be made. The same screen elements can be used in a Directronic version to permit reception of VHF high-band and UHF signals arriving from differing directions.

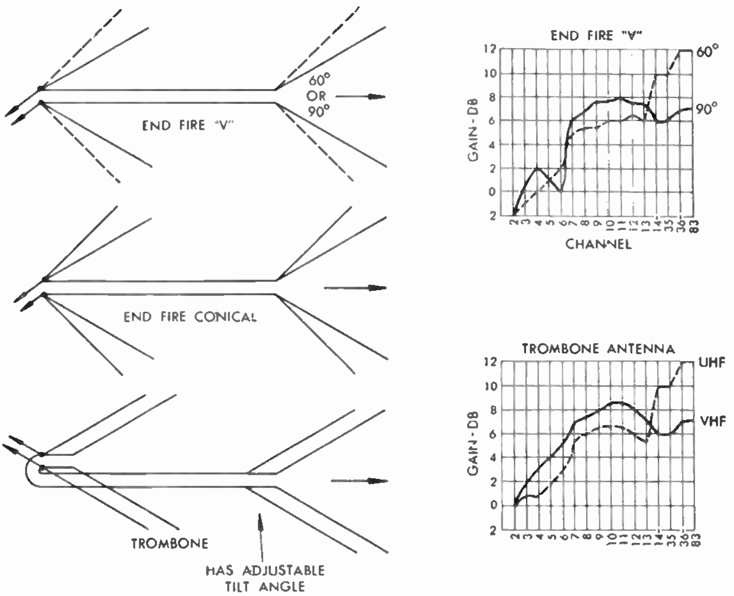


FIG. 364 VHF-UHF Antennas

UHF-VHF ANTENNA

One of the more perplexing problems of antenna evolution today is the development of a basic type that can offer all-band performance of a quality sufficient to approach the performance of a dual VHF-UHF installation. The common approach has been to modify a present VHF type to a combination VHF-and-UHF type. To date, such an approach has met with only minor successes for many reasons—poor UHF pattern (narrow and difficult to orient and with high susceptibility to reflections), loss of efficiency on the VHF band (loading and tendency to smear), and poor UHF gain (poor standing-wave conditions and erratic results from channel to channel as a function of frequency).

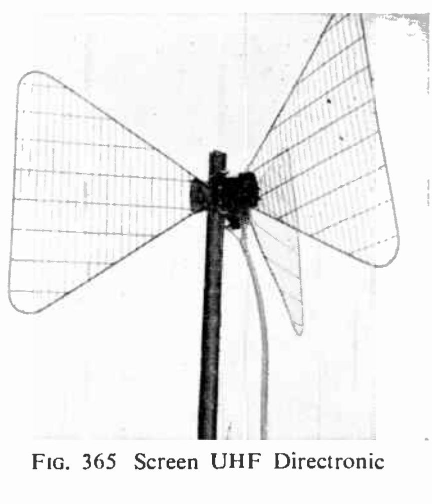


FIG. 365 Screen UHF Directronic

A better approach to the combination VHF-UHF antenna is one complementary to that used. Instead of designing and modifying a VHF antenna for UHF reception, it is possible to redesign a good UHF antenna which will permit VHF reception. In this technique one need make little or no compromise of the antenna performance on the UHF band. Thus the gain and the pattern of the UHF antenna are not disturbed by an alarming amount, and the performance on the UHF band (where performance is at present most critical because of limited propagation range and prevalence of dead spots) is at peak efficiency. The VHF performance need be sacrificed just slightly with respect to the higher-gain VHF antennas. In fact, the VHF performance is, in general, superior to the type of VHF modification that permits UHF reception. Thus a complementary approach permits the development of a combination VHF-UHF antenna that is generally superior to the type antenna modification that evolves from the VHF antenna.

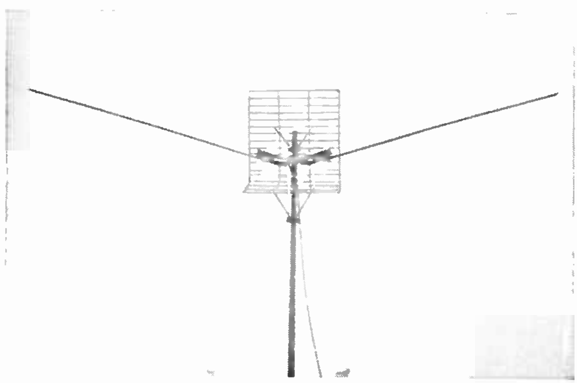


FIG. 365a Snyder VHF-UHF Antenna

A typical VHF-UHF antenna evolved from a good UHF antenna is illustrated in Fig. 365a. In this antenna a standard corner reflector with a bow-tie driven element is used for UHF reception. By properly positioning a single and longer VHF dipole in the corner reflector, VHF reception can be added without sacrificing UHF performance. The success of the addition in terms of minimum influence on UHF reception is to add as little as possible that might disturb the field of the corner reflector. Thus a simple dipole coming straight away from the back of the dipole adds a minimum of disturbance. The addition of other VHF elements that come into the center of the corner reflector, or improper metallic or insulated supports inside the corner reflector or near the point of transmission-line attachment, results in an increasing deterioration of UHF performance. To sustain UHF performance, therefore, the addition of the VHF elements should be done simply and with a minimum of physical disturbance within the corner reflector. The simple dipole addition prevents loss of UHF gain and causes only the addition of some minor lobes to the UHF antenna pattern. When VHF conicals or other basic VHF driven elements are added to the corner reflector using the same method, there is a deterioration in UHF performance.

On the VHF band the antenna functions as a dipole and reflector, but the dipole has an improved bandwidth because of the influence of the taper of the UHF bow-tie driven element and the presence of the close-spaced screen. Thus the bandwidth of the antenna on the VHF band is retained, and performance is comparable to a standard conical. The dipole element is tilted forward to align the high-band lobe and permit peak VHF high-band gain.

The combination VHF-UHF antenna can be used as a simple, metropolitan area corner reflector and single, tilted dipole. This is a simple antenna for metropolitan areas, permitting reception of the powerful VHF channels; at the same time, it has a high gain for reception of the local UHF channels, which can be troublesome even within the supposedly strong signal area of the UHF station. With the addition of the reflector this combination can be used as an ideal antenna, for suburban and near fringe areas, while the same antenna, stacked, can be used for a fringe-area antenna with high VHF gain and superior UHF performance.

TRANSMISSION LINES AND INSTALLATION

The same basic types of transmission lines are used on the UHF band. However, the quality of line and the amount of air dielectric is a more important consideration, since weather and moisture particularly have a more adverse effect on the UHF than on the VHF signal transfer. The conventional 300-ohm line of good quality has a measured loss of only 3 decibels per-hundred-foot in the 500-megacycle range. However, in rainy, damp weather the loss doubles and triples. It is apparent that in rainy, damp weather lines subject to absorption of moisture will be unreliable on the UHF band. Likewise, dirt, salt, or any type of coating deposited on the line over a period of time will result in a substantial signal loss. Thus frequent inspection and replacement of antenna components is advisable.

Good quality 300-ohm tubular line has good UHF characteristics. This type of line has a loss of 3 decibels per hundred but is less subject to weather conditions. Tubular line has a greater air-dielectric spacing between conductors, and attenuation is less influenced by the dielectric material of the line. This will continue to be the most common type of line used for UHF reception. Open-wire line also functions well on the UHF band (2.5 decibels per hundred) but is quite difficult to route, as metallic objects near the line disturb impedance relations or absorb energy.

In fact, all types of line need to be routed carefully. A minimum spacing of four inches from nearby metallic objects should be retained. Even with this separation, resonant lengths in nearby metallic objects can disturb a line. Spacers must be of good quality and low loss—and must not display adverse resonant effects. A sensitive field-intensity meter can be used to advantage in testing UHF antenna positions and for chasing down line defects.

Coaxial lines which can also be used on the UHF band offer good shielding characteristics but rather high attenuation per cost. The common *RG59U* has a loss of some 9 decibels per hundred. Nevertheless, there will be many locations that require this type of line or more expensive coaxial because of its shielding and isolation features. This type of line offers shielding from ignition and electrical noises, although this type of interference has much less influence on UHF reception. In fact, if one can build up the UHF signal to a level comparable to a VHF signal at antenna input, the picture will be less subject to interference than the corresponding VHF signal. The coaxial line in UHF service offers more complete isolation from nearby objects and is less difficult to route.

UHF TRANSMISSION LINE

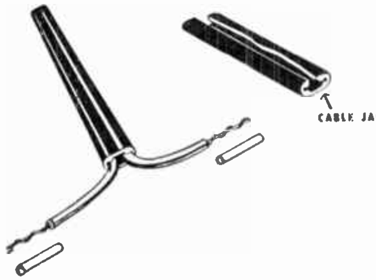
The transmission line is an important cog in obtaining strong and consistent UHF reception. By consistent reception we mean no sharp decline in signal level when the line becomes damp, coated, or wet. The better lines (in the present state of development of dielectric materials) have a large percentage of air-dielectric spacing. To derive full benefit from these better lines this air-separation area must be kept dry and undisturbed by proper sealing off of the line at the ends; proper vent-arrangement is necessary to prevent condensation.

Anaconda has developed a special 270-ohm line for UHF operation. It consists of two polyethylene tubes surrounded by a weather-resistant, outer polyethylene jacket. Inner construction and shape of the jacket permit only a limited-area physical contact (Fig. 366) with the inner tubes (large percentage of air-dielectric separation). Inside each of the tubes is a solid-wire conductor that is held centered by a polyethylene lacer, affording a large percentage of air-dielectric spacing and only a limited surface contact between the conductor and dielectric material.

FIG. 366 Anaconda UHF Line

It is apparent that the physical construction of the line maintains uniform spacing between conductors and offers much air spacing. Line-loss is 3.6 decibels per-hundred-foot at 500 megacycles and 5.1 decibels per hundred at 900. Good protection from weather offered by a line can be attained only by proper sealing, which prevents dampness and condensation from penetrating into the internal air-separation areas. The recommended sealing procedure is illustrated in Fig. 367.

The Amphenol tubular-type line affords a very large percentage of air-dielectric spacing, particularly in the area directly between conductors where this is most significant (shortest path between conductors). Dielectric material



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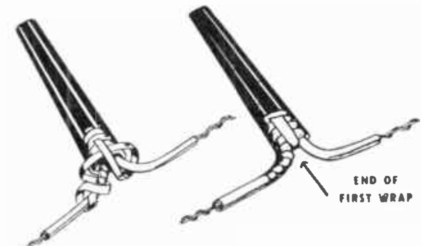
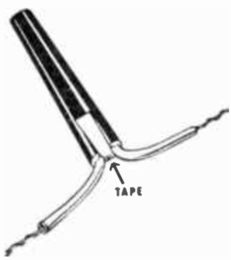
Make a circular cut around, but only partly through the cable jacket 4 inches from the end. Next make a longitudinal cut entirely through the jacket from the initial cut to the end of the cable. The jacket is removed back to the circular cut as shown in step 1.



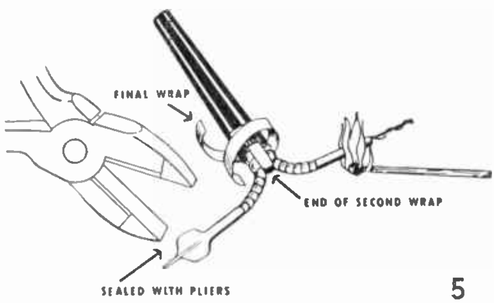
3 4

Next take a 7" piece of tape and begin wrapping it in a clockwise direction one inch below the crotch. When the crotch is reached, pass the tape through the crotch and continue wrapping for ¾" along one conductor. Then wrap down the conductor and through the crotch to near the starting point as shown above in steps 3 and 4. Repeat this for the other conductor, wrapping the tape counterclockwise in this case. The seal is finished as shown in step 5 with a 4" piece of tape.



2

Spread the conductors apart and place two 1" pieces of sealing tape across the crotch as illustrated in step 2. The tape should be applied under tension.



5

Remove one inch of insulation from the conductor ends and with a lighted match soften next inch and press firmly with pliers to complete the seal around the conductor as shown in step 5.

FIG. 367 Sealing Procedure for Anaconda Line

is reasonably tough, and rigid as well, thus allowing for maintenance of uniform spacing and minimization of moisture penetration. The attenuation factor is approximately 3 decibels per-hundred-foot in the 500-megacycle range.

All of these line types must be sealed and looped at the antenna ends to prevent moisture penetration; they should also be elbowed (Fig. 368), and pierced at the point where the line enters a building. The line should be dipped into an elbow or drip loop and dielectric pierced (¼-inch hole) at the lowest point; internal condensation is thus minimized and drainage provided. If the line is allowed to fill with water all the advantages of its construction are sacrificed.

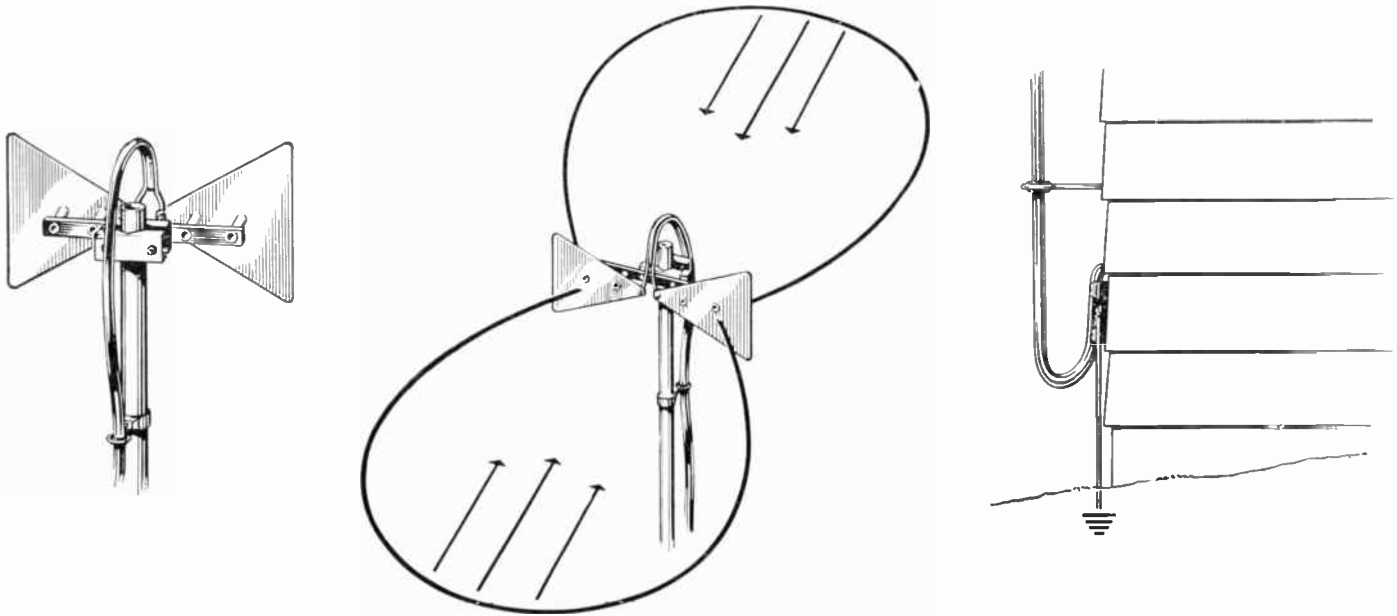


FIG. 368 Amphenol Recommended Line-Installation Procedure

The characteristics of Amphenol tubular line and the influence of weather on various lines are tabulated on the chart below:

Type of Line	*100 Megacycles*		*500 Megacycles*		*1000 Megacycles*	
	Dry	*Wet*	*Dry*	*Wet*	*Dry*	*Wet*
1. Flat 300-ohm	1.2	7.3	3.2	20.0	5.0	30.0
2. Tubular 300-ohm	1.1	2.5	3.0	6.8	4.6	10.0
3. *RG-59/U* Coax	3.7	—	9.6	—	14.5	—
4. *RG-11/U* Coax	1.9	—	5.2	—	7.8	—

Nominal Impedance, 300 ohms
Velocity of Propagation, 34 per cent
Attenuation: Decibels per 100 feet 400 megacycles 2.7, 500 megacycles 3.0, 700 megacycles 3.6, 900 megacycles 4.2.

An improved open-wire line, marketed by Gonset for UHF operation (Fig. 369), has a lower impedance (375 ohms) and less spacing between conductors. Consequently, separation between conductors is a lesser percentage of the UHF wavelength as compared to the usual open-wire line. The electric field is better confined and less disturbed by surrounding metallic surfaces. Loss is approximately 2 decibels per hundred at 500 megacycles.

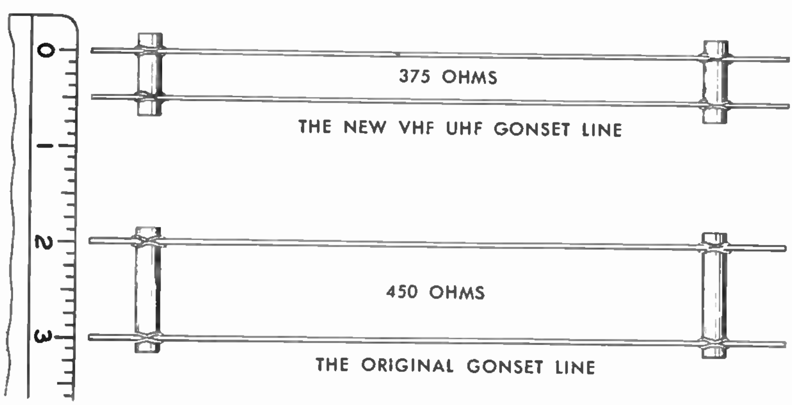


FIG. 369 Gonset UHF Open-Wire Line

In the present state of development of antennas and input circuits, it is advisable to tune the line for peak UHF reception on a given channel. A simple method is to wrap a two- or three-inch piece of aluminum foil around the line and slide it along the line for the best and strongest picture.

Crossover networks are used in VHF-UHF installations to isolate VHF and UHF segments and to prevent one from interacting with the other. For example, with a crossover network it is possible to attach a UHF antenna and a VHF antenna to a mast to connect both antennas to the crossover network and then to a single transmission line leading down to the receiver. Without the network two transmission lines would have to be run to the receiver.

In fact, there are four basic applications for such a network, Fig. 369a. A second application would be to use two networks; one at the antennas and a second at the receiver when the receiver has dual VHF-UHF inputs. When a single UHF-VHF antenna is installed, a single crossover network is connected at the receiver to permit feeding of dual inputs. A network can be used when a dual antenna installation is made and the receiver has a single antenna input for both UHF and VHF reception.

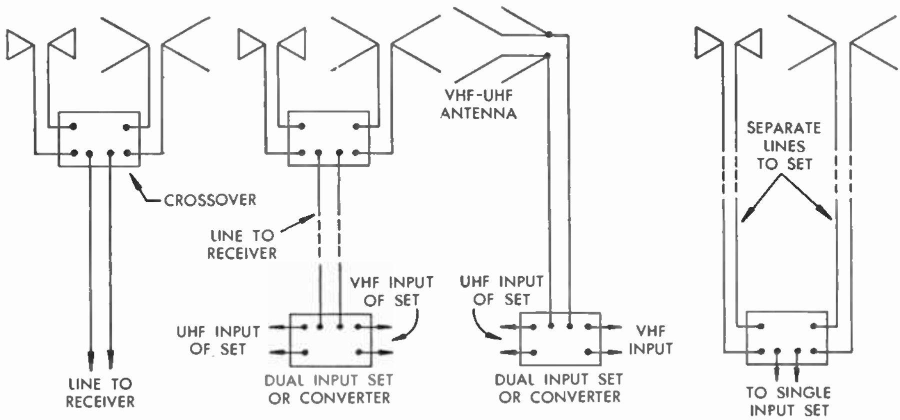


FIG. 369a Applications for VHF-UHF Crossover Networks

The electrical functions of the crossover network are to pass the desired frequency in each leg but to prevent transfer of the frequencies present in the opposing leg without attenuating those frequencies. A simple network would be the reactive type shown in the top drawing of Fig. 369b. The receiver transmission line would be attached to the two center connections, while the UHF antenna would connect to the left and the VHF antenna to the right. Capacitor *C1* would be chosen to have a very small reactance in the UHF band, and UHF signals would pass without attenuation to the line running into receiver.

The capacitors would have a high reactance at the much lower VHF frequencies and would prevent the VHF signals from being lost or attenuated in the UHF antenna system. The inductor *L1* has a low enough reactance to permit the transfer of VHF signals to the transmission line but a high reactance to the UHF signals (reactance of the inductor rises with the frequency) which prevents the VHF section from attenuating the UHF signals. Thus, looking in from the UHF terminals, the UHF signal sees a low-impedance path to the line and a high impedance out the VHF leg. From the VHF side, the VHF signal sees a low impedance to the line and a high impedance out the UHF leg. Likewise, the network must terminate each antenna in its characteristic impedance and match the impedance of the line as well. This is a difficult task over the great range of frequencies required, and attenuation, because of mismatch, is often severe.

In the second arrangement, parallel resonant circuits are used to isolate sections. For example, *C1-L1* is tuned to the VHF range and blocks VHF frequencies from the UHF section, while *C2-L2* tunes to the UHF range and blocks UHF signals from the VHF section. The reactance of the tuned circuit off of resonance is so low that it permits transfer of desired frequencies in each leg to the transmission line. Again it is difficult to retain this resonant condition over such a wide band and at the same time present proper impedance match.

The final method, using high- and low-pass T-filters, is perhaps most successful, due to its ability to retain impedance constant over the wide frequency range. In the left leg the low reactance of the series capacitors and the high reactance of the compensating shunt inductor permit transfer and match of the UHF signals to the line. The lower reactance of inductor *L1* and the higher reactance of capacitor *C1* block the lower VHF signal frequencies, but at the same time, do not upset the match of the VHF signals to the line. The left side is thus a high-pass filter section. The right side functions as a low-pass filter, matching and transferring VHF signals to the line. While it has a high impedance in the UHF range, it does not disturb the match on the UHF side.

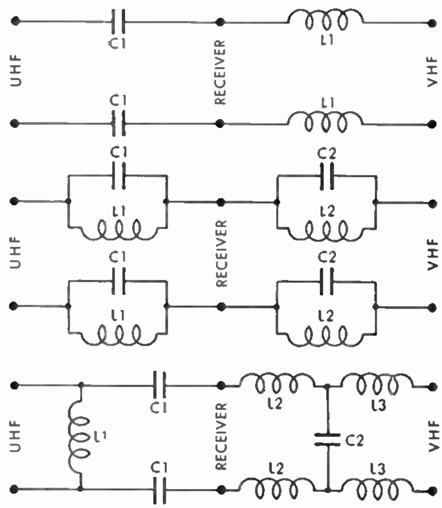


FIG. 369b Crossover Networks

Any crossover network must be small and compact in order to prevent leads and large elements from forming spurious resonant conditions at the very short UHF wave lengths. Thus, printed circuits are common because they lend themselves to small, compact, mounting assemblies with minimum lead lengths.

180. *UHF Conversion Methods*

The preparation of a standard VHF receiver for UHF reception is not difficult. It consists of either replacing a presently unused VHF channel strip with a UHF version (Fig. 370), or adding an external converter that will supply signal to the antenna input of VHF receiver, Fig. 371. Consequently, with a few connections and adjustments UHF reception can be added to a VHF receiver. At many sites it will be far more difficult to overcome the vagrancies of propagation and of antenna performance than it will be to handle the signal after it reaches the end of the transmission line.

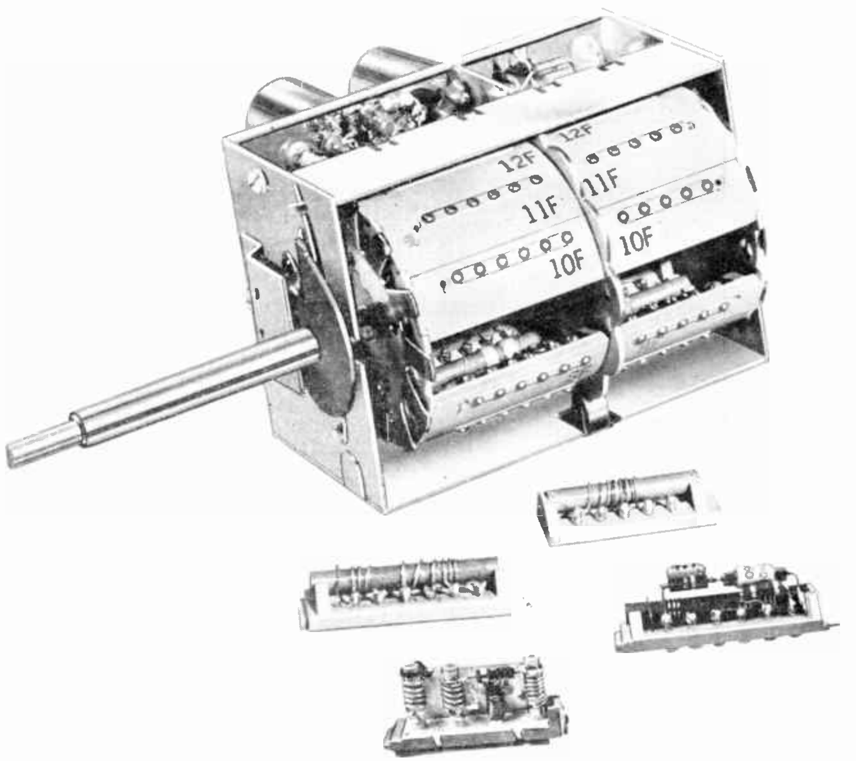


FIG. 370 Standard Coil UHF Strips

UHF PLUG-IN CHANNEL STRIPS

Certain VHF receivers, as a function of the type of tuner employed, can be prepared for UHF reception with a plug-in adapter strip. When a turret-type tuner is used, an unused VHF channel strip can be removed and a UHF replacement strip inserted in its stead. If there is more than a single UHF allocation in the area, additional unused strips can be removed and replaced with UHF channel strips. In general, a separate strip will be required for each UHF channel to be received in a given area.

Instead of vacuum tubes, the UHF channel strip employs two crystals: a crystal mixer and a crystal harmonic generator. The mixer crystal, a high-frequency type, is small in size; it has low capacity and an excellent signal-to-noise ratio, while the crystal harmonic generator is larger physically and a more common type (such as *1N34*). This latter type of crystal is often biased carefully to enhance its harmonic-generating ability.

The operating principles of the UHF plug-in strip (Fig. 372) are first to mix the incoming UHF signal with a UHF local-oscillator component derived

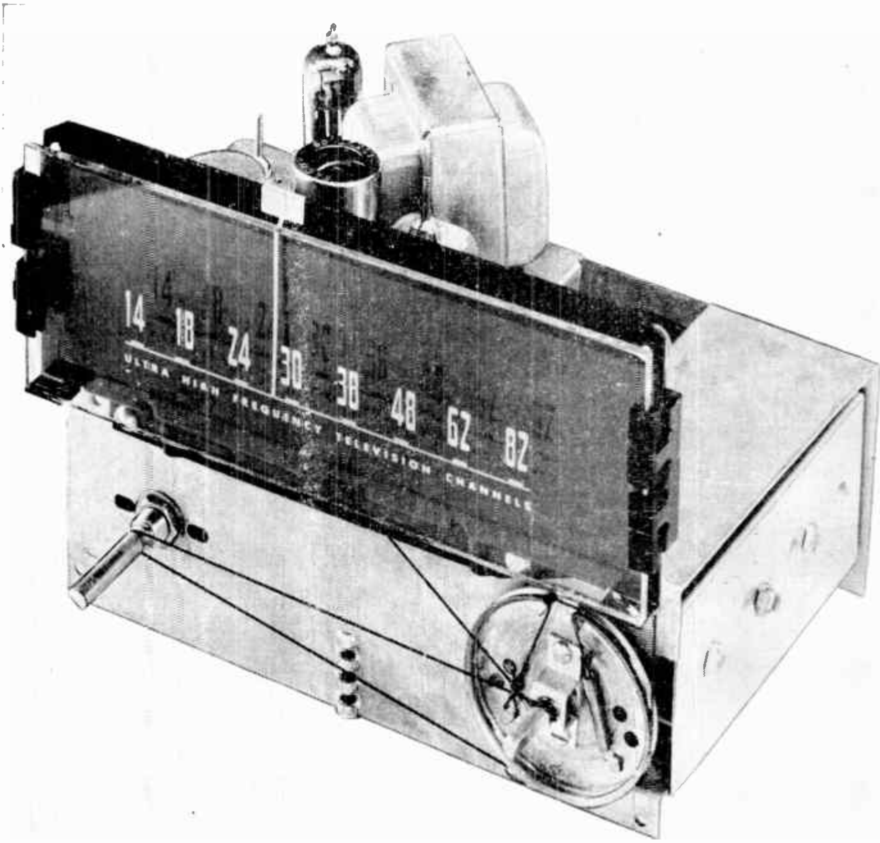


FIG. 371 Mallory External Converter

from the VHF local oscillator through harmonic generation. The difference frequency at the output of the crystal mixer matches the frequency on which a VHF tuner must be set for UHF reception. Thus the r-f amplifier of the VHF tuner acts as a *high-frequency* i-f stage. The signal is mixed again but this time with the fundamental VHF-tuner local-oscillator frequency to produce the standard i-f output range (can be considered the low frequency i-f section of a double-conversion process).

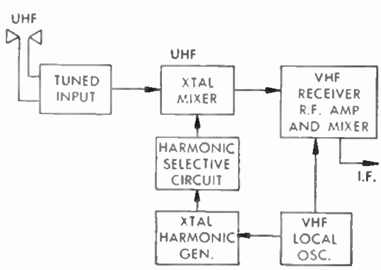


FIG. 372 Block Diagram Plan of UHF Plug-In Strip

A mathematical interpretation will clarify the technique. Let us assume we want to receive a UHF signal on the first UHF channel—picture-carrier frequency of 471.25 megacycles. If a 98.5-megacycle high i-f frequency is decided upon, the UHF local-oscillator frequency would be 471.25 minus 98.5 or

372.75 megacycles. UHF local oscillations must have a frequency lower than the UHF signal frequency in order to obtain correct relative positioning of sound and picture carriers at the output of the VHF mixer. Picture- and sound-carrier frequencies at the output of a UHF mixer are:

$$471.25 - 372.75 = 98.5\text{-megacycle picture carrier}$$
$$475.75 - 372.75 = 103\text{-megacycle sound carrier}$$

This output is applied to the r-f amplifier input of the VHF tuner. The tuned circuits of the r-f amplifier and mixer have previously been set on this frequency by the VHF coils of the UHF plug-in strip. Thus the r-f stage acts as a high-frequency intermediate amplifier. At the VHF mixer the signal is mixed with the VHF local-oscillator frequency to produce standard i-f frequencies of the VHF receiver or (in megacycles):

$$124.25 - 98.5 = 25.75 \text{ picture carrier}$$
$$124.25 - 103. = 21.25 \text{ sound carrier}$$

Observe that the UHF local oscillations represent the third harmonic of the VHF local-oscillator fundamental frequency. It is the function of the crystal harmonic generator to develop a strong third harmonic from the fundamental local-oscillator frequency. An arrangement such as this, employing a VHF harmonic for UHF-mixing, means that the UHF strip requires no separate UHF oscillator tube and associated circuit components.

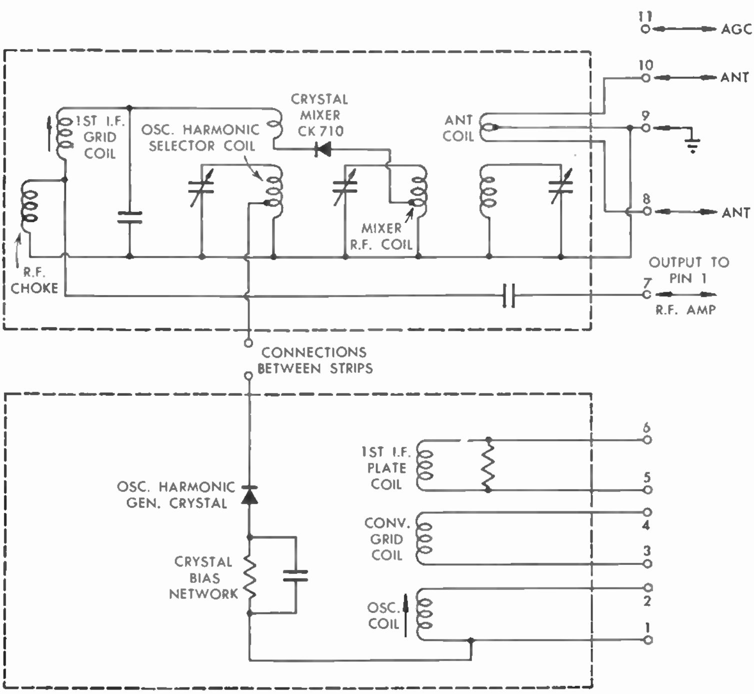


FIG. 373 Schematic of Standard Coil Plug-In Strip

181. *Commercial* UHF *Plug-In Strips*

The UHF channel strip arrangement can be used with the Zenith turret tuner or any receiver that employs a Standard Coil tuner, Fig. 373. The Zenith plug-in strip has three basic functional sections. There is a UHF section consisting of input filter, double-tuned input transformer (coupling between antenna and crystal mixer), and UHF crystal mixer.

The UHF antenna system is coupled into the strip through a loop (Fig. 374), at the low-impedance end of the primary of the double-tuned transformer. Consequently, an impedance match and voltage step-up are obtained through transformer action. The double-tuned transformer between the antenna and mixer permits peak selectivity, high Q, and excellent off-frequency signal rejection. The antenna, mixer, and multiplier resonant circuits are completely shielded in individual cavities, thus minimizing spurious resonances and direct feed-through of undesired signals. Coupling between the two sides of the double-tuned transformer is a mutual inductance in the form of a small cylindrical bushing at the low-impedance ends of the resonant windings. No resistive damping is required, as suitable bandwidth is attained with the damping influence of the antenna a crystal mixer.

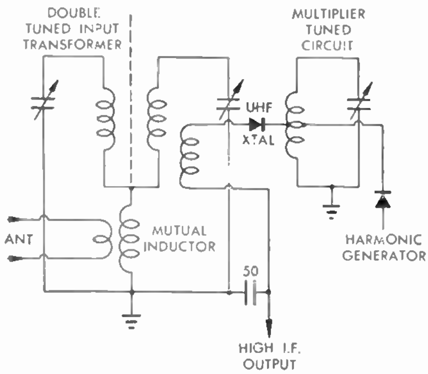


FIG. 374 UHF Section of Zenith Strip

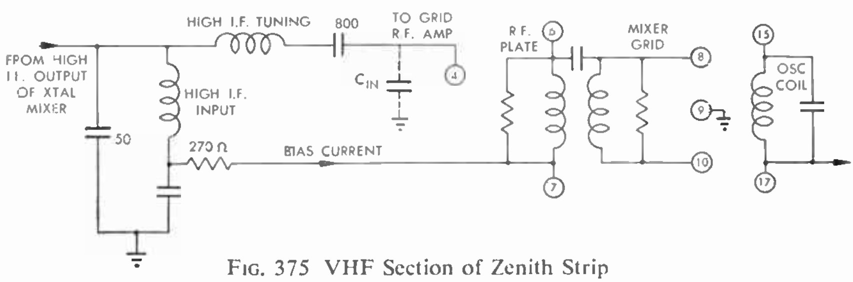


FIG. 375 VHF Section of Zenith Strip

A small loop links the secondary or mixer-tuned circuit with the crystal circuit, while the mixer circuit also includes a small portion of multiplier inductor, serving as a means of introducing UHF local oscillations.

The i-f output of the crystal mixer is coupled into the VHF section of the tuner through a low-pass section that filters out UHF variations. Signal is applied to the grid of the r-f amplifier and is amplified by this stage, the VHF-tuner r-f amplifier plate and mixer grid circuits being resonated to the i-f frequency of the crystal mixer output. Inductors of these resonant circuits are mounted on UHF strip, Fig. 375. These high-frequency i-f and VHF local-oscillator signals heterodyne to produce a still lower frequency i-f that corresponds to the i-f range of the VHF receiver.

The VHF local oscillator is also the fundamental source or signal for the UHF harmonic generator section of the tuner strip. Excitation is obtained from the plate circuit of the VHF oscillator, Fig. 376, and is applied through an r-c time constant to the multiplier crystal. A time constant biases crystal with the result that its current flows only on crests of excitation, producing crystal current bursts that are rich in harmonics. These current bursts excite the multiplier resonant circuit which is tuned to a harmonic of the fundamental repetition rate. Thus a harmonic sine wave is formed that serves as the injection signal for the UHF crystal mixer.

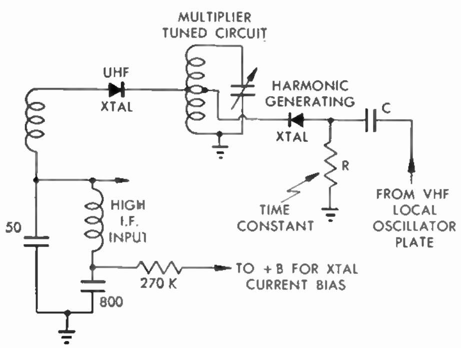


FIG. 376 Harmonic Generating Section

To produce a strong UHF harmonic component, some biasing of the crystal is helpful. Bias current, as obtained through the *270K* resistor to plus B, selects an operating point near the maximum curvature of the crystal characteristic. Nonlinearity at this point accents harmonic levels.

The Standard Coil plug-in strip has the same general plan. Note that the terminals on the right (Fig. 373) correspond to the numbered terminals of the Standard Coil tuner. The antenna is connected through terminals 8 and 10. The antenna coil is coupled to the primary of the double-tuned mixer transformer, while the crystal mixer taps off a low-impedance point of secondary. It is also coupled to the harmonic-selector resonant circuit, thus obtaining UHF injection signal.

The mixer output circuit is resonated to the high i-f frequency, and the i-f signal is applied to the grid of the r-f amplifier via terminal 7. The r-f amplifier plate, mixer grid, and VHF local-oscillator circuit inductors are also a part of the VHF strip. The local oscillator signal must be applied also to the crystal harmonic generator through the r-c biasing network, the proper harmonic being selected by the UHF resonant circuit to cause development of an exciting UHF sine wave for the UHF crystal mixer.

182. *Basic* UHF *Converter Types*

The most common type of UHF converter (Fig. 377) consists of a crystal mixer, UHF local oscillator, and single-cascode i-f amplifier. At present this type of converter permits the production of an economical and effective UHF unit, capable of tuning over the entire UHF band. It can be attached to all types of VHF receivers. A second possibility, like the plug-in arrangement, would employ a system of harmonic generation, using a low fundamental local-oscillator frequency and proper harmonic mixing or doubling in order to derive the UHF local frequency.

R-F amplifiers will not be immediately prevalent because of the expense involved in establishing a suitable signal-to-noise ratio using vacuum tubes in the UHF band. A crystal mixer without preceding r-f stage lends itself well to converter application because of its favorable UHF performance. A crystal mixer has a low-noise content and requires only a low local-oscillator injection level. Therefore, a high signal-to-noise can be retained and the local oscillator radiation problem minimized.

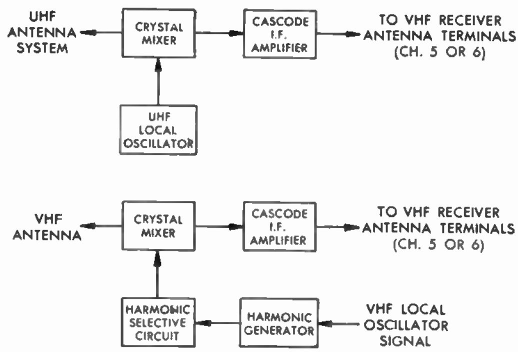


FIG. 377 Basic UHF Converter Methods

A crystal mixer attenuates, rather than amplifies, an applied signal. The usual vacuum-tube mixer, although it does not have as high a possible gain as when employed for straight amplifier use, at least has some gain when connected as a mixer. The crystal mixer introduces a 6- to 9-decibel loss, the signal level at the output of the crystal mixer being substantially less than the input signal level from the antenna system. It is significant that the output frequency is also much lower, however, and is easier to establish favorable signal-to-noise relations.

Insofar as vacuum-tube operation is concerned, the higher the frequency of operation, the higher will be the noise content of a given tube and the poorer will be the signal-to-noise ratio. Consequently, if the signal can be introduced to the first vacuum tube at a low frequency, a better ratio can be obtained. Inasmuch as signal output is weak at the crystal mixer (and crystal mixing has not added a high noise level), the signal-to-noise ratio of the converter is set by the first i-f tube. A satisfactory ratio can be obtained because of the lower applied frequency. If an r-f amplifier stage or vacuum-tube mixer were employed, it would be more difficult and more costly to obtain a comparable ratio because of the much higher signal frequency.

To summarize, a higher signal-to-noise ratio per cost factor can be obtained by introducing a lower frequency signal to the first vacuum-tube circuit. Thus, vacuum-tube amplification or mixing in the UHF range is not used. Instead, a low-noise crystal mixer is used to reduce the signal frequency before it is introduced to the first vacuum-tube circuit at a much lower frequency. To obtain the very best signal-to-noise relation with this system, the i-f stage must be designed with care and in such a way as to keep hum and noise level at a minimum. The new cascode type of amplifier with its low-noise content and effective shielding is common in UHF converters.

Local-oscillator radiation is not as trying a problem as in the VHF band, despite the absence of an r-f amplifier stage. The crystal mixer, in addition to its attenuation characteristic, also requires much less local-oscillator excitation than a corresponding vacuum-tube mixer. Furthermore, there is a greater frequency separation between the signal and local-oscillator frequency, and the local-oscillator component is not as likely to pass out through the frequency-selective input circuits to antenna.

CONVERTER OUTPUT AND SWITCHING SYSTEMS

The output signal frequency of the UHF converter is in the channel frequency range of the present VHF receiver, and a converter can be attached to a receiver much as we attach a booster today. Antenna or antennas are attached to the converter, while output of the converter is applied to the antenna input terminals of the VHF receiver. The most common output-frequency range will be over the spectrum of channels *5* and *6* with the output ranges of channels *2* and *3*, channels *12* and *13*, or channels *9* to *12* present for some converters.

Two adjacent channels are chosen to permit the converter output frequency to be set to whichever channel of the two is unused in a given area. For example, in New York with a channel *5* station in operation, the converter output would be set on channel *6*; in Philadelphia, with a channel *6* station in operation, the converter output would be set on channel *5*. The receiver is set on this pre-set channel whenever UHF reception is desired.

A switch is included with which to switch the converter output to the receiver antenna terminals when UHF reception is wanted. At the same time the UHF antenna is attached to the input of the converter. The same switching arrangement removes the converter output from the VHF receiver antenna terminals and applies the VHF antenna system to these terminals for normal VHF reception.

Some other systems for UHF converters (Fig. 378) will be the use of an r-f amplifier-preceding crystal and/or a vacuum-tube mixer for more expensive types. Amplification in the UHF frequency range requires more critical design, expensive components, and higher costs. If more sensitivity, selectivity, and image-rejection are desired, the second method is preferred. This system, using a double-superhet technique, has a first mixer-oscillator that brings signal frequency down into the 150- to 250-megacycle range. The signal is then amplified in a selective multi-stage i-f strip before its presentation to a second mixer-oscillator. The output of the second mixer has the same frequency range as those of low video channels (*2–3*) or the i-f frequency range of a VHF receiver. For the latter arrangement, actual VHF-UHF switching is done at the input of the i-f strip of the VHF receiver, keeping the VHF and UHF tuners separate from each other. This system might at first appear elaborate, but with small component parts the unit can be mounted on a subchassis that can be attached to the main chassis of a VHF receiver.

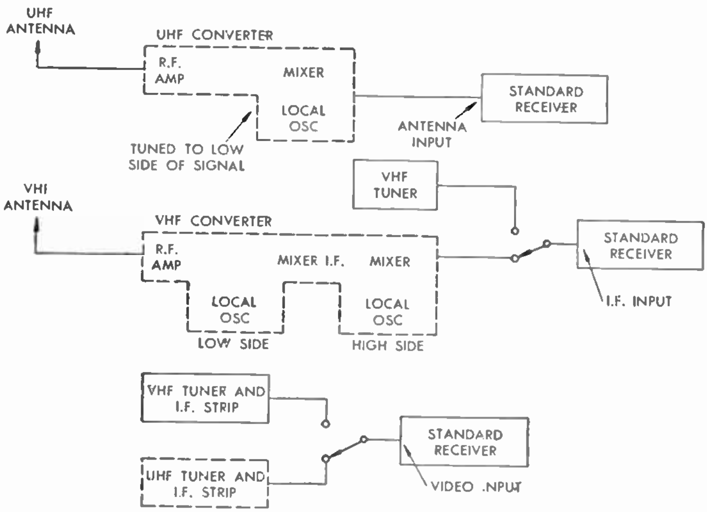


FIG. 378 Additional UHF Conversion Systems

Still another possibility is that of a complete UHF tuner and i-f strip assembly with the switching done at the input to the receiver's video amplifier. These latter two systems may be found in some factory-built combination VHF-UHF model receivers.

183. UHF *Tuned Circuits*

A number of problems need to be overcome in the development of a tuner-converter for the UHF 470- to 890-megacycle spectrum. The scope of these difficulties is apparent when we consider that this span of frequencies contains 70 television channels, bringing the total count of VHF and UHF channels to *82*. Certainly a 70- to 82-selector switch arrangement presents problems. Continuous tuning with vernier adjustment is a difficult problem to overcome over this range of operation. To add to the difficulties, the frequency range is somewhere between the end of the range at which lumped constants can be made to operate efficiently and the range at which microwave practices can be introduced. Choice of tuned circuits and tubes is critical. To obtain the necessary frequency range, tuned circuits often take on rather peculiar shapes.

The task of tracking a number of these tuned circuits is present. One-knob tuning of the various resonant circuits requires mechanical ingenuity. Absence of an r-f stage and use of a crystal mixer have helped to simplify the tracking problem.

The factors that restrict the use of standard miniature tubes and lumped constant-tuned circuits are tube capacities, lead inductances, and transit time. Tube capacities, small as they may be, present a very low reactance on the UHF band—for example, a 5-micromicrofarad capacity has a reactance of just approximately 40 ohms at 800 megacycles. A small lead inductance has a high reactance on the UHF band. These two factors, therefore, set and limit the highest frequency to which any externally attached resonant circuit can be tuned. Acorns and sub-miniature tubes show possibility because of their smaller parameters (size, capacity, and inductance). The 6F4 acorn and its miniature tube base counterpart, the 6AF4, are common in the initial lines of UHF converters.

As far as influence of transit time is concerned, it places a low-impedance resistive-capacitive shunt across the grid input of the tube, limiting the highest frequency at which a tube can oscillate or function as an amplifier. Transit-time loading is caused by in-phase (resistive) and leading (capacitive) components of the grid current. This current flow is the result of the finite time of travel of electrons flowing over the cathode-grid-plate path in the tube. Ordinarily this time of travel is insignificant, but at the UHF frequency it does become appreciable with respect to the period of the UHF wave. Consequently, flow of current within the tube cannot follow coincidently with grid voltage changes, and grid current flow results.

TUNED-CIRCUIT TYPES

The UHF resonant circuit is a critical combination in establishing efficient UHF operation. There are a number of basic types.

1. Perhaps, initially, the most common type of resonant circuit will be the quarter-wave shorted section of line, Fig. 379. Parallel resonance at a given UHF frequency is attained by moving a shorting bar along the line. A mechanical arrangement permits simultaneous movement of the shorting bars on the two or three tuned lines needed in a typical UHF converter.

Dimensions of lines must be chosen in such a way as to permit operation over the wide span of frequencies. The line lengths and lead inductances determine the total inductance of the resonant circuit, while the total capacity is a function of lines, circuit components, and tube capacities. To permit tuning to the very high end of the band, the total capacity must be held to a minimum. This circuit can then be brought to resonance with a satisfactory value of inductance (though small) for establishment of a suitable L to C ratio and circuit Q at the high end of the band, Fig. 380. Consequently, the

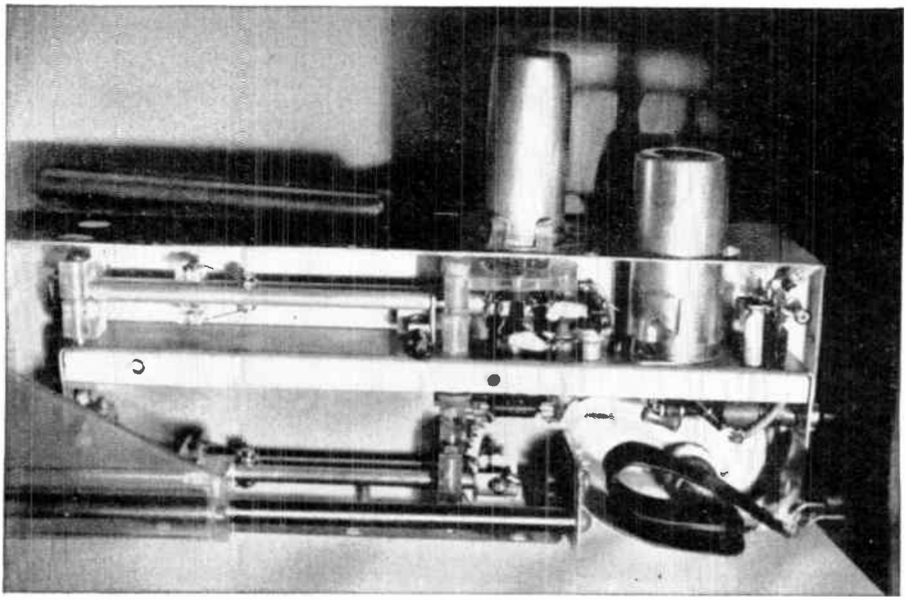


FIG. 379 GE Tuned Lines

high-end frequency limit can be tracked on each line by adjusting for the minimum inductance required to bring the line to resonance with whatever circuit capacity exists. Low-end frequency limit is attained by using over-all length of line and whatever added capacity is needed in order to reach the low-frequency point with suitable Q and bandwidth.

Another factor in the use of lines as resonant circuit elements is the local oscillator which must operate on a frequency lower than that of the mixer and/or antenna input circuits. Lower-frequency range operation can be obtained by changing dimensions of the line, or if for mechanical reasons it is desired to keep line dimensions uniform, a tunable capacity can be added across the line. High- and low-frequency end adjustments must also be incorporated if proper oscillator tracking and calibration are to be established.

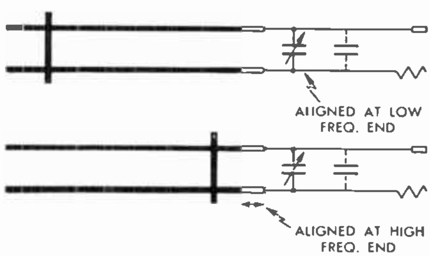


FIG. 380 Tuned Lines as Resonant Circuits

Tuned lines can also be formed into an arc instead of being stretched out in a straight line. This expedient conserves space and serves as a convenient means of ganging and tracking a number of tuned UHF circuits. In this arrangement (Fig. 381) a shorting contact at the end of an insulated arm slides

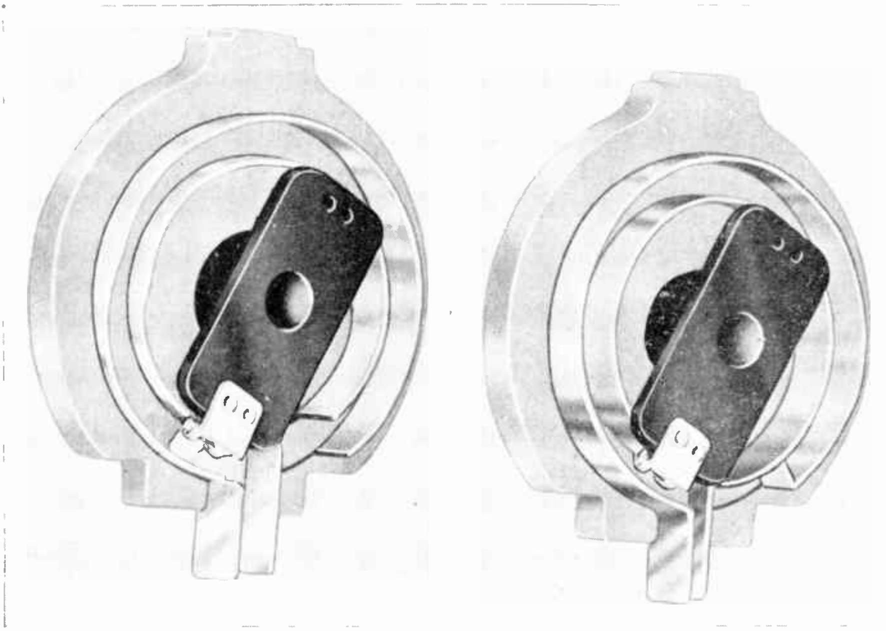


FIG. 381 Mallory UHF Tuned Lines

along the circular lines. In a typical converter, three of these tuned sections with associated arms and shorting contacts are attached to a single shaft and provide for continuous tuning over the UHF band.

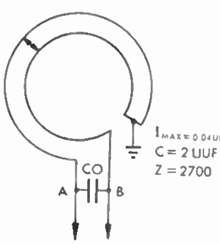


FIG. 382 Tuned Lines in Form of Arc

A typical resonant circuit (Fig. 382) of this type has an induction of 0.04 microhenry and a capacity of 2 micromicrofarads or a self-resonant frequency of approximately 530 megacycles—with just one micromicrofarad of added circuit capacity, the low-frequency minimum becomes approximately 460 megacycles near the low-frequency limit of the UHF band. Proper frequency range and tracking are attained by controlling the width and shape of the strip-tuned lines of each resonant section. Minor tracking and bandwidth adjustments can be made with small external capacitors.

2. The *butterfly*-tuned circuit, so named because of its appearance, is an effective high-frequency tuned circuit, because as the rotor is turned both the inductance and capacitance of the resonant circuit change. Thus a more uniform *Q*, bandwidth, and positive tracking can be established. The stator plates of the butterfly (Fig. 383) act as an inductive turn or loop, while the rotor acts both as a variable capacitor and an inductance shunt. Consequently, as the rotor moves the capacity, as well as the inductance of the stator, changes.

In the first position the inductance is minimum, since the rotor almost completely closes the inductance loop. For example, a flat washer-shaped ring, as opposed to a turn of thin wire with the same circumference, would have substantially less inductance. Capacitance is also minimum in this position, since there is very little capacitive linkage between the rotor and stator. This position, then, represents the highest resonant frequency of the butterfly. When the rotor is in a position directly over the stator, as in Fig. 383, the resonant frequency is minimum. In this position there is maximum capacitive linkage between the rotor and the two sides of the stator. At the same time, the air space in the stator has been opened, and there is maximum inductance. The center drawing of Fig. 383 shows an intermediate frequency point. With this type of tuned circuit, both capacitance and inductance decrease with tuning toward the high-frequency end, such decrease permitting a greater frequency range with a uniform Q and bandwidth. Another advantage of this type of circuit is the absence of moving contacts, noise, and positive contact problems.

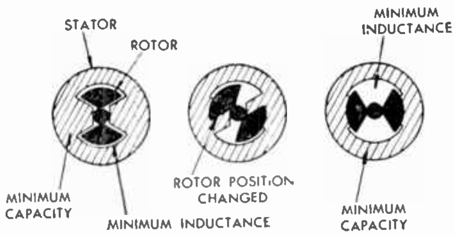


FIG. 383 Butterfly-Tuned Circuit

3. A modified or semi-butterfly has been used by DuMont in an experimental UHF converter. With a semi-butterfly, too, the stator acts as a single-turn inductor, Fig. 384, while the rotor changes circuit capacity as well as inductance of the stator. As illustrated in drawing A, the stator is a single-turn coil with an open air space at the bottom and a definite capacitive linkage between the rotor and two sides of the stator (each side of slot). This is the maximum inductance, maximum capacitance, low-frequency position. When the rotor is moved through 180 degrees, drawing B, it closes in the air space, increasing the effective surface area of the stator turn and lowering the inductance. Likewise, there is a minimum capacitance linkage between two sides of the stator and the rotor in this position. This is the high-frequency limit of the tuned circuit. Other positions of the rotor tune to intermediate frequencies.

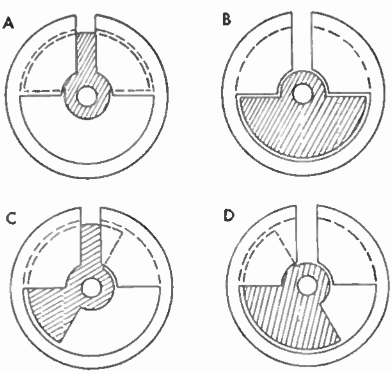


FIG. 384 Basic Semi-Butterfly

To lower the minimum inductance further, the surface area of the stator turn can be increased until it forms a cylindrical shape, Fig. 385, thus raising the high-frequency limit of the tuned circuit still higher. When the top of the cylinder is flattened a mounting position for the tube socket and associated components is provided.

4. Still another type of tuned circuit is the cylinder type, Fig. 386. It consists of two slotted cylinders, one rotating inside the other. Cylinders represent a simple means of circuit mounting, too, as tube and components can be attached across the slot of the outer cylinder. As indicated schematically, Fig. 387, one side of the slot is connected to the grid, the other side of the slot, capacitively-coupled to the plate.

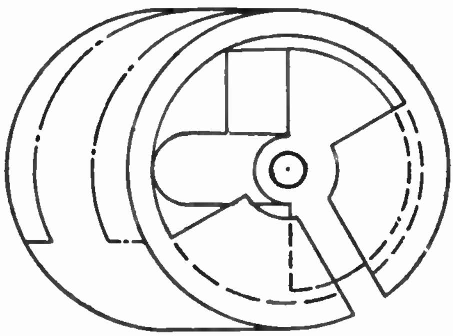


Fig. 385 DuMont Cylindrical Semi-Butterfly

In a cylinder circuit, frequency is maximum when two slots are coincident. In this position there is minimum capacity across the slot of the outer cylinder, such capacity permitting a very high-frequency limit because of the very low inductance of the cylindrical turn. As the inner cylinder is rotated a capacity exists between the outer cylinder and the inner cylinder. The inner cylinder now spans across the slot of the outer cylider, introducing a capacity across this slot. In effect we have two capacitors in series across the slot—each side of the slot to the inner cylinder. The effective capacity across the slot is highest when the inner slot faces exactly away from slot of outer cylinder. This position represents the low-frequency limit of the tuned circuit.

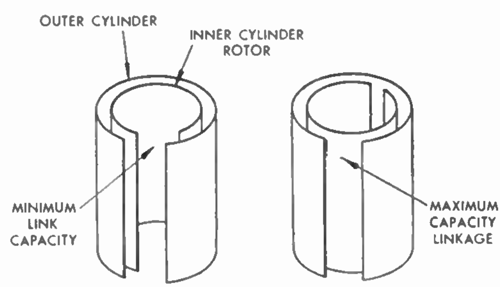


Fig. 386 Cylinder-Tuned Circuit

We can expect to find any one of these basic types of tuned circuits appearing in UHF converters, UHF test equipment, and eventually in combination UHF-VHF receivers.

5. In some of the RCA experimental models, a long, thin cylinder (¼-inch diameter) is used as a tuned element, Fig. 388. This section is largely inductive, resonating with external capacities and also its own end capacity at operating frequency. The element is tuned by a brass core that changes the element inductance and the end capacitive reactance as it moves into the cylinder. The cylinder itself is a paper-base bakelite tube with an inside diameter of 1.251 inches to which has been cemented strips of 2.5-mil copper foil.

To permit adjustment, the copper core is attached to a piece of kovar wire (minimizes drift due to thermal expansion) which has been broken up into short segments by insulating glass beads in order to prevent resonant suck-outs. Tuned elements of this type can be ganged together to permit tuning of the various resonant circuits of the converter.

6. As far as UHF oscillators are concerned, a new approach (developed separately by Pettit and Pan) permits operation at a substantially higher frequency. Customarily we use a parallel-resonant circuit so tuned as to present an inductive reactance to match the capacitive reactance of the tube (dominantly C_{gp} for the usual ultra-audion oscillator) at the oscillating frequency. Eventually, with just a short circuit attached externally, this tube would be operating at its highest or self-resonant frequency. Inductance at this frequency consists of tube plate and grid lead inductances.

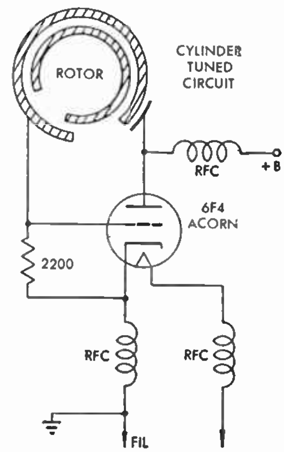


Fig. 387 UHF Oscillator with Cylinder-Tuned Circuit

If the external inductive parallel-resonant circuit is replaced by a capacitor, Fig. 389, a series-resonant circuit is formed (with lead inductances); this effectively reduces the equivalent inductive reactance and allows operation at frequencies higher than the self-resonant frequency of the tube. As far as the tube is concerned, it still "sees" a parallel-resonant circuit—the external capacitor merely tuning the lead inductances which, though still parallel to C_{gp}, are reduced in effective value.

INPUT SELECTIVITY

The input selectivity of the UHF device is an important operating characteristic, as it determines the degree of sensitivity the UHF device has to the desired signal and also its ability to reject interference. Possible sources of interference that can reach the crystal mixer circuit of the converter are image interference, VHF harmonics, higher harmonics of the local oscillator frequency, and multiple harmonics of the insert method of conversion, which employs harmonics of a lower-frequency VHF oscillator. The amount of local oscillator radiation is also influenced by the input selectivity of the UHF device.

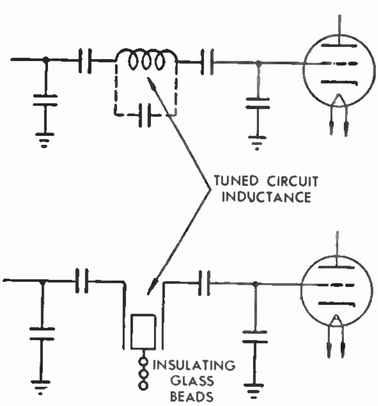


Fig. 388 Reactance Section as Tuned Element

In most of the double-conversion UHF units, the UHF local oscillator tunes lower than the applied signal frequency. Nevertheless, the converter has some sensitivity at a signal frequency point which is below the UHF local oscillator frequency by the amount of the i-f frequency. For example, with the converter tuned to some frequency at the high end of the UHF band there can possibly be interference with another signal having a frequency at the low end of the UHF band (an interference signal frequency that would be the local oscillator frequency minus the high i-f frequency). It is also possible to have interference from a signal that beats with the second harmonic of the UHF local oscillator frequency. In this case, when a UHF signal at the low end of the UHF band is received, it is possible to have interference from some signal at the high end of or outside of the UHF band that is beating with the second harmonic of the UHF local oscillator and producing an i-f frequency output. A few sample frequencies demonstrate this condition—for example, if a 500-megacycle signal is received and the high i-f frequency is 150 megacycles, the local oscillator would have to be set on 350 megacycles. The second harmonic of this local oscillator frequency is 700 megacycles, and it is possible that an interfering signal of 850 megacycles (high end of UHF band) can produce an interfering 150-megacycle i-f signal. In fact, in testing one particular converter on a UHF channel at the very high end of the UHF band it was found that a stronger picture was obtained when the converter was tuned to the very low end of the UHF band, at which point the second harmonic of the UHF local oscillator frequency was supplying the injection signal for the mixer.

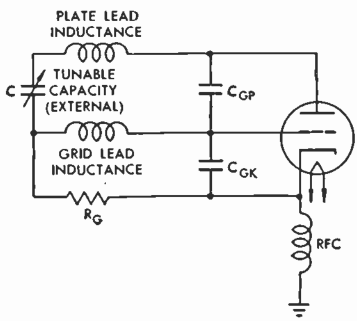


FIG. 389 Series-Tuned Oscillator

A strong source of interference in the UHF converter is the harmonics of the low-frequency oscillator—the harmonics of the VHF local oscillator that feed into the input circuit of the mixer through the antenna input system or via the back door through the high i-f amplifier section. When a channel 5 high i-f frequency is employed, there are four to five harmonics of the VHF local oscillator that fall within the UHF band. Likewise, there is an equal number of image interference points from the same VHF harmonics, and consequently, there are almost a dozen interference points in the UHF spectrum from the VHF source. These points can be located by tuning any UHF converter over the UHF band, and noticing the frequencies at which hum or blocking occurs on the VHF receiver screen.

When a crystal harmonic generator is used for double-mixing action, as it is in many of the insert UHF devices, it is possible that a harmonic, other than that desired for a given channel-mixing, can beat with interfering signals to produce the i-f frequency employed. Thus, these many possible sources of interference stress the importance of the selectivity of the UHF device if the desired signal is to dominate. Likewise, to prevent interference between UHF installations, it is important that local oscillator radiation be minimized. Inasmuch as r-f stages are not employed in many UHF devices, the input selectivity, to a degree, also determines the amount of local oscillator radiation. When vacuum-tube mixers are employed in the UHF band, peak selectivity and use of an r-f stage are mandatory for prevention of local oscillator radiation, because of the much higher injection level required by a vacuum-tube mixer.

The ideal UHF device then includes a selective input circuit with a sharp skirt but adequate bandwidth, a high-pass filter input circuit to prevent the entrance of VHF sources of interference at the input circuit, and a low-pass filter in the exchange path between the output of the UHF device and the input of the VHF receiver, which prevents the exchange of high-frequency harmonics of the VHF local oscillator (no frequencies higher than the high i-f frequencies). A typical high-pass filter for the input circuit is illustrated in Fig. 389a. Choice of i-f frequency, whether it be high or low, influences the design of the unit. For example, a high i-f frequency (in the high-band VHF range) has specific advantages and disadvantages, as compared to a lower i-f frequency (low-band VHF range or 45-megacycle i-f range). When a high i-f frequency is chosen, the unit can be designed to have somewhat better off-band rejection. Response can be better controlled because of the greater frequency separation between the desired signal frequency and image frequencies. The high i-f frequency, however, involves a higher noise content as compared to a lower-frequency i-f range. It is more difficult to obtain the same degree of image-rejection with the lower i-f frequency; at the same time, because of the much lower UHF local oscillator frequency used with the higher i-f frequency, it is more likely that the second harmonic of the local oscillator can create second-harmonic interference. However, the lower UHF local oscillator frequency presents less of a stability problem because of its lower frequency of operation. Likewise, the greater separation between the local oscillator frequency and the signal frequency minimizes the local oscillator radiation problem. Thus choice of the i-f range is a series of pros and cons; performance of various types can be almost equalized by emphasis on proper design as a function of the i-f frequency used.

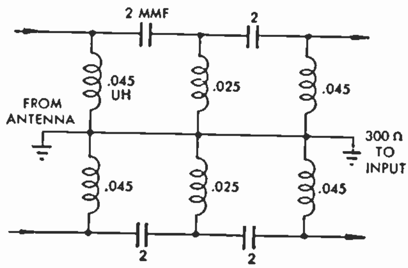


FIG. 389a High-Pass Filter to Block VHF Interference

Choice of single- or double-conversion for UHF devices also presents a series of advantages and disadvantages. With a single-conversion system and only one local oscillator the problem of interference from local oscillator harmonics is eliminated. However, the lower frequency i-f that must be used with a single-conversion system necessitates greater emphasis on proper image-rejection and minimization of local oscillator radiation. Likewise, with a high-gain single-frequency amplifier that is to have a sensitivity comparable to that of a double-conversion system, greater stress must be placed on preventing oscillation and instability in the high-gain single-frequency i-f amplifier.

184. UHF *Crystals and Tubes*

There are two basic types of UHF mixers: germanium and silicon. The silicon crystal has a lower insertion loss and, consequently, a more uniform and somewhat higher sensitivity. The germanium crystal is less expensive and perhaps a bit more rugged, but, in the present state of the science, has a higher insertion loss and less uniformity. However, improved production techniques are bringing germanium performance up to a higher and more consistent level. Although the silicon-type diode (*1N82, 1N21A-B*) is more subject to acci-

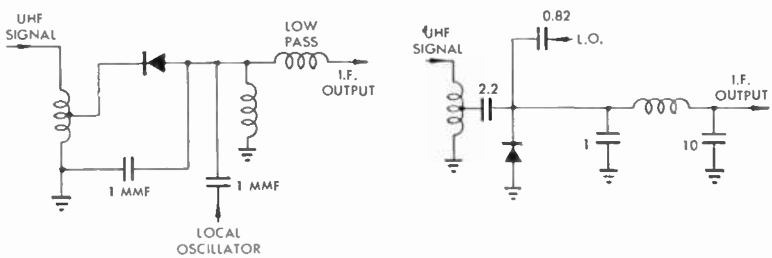


FIG. 389b Crystal Mixer Circuits

dental burn-out by application of excessive voltage, it has a somewhat lesser insertion loss as compared to the germanium crystal (*1N72, CK-710*). This insertion loss of a crystal mixer of approximately 50- to 75-per cent means that the i-f output signal of the crystal mixer is substantially less than the applied r-f signal at the input of the crystal mixer. Therefore, the crystal mixer and the first i-f amplifier stage must be carefully designed to obtain the very best noise factor.

A crystal-mixer stage, Fig. 389b, can be connected in a shunt or series arrangement. To obtain the very best signal-to-noise ratio and sensitivity, the input circuits must be matched to input impedance of the crystal, and the output impedance of the crystal must be matched by the i-f resonant output circuit. Input impedance to the crystal diode is in the 250- to 300-ohm range, while output impedance varies between 300 and 450 ohms, impedance being a function of local-oscillator injection level. Thus crystal injection level not only influences the sensitivity of the mixer, but its bandwidth as well. Injection alignment controls, which are common in newer UHF devices, establish an optimum level over the UHF spectrum. In fact, the injection level does vary input and output impedances, conversion loss, and the noise factor. It is apparent that maintenance of a constant and proper injection-level over the entire UHF spectrum is a difficult problem in design of UHF devices. Consequently, it is common to find that the relative sensitivity between two converters varies from channel to channel as a function of injection level.

For peak conversion efficiency the proper level of the local oscillator signal must be loosely coupled to the crystal mixer. A crystal d-c current flow of 0.5-to-1 milliampere indicates proper injection (biases crystal on the most sensitive part of its characteristic). In UHF practice, the output side of the mixer stage functions as an i-f selective circuit, a filter for r-f and local oscillator UHF components, and a low-pass filter to prevent the exchange of high-frequency harmonics from the VHF local oscillator into the UHF mixer.

In summary, the advantages of the crystal mixer are: high signal-to-noise ratio in the UHF spectrum, a low input capacity (less than 1 micromicrofarad), no contact potential to bias the stage with relation to the signal level improperly, and no heater supply, whose lack permits simple design without the difficulties of a-c hum and high r-f losses.

UHF OSCILLATOR

Two most important considerations in the planning of the local oscillator stage are stability and constant injection voltage to the mixer. A third factor is the combination of proper shielding and minimization of local oscillator radiation. Minor circuit changes have an appreciable influence on local oscillator frequency. For example, if the local oscillator tube is not seated properly with clamp and shield, a shift in its physical position can produce a multichannel shift in frequency. Even with a good local oscillator tube there is still possible a drift of 25 kilocycle-per-volt change in anode potential and a 50-kilocycle drift per 0.1-volt change in filament voltage. Thus it is important to keep the supply potentials constant to the local oscillator tube and to mount the tube and circuit firmly and carefully in order to prevent temperature changes from causing serious oscillator drift.

Of course, the resonant circuit and tube of the oscillator must be properly shielded to prevent local oscillator radiation or feed-through into the antenna circuit. Unfortunately, this problem becomes more serious as vacuum-tube mixers are developed for UHF work, because the vacuum-tube mixer requires a much higher injection signal than those of the crystal mixers in use.

The internal and external parameters of a UHF oscillator stage are illustrated in Fig. 389c. In the UHF range the tube capacities and lead inductances determine the frequency limit of operation of the tube. Design of the external circuit must allow it to approach the limits of the tube parameters. The inductance of the plate and grid leads of the tube add substantially to the inductance of whatever type of resonant circuit is employed in the oscillator stage. In the illustration a transmission-line section is used as a resonant circuit, and this line has a wavelength factor substantially greater than its physical length, because of the loading influence of lead inductance. A quarter-wave shorted section of line can serve as a resonant circuit, or a half-wave open section can function as that circuit and permit the use of a longer physical length of line and fewer complications in tuning.

Feedback is established by the Colpitts-like division of voltages by the electrode capacities of the tube—C_{pk} and C_{gk}. To prevent external portions of the circuit from upsetting the tuned resonant circuit and internal parameters of the resonant circuit, the plate and grid lines, as well as the cathode heater circuits, must be adequately filtered and de-coupled. Supply potentials must arrive through actual inductors or resistor-capacitor filters, and external wiring must be planned carefully in order to avoid setting up resonant sections that cause parasitic oscillations or resonant suck-out. Such a resonant condition in the vicinity of the oscillator pulls out the energy from the resonant tank and either stops oscillation or reduces the injection voltage level. A cathode filter must be employed to keep the cathode-lead inductance and the capacities to heater and ground from setting up resonant conditions that may cause instability or parasitics. It is preferable to keep the heater at cathode potential and thus prevent cathode filament capacities from influencing the circuit stability.

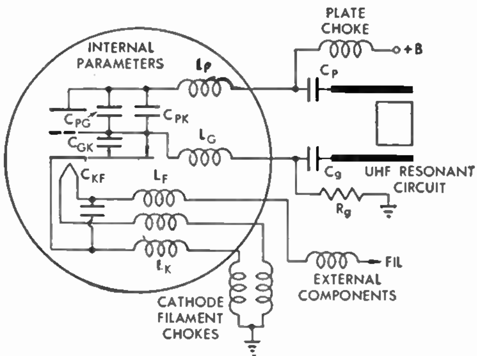


FIG. 389c Local Oscillator Stage

Local oscillator injection must be planned carefully to present only a light load to the oscillator, the light load preventing instability caused by circuit variations at the oscillator signal termination and also blocking at certain frequency ranges. Oscillator injection must be supplied to the crystal mixer through a loose-coupled link arrangement or some sort of a resistance-capacitance feed that keeps the oscillator injection reasonably constant with the output strength of the oscillator, itself, and the frequency.

UHF VACUUM TUBES

To permit efficient operation in the UHF band, the tube elements and spacing must be extremely small, and the inter-electrode capacitances and lead inductances must be held at a minimum. Likewise, the tube electrodes must be mounted as near as possible to the stem that feeds the leads to the external pins. The 6AF4 UHF oscillator tube represents a typical example, Fig. 389d. The nickel plate of the tube is approximately ¼ inch long, and to minimize grid-to-plate capacitance the plate area is held small enough barely to cover the active area of the cathode with a spacing of 0.004 inch. The spacing between the grid (formed of extremely fine, gold-plated wire) and the nickel cathode is just 0.001 inch. Thus inter-electrode capacitances are held at a minimum by the small physical side of the element, and lead inductance is held at a minimum by the low-mounting position, the use of direct, straight connections to the electrode, and the double connection to the grid and plate. All leads are silver-plated for the purpose of minimizing skin losses at the extremely high frequency. Very small electrode capacities and high mutual conductance are important in obtaining adequate gain and sensitivity for broadband high-frequency operation.

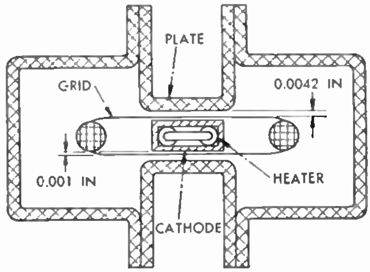


FIG. 389d UHF Tube Construction, RCA 6AF4

An unusual UHF r-f amplifier tube is the General Electric-developed 6AJ4 which can be used as a grounded-grid r-f amplifier, Fig. 389e. The tube elements are mounted horizontally instead of vertically and positioned very near to the tube base. With this particular mounting structure a very effective internal grid-shielding arrangement with five separate grid leads maintains excellent isolation between the input and output resonant circuits. If this isolation is not maintained or if it is lost in the external construction, there is a possibility that the input impedance to the stage would vary substantially over a received bandpass. The same effective isolation minimizes local oscillator radiation and regenerative effects.

A single cathode and a single plate lead are used, with a five-lead connection to the grid being of great significance. The 6AJ4 has a very high transconductance, as well as a high ratio of transconductance-to-plate-current, and attains a very low noise figure of operation in the UHF range.

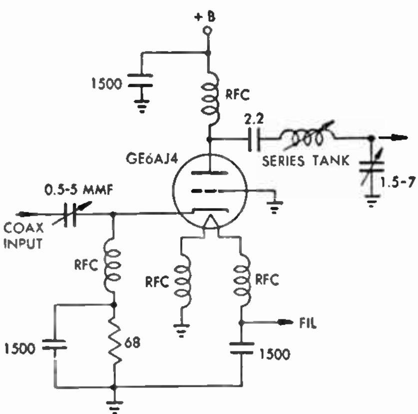


FIG. 389e Grounded-Grid R-F Stage

It is possible to use a vacuum-tube mixer in the UHF range, Fig. 389f; when a suitable noise factor is attained and local oscillator radiation held within safe limits, there are a number of attractive advantages to the use of a vacuum-tube mixer. A typical UHF grounded-grid mixer tube is the Sylvania 6AN4. The vacuum-tube mixer has the following features:

1. A conversion gain in contrast with a conversion loss. The conversion gain, of course, varies with the injection voltage but can be as high as 1.5 with a 2-volt injection signal.

2. A relatively constant input impedance and conversion gain with varying oscillator injection. This allows sensitivity to be held at a more constant value for a like change in oscillator injection voltage, as compared to the action of a crystal mixer.

3. A relatively constant impedance seen by the i-f input stage with varying oscillator injection.

185. *Commercial* UHF *Converters*

Descriptions of some of the commercial types of converters will help in understanding the circuit techniques used. In the General Electric converter, Fig. 390, the tuned-mixer and local oscillator lines are employed with a screw mechanism that moves a shorting contact along the line as the tuning control is varied. Proper tracking is assured by high- and low-frequency end alignment adjustments (*C5*, *C21*, and end adjustments of *L1* and *L2*).

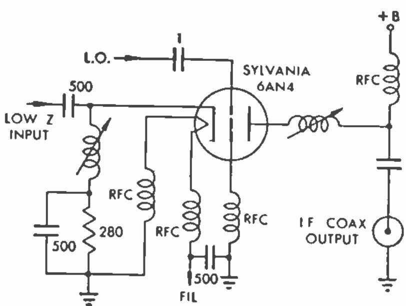


FIG. 389f Grounded-Grid UHF Mixer

The antenna input circuit, consisting of capacitors *C1* through *C7*, is a high-pass filter and blocks low-frequency interference and spurious beats. Local oscillations from *6AF4* ultra-audion are applied via 100-ohm termination *R12* and small 1-micromicrofarad capacitors.

As in many UHF converters, a low-noise triode i-f amplifier is used. As mentioned previously, signal-to-noise ratio is established at the first i-f stage when a crystal mixer is used and no UHF r-f amplifier is present. This i-f stage is peaked (*L10*) either on channel *5* or *6*, depending on which is unused in the district. The signal reaches output transformer *T1* through a low-pass filter that keeps undesired higher-frequency interference (VHF harmonics) from feeding into the VHF receiver and setting up spurious beats.

A versatile switching arrangement is required with UHF converters. Both the VHF and UHF antennas are attached to the UHF converter. When the switch is set on OFF or VHF position, the VHF antenna is shuttled around the converter to the antenna terminals of the VHF receiver. On VHF position, as opposed to OFF position, the filaments of the converter are on and keep the tube cathodes heated for immediate UHF use. On UHF position, the UHF converter is put into operation, using UHF antenna and with the converter output attached to the VHF receiver antenna input. Two UHF switch positions are for turning the pilot light on and off. When the same antenna is used for both VHF and UHF reception, it is advisable to use a crossover network to link the antenna with both converter inputs (refer to Fig. 369a).

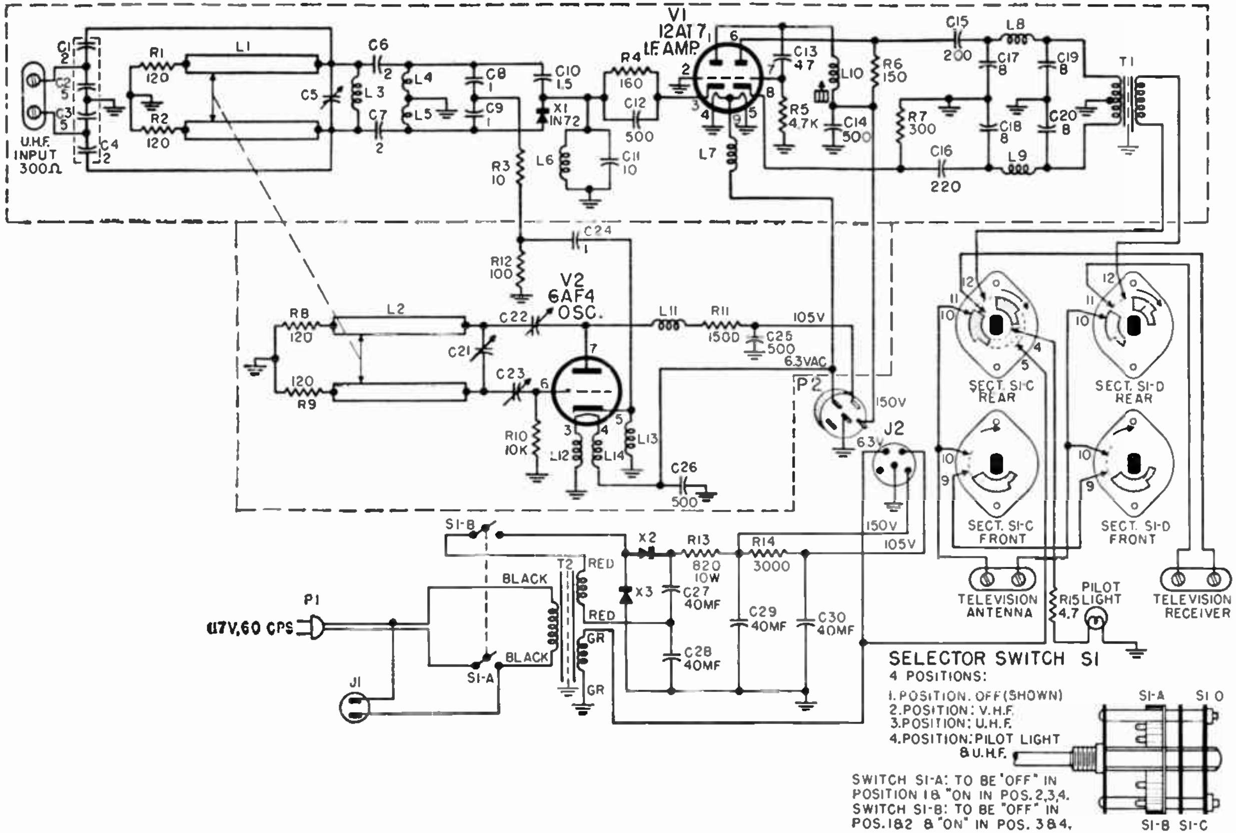


FIG. 390 Schematic, General Electric UHF Converter

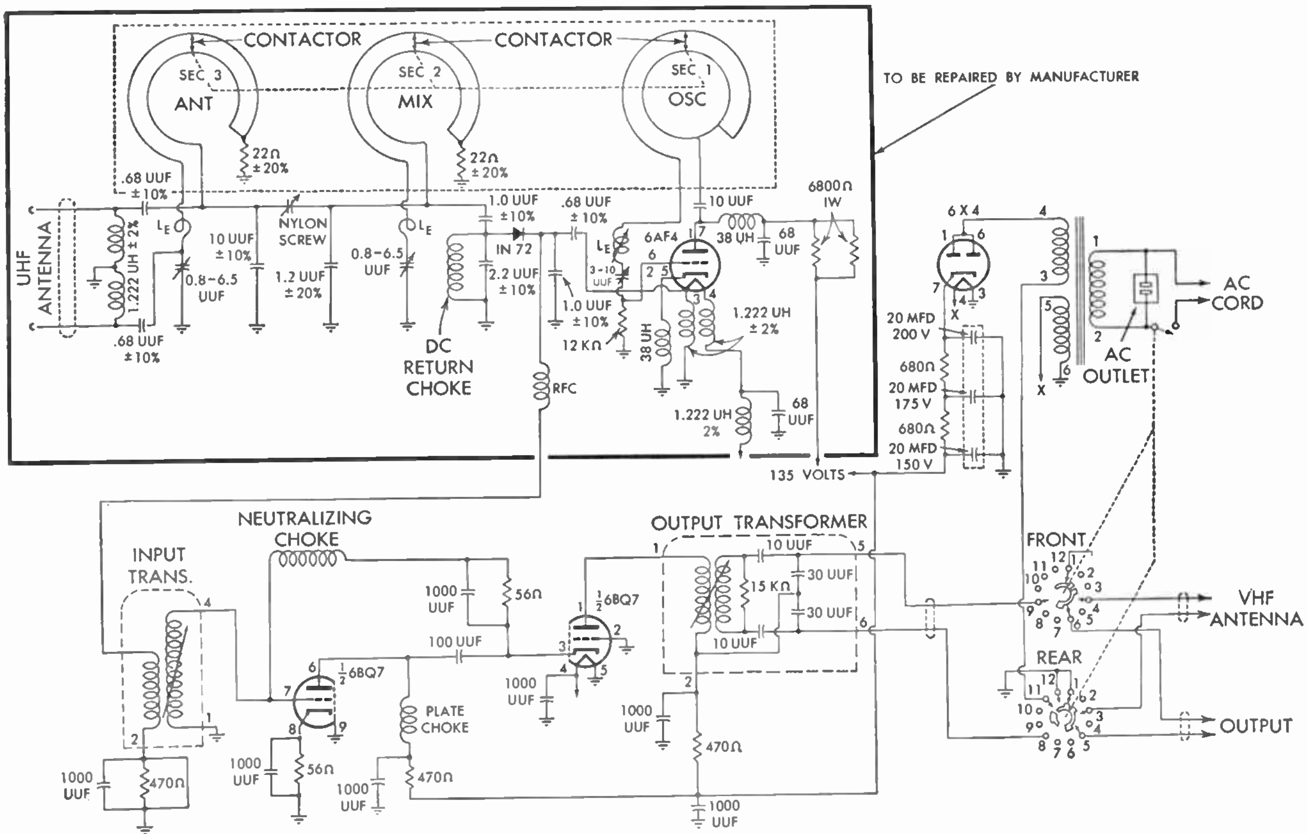


Fig. 391 Schematic, Mallory UHF Converter

MALLORY UHF CONVERTER

A Mallory-type UHF converter (tuned circuits which have widespread application in the converters of a number of manufacturers) uses the circular type of tuned lines, Figs. 391 and 392. Three UHF-tuned circuits are ganged to form a double-tuned input transformer and a local oscillator tank circuit. The double-tuned input transformer provides necessary bandwidth with good off-frequency signal-rejection. Small shunt capacitors and end-coils provide tracking adjustment, while a small capacitor between the high sides of the transformer regulates bandwidth.

Local oscillator injection to the crystal mixer arrives via the filament circuit of the UHF oscillator. This method of coupling via tube capacity (cathode-to-filament) provides uniform injection over the UHF range, with light oscillator loading and good stability.

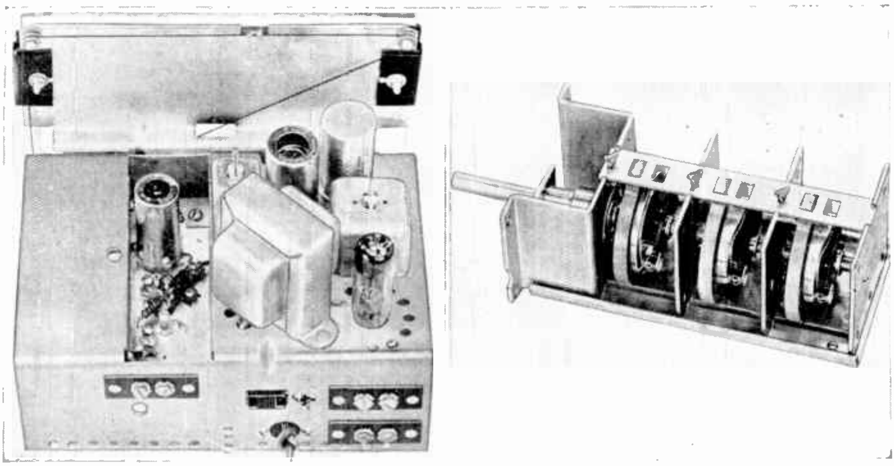


FIG. 392 Mallory UHF Converter and Tuned Lines

A cascode-type i-f amplifier (very common in UHF converters of all types) is used to establish the very best signal-to-noise ratio with excellent i-f stability.

The Regency converter, Fig. 393, also uses curved resonant lines, but they are longer, wider, and have greater spacing. The lines actually surround the oscillator and the i-f amplifier tubes and circuits. The pre-selector line is an effective half-wavelength (shorted at each end), presenting a low-impedance coupling to the antenna system at one end and a similar, low impedance at the opposite end for matching into the crystal mixer. The line is tuned with a shorting clip, Fig. 394, that moves away from the antenna end and has a total travel of four inches in covering the UHF band—at the same time, a second shorting clip also moves along the effective quarter-wave shorted section of the local oscillator line.

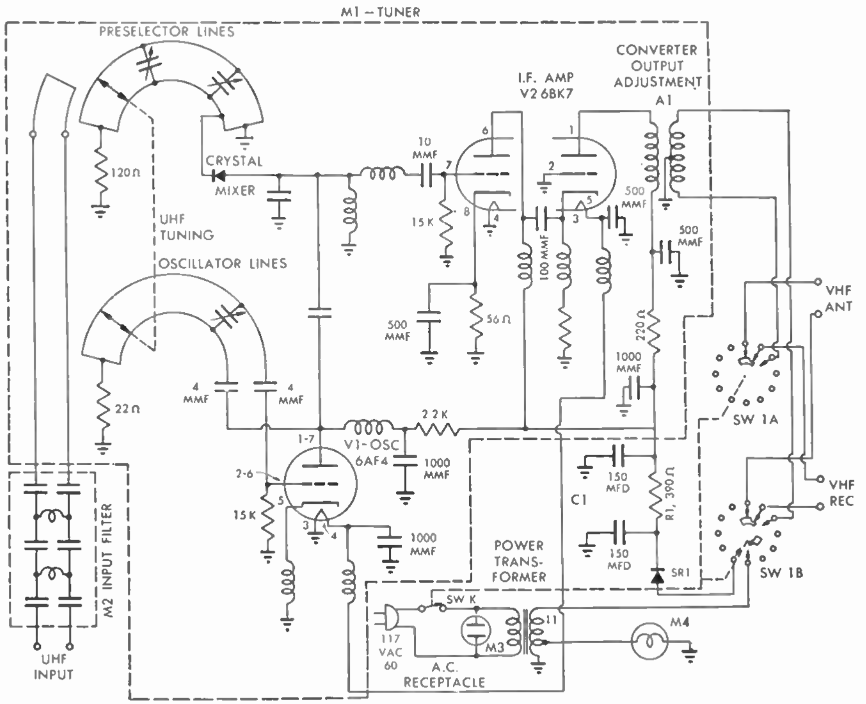


Fig. 393 Regency Converter

Three special trimmer capacitors (two shunting pre-selector lines and one across the oscillator lines) provide tracking adjustments with separate high- and low-end settings for the pre-selector lines.

The UHF antenna input is again by way of a high-pass filter in order to

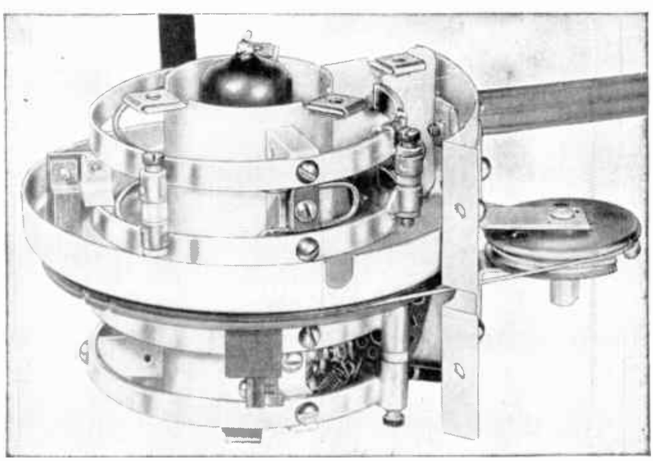


Fig. 394 Tuned Resonant Lines

minimize the possibility of VHF interference. A crystal mixer, a 6AF4 local oscillator, and a cascode i-f stage are again employed. The i-f amplifier frequency range, however, extends from channels *8* through *12* and is tunable by means of the alignment control at the rear of the converter.

COMBINATION BOOSTER-CONVERTERS

An excellent UHF- and fringe-combination plan has been developed by Sutco. This unit, Fig. 395, consists of a VHF booster and UHF converter on one chassis. Consequently, only a single external unit is needed in VHF fringe areas where boosters are necessary; the additional wiring, switching, and space required by a separate booster and UHF converter are eliminated.

In the installation of the Sutco converter, separate VHF and UHF antennas are attached to the terminals provided. The output of the converter connects to the antenna terminals of the VHF receiver. In VHF position, the incoming VHF signal is amplified by a push-pull neutralized *6J6* triode with tuned input and output circuits and is then passed on to the antenna terminals of the receiver. On UHF position, the incoming UHF signal is applied to a crystal mixer via a tuned input transformer, which uses Mallory continuously-tunable UHF-resonant circuits. A *6AF4* UHF local oscillator supplies the injection voltage. The i-f output of a crystal mixer is in the channel *5* to *6* range and is amplified by a VHF triode booster which is set to this range for UHF operation. It is apparent thta the VHF-booster section has a dual function, since it also acts as an i-f amplifier for UHF operation.

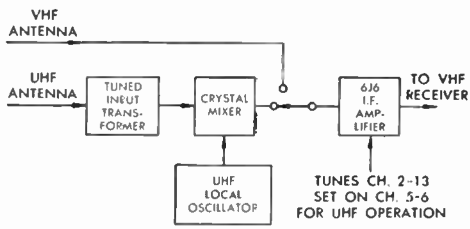


FIG. 395 Sutco VHF-Booster UHF-Converter Plan

A similar plan is used by Astatic, Fig. 396, which includes a UHF converter and a two-stage high-gain i-f amplifier. This amplifier is tuned to channels *5* and *6* for UHF reception and over the entire channel *2* to *13* range for high-gain VHF reception.

The Astatic UHF section uses a double-tuned pre-selector and oscillator-curved lines. Spaced, sliding contacts move around the insides of all lines simultaneously. This spacing between contact points prevents weak-sensitivity spots (suck-outs) at the high end of the band where unused portions of lines become a substantial percentage of the tuned wavelength. Low-end tracking and calibration capacitors are present on pre-selector lines. Proper tracking over the band is obtained with identical physical construction of lines, 27-ohm resistors (chassis isolation), and a 0.47-micromicrofarad capacitor. A silicon crystal mixer is used, and i-f output is coupled via the low-pass filter and unbalance-to-balance auto-transformer to the i-f amplifier input.

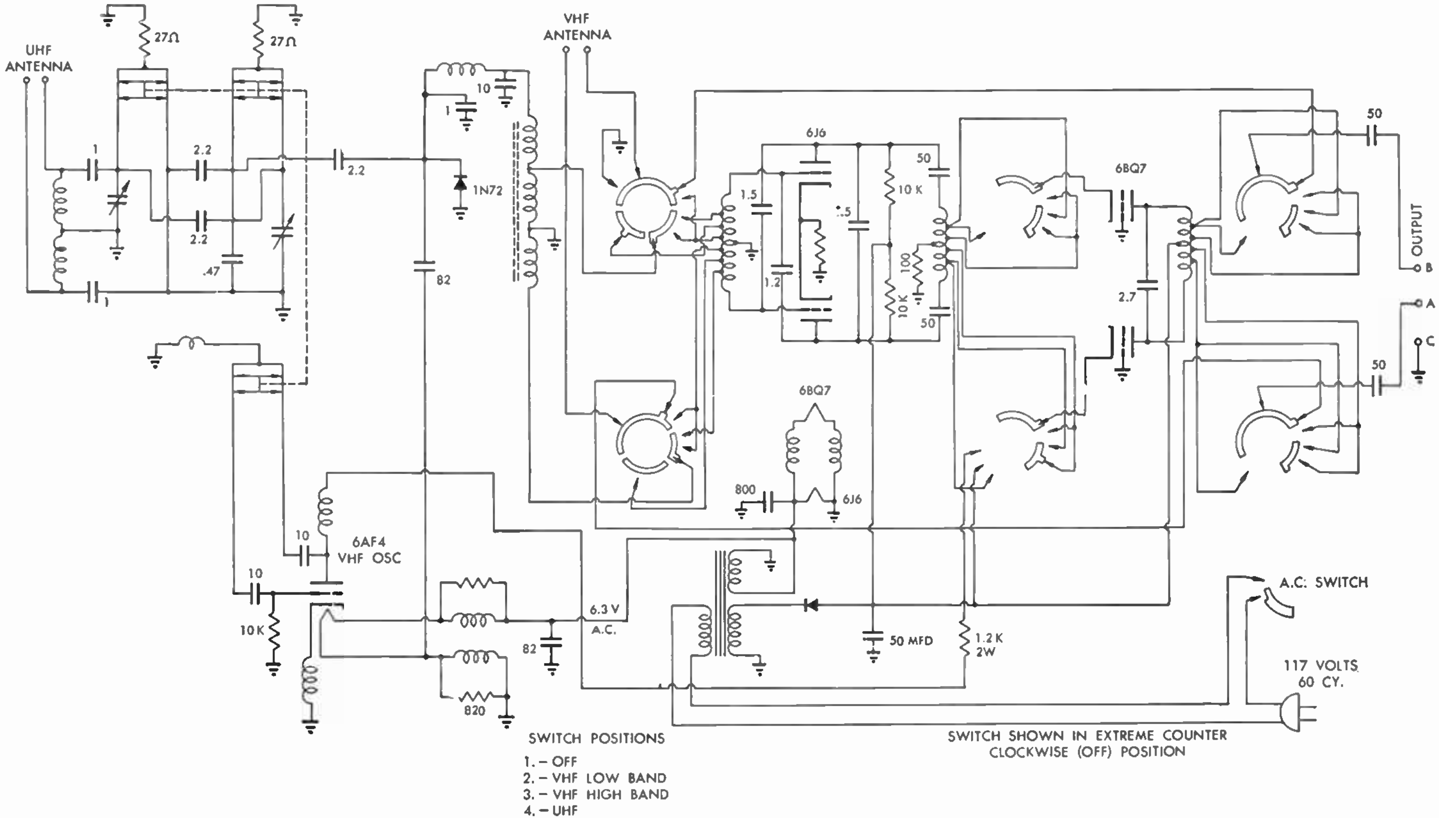


FIG. 396 Astatic Booster-Converter

The i-f amplifier and VHF-booster section uses a push-pull neutralized triode input stage and a grounded-grid output stage that feeds a balanced signal to the antenna input of a VHF receiver.

One of the newer converters employed in a number of combination UHF-VHF receivers is General Instrument's unit, Fig. 397. It employs three coaxial (a quarter-wavelength shorted) resonant circuits—a double-tuned input trans-

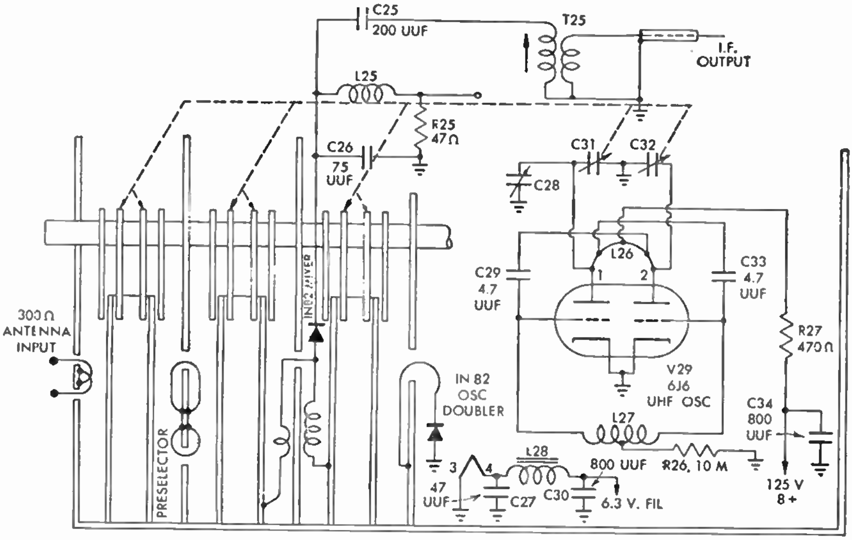


Fig. 397 General Instrument Converter

former and a local oscillator. The coaxial resonant circuits are end-tuned over the UHF range by ganged capacitors. This type of tuned circuit has a rising Q with the frequency, and a more uniform bandwidth can be attained over the UHF band.

Coupling between resonant circuits of the input transformer is accomplished via small coupling links—a large one for optimum coupling at the low end of the UHF band, a small one for the high end. The secondary is coupled to a crystal mixer through a small choke that is self-resonant below the UHF band; lower-frequency interference is rejected and crystal circuit impedance held uniform. The local oscillator injection signal is applied to the crystal via an identical coil.

The output of the crystal mixer is applied to the i-f output transformer. The unit can be designed to supply any desired VHF output frequency—the General Instrument unit uses channel 5 to 6 range with selection governed by a rear slider switch. A test point is provided for attachment of a scope for alignment or a VTVM for testing the crystal current and injection.

Crystal injection has much to do with the sensitivity and noise factor of UHF devices, and there is a general trend toward including points of test and even injection controls for adjustments to specified injection levels.

The unusual feature of the General Instrument converter is the use of a half-frequency local oscillator. A conventional and stable oscillator can be designed which uses a 6J6 operating at a rather low frequency. Output of the oscillator is applied to a doubler resonant circuit (third-coaxial tank circuit) that develops the proper crystal injection frequency. A second crystal is employed to emphasize the second-harmonic output of the local oscillator.

RECEIVER-MANUFACTURER TYPES

Most manufacturers of receivers have available not only UHF devices for attachment to the rear of present receivers ("back-porch unit"), but also external units and optional UHF tuners that can be supplied as internal units for the new receivers.

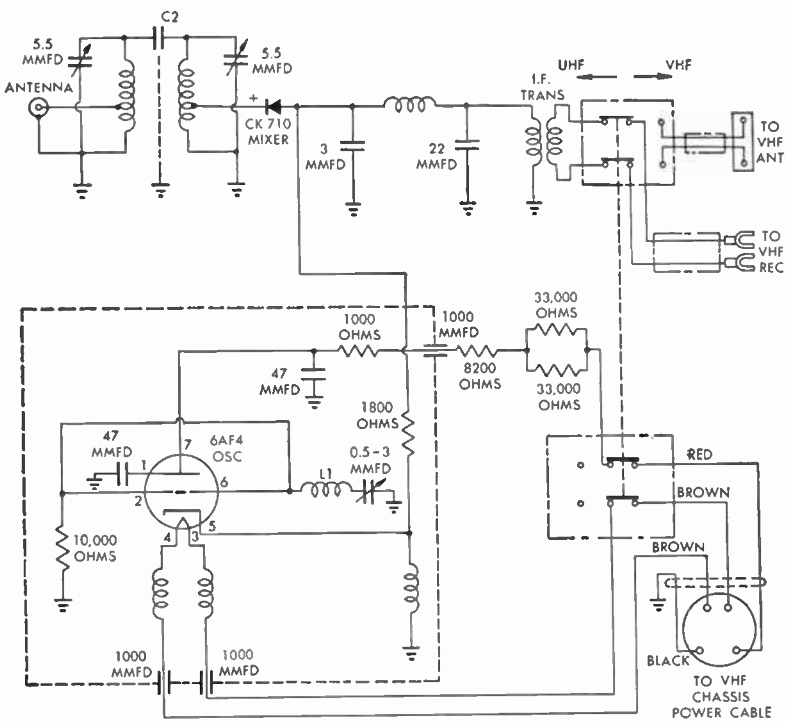


FIG. 398 RCA Single-Channel UHF Converter

As a typical example, RCA has available four separate UHF devices—single-channel, dual-channel, and all-channel external or internal UHF tuners.

The single UHF selector consists of a tuned input transformer and crystal mixer. After UHF mixing, Fig. 398, the difference signal (channels 5 to 6) is applied to the antenna terminals of the VHF receiver via the output i-f transformer. A *6AF4* local oscillator is employed and derives its power via an adapter that must be inserted into the tube socket of the audio output stage of an RCA receiver.

The selector chassis is attached to the rear of the receiver chassis. Its output is connected to the antenna terminals of the receiver. Separate terminals are provided for the attachment of VHF and UHF antennas. The audio output tube of the receiver is removed, and the adapter of the UHF selector is plugged into its socket. The tube is then inserted into the adapter.

Three pre-set trimmers and a jumper are adjusted, prior to installation, in accordance with the instruction chart and the channel to be received. After installation, trimmers are peaked for best picture and sound.

The dual-channel selector has the same general plan except that it includes its own power supply and an i-f amplifier stage. This unit can be used with any type of receiver and is positioned on top of, or to the side of, the present VHF receiver. A dual triode *6BQ7*, Fig. 399, provides two UHF oscillators that are pre-set for the two UHF channels to be received. Two tuned input transformers are also included, and each is pre-set for the two UHF channels to be received. An unusual feature of the UHF local oscillator system is that the second harmonic of the fundamental oscillator frequency is used as a mixer injection signal. This expedient simplifies design by lowering the oscillator fundamental frequency and reducing the frequency range over which the oscillator must be varied.

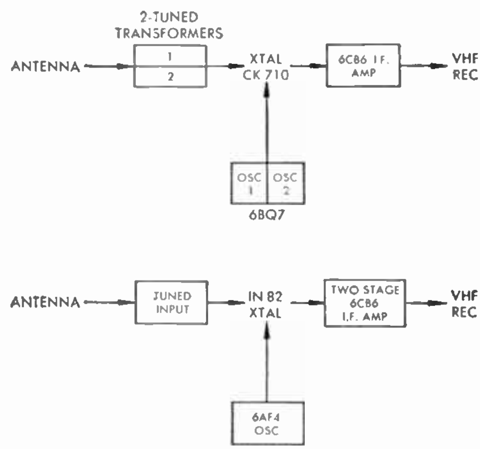


Fig. 399 Additional RCA UHF Devices

Sarkes Tarzian has a number of versatile UHF devices, Fig. 400, both for attachment to a present VHF receiver or for insertion in new models of receivers. Another unit permits the replacement of a present tuner with a VHF-UHF combination for either 45- or 25-megacycle i-f.

A single-channel but self-powered "back-porch" unit, Fig. 401, can add any single UHF channel to any receiver without any receiver change whatsoever, except those of attaching both the VHF and UHF antennas to the converter and the converter output to the antenna terminals of the receiver. Output of the converter can be adjusted to any unused low-band VHF channel. A Sarkes Tarzian external all-channel UHF converter can provide all-channel reception on any receiver. The converter output can be adjusted to any low-band channel.

Still another combination consists of an internal all-channel UHF and VHF

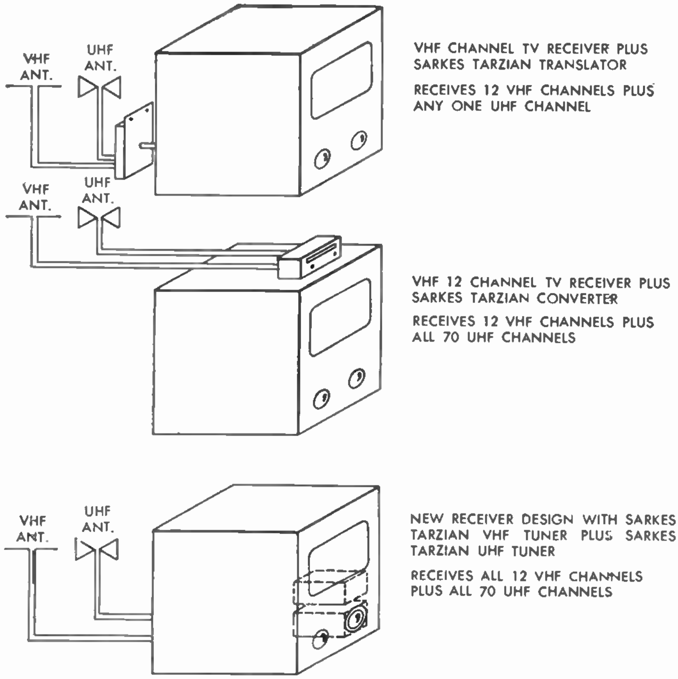


FIG. 400 Sarkes-Tarzian UHF Plans

tuner. Two small units provide all-channel coverage for new receivers or an internal replacement for present VHF receivers. In UHF position, the VHF-tuner section acts as a 45-megacycle two-stage i-f amplifier before the signal is applied to the standard receiver 45-megacycle i-f amplifier. Power for the converter section is obtained from the VHF tuner when it is switched to UHF. General Electric supplies a 3-channel UHF converter for attachment to the rear of present receivers. It has its own power supply and a long, insulated shaft that permits switching from the front of a General Electric receiver. The converter consists of six individual resonant circuits—three for pre-selector

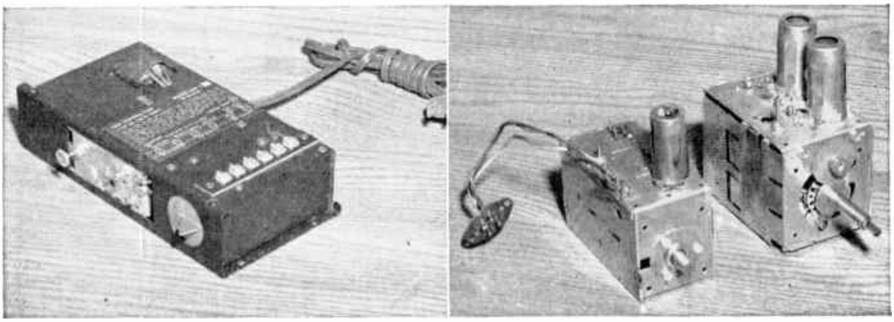


FIG. 401 Sarkes-Tarzian UHF Units

and three for the local oscillator. Each of the three pairs can be pre-set to any UHF channel, and therefore, a unit can be pre-set on any three UHF channels at installation. Resonant circuits consist of quarter-wave lines formed into a spiral on the inside wall of a plastic coil form. The frequency is varied with a shorting core that winds into a coil form.

An external-type converter developed by Granco does not use any sort of moving or wiping contact to tune over the UHF band, Fig. 402. It consists of three cavity-resonant circuits—a double-tuned input transformer cavity and

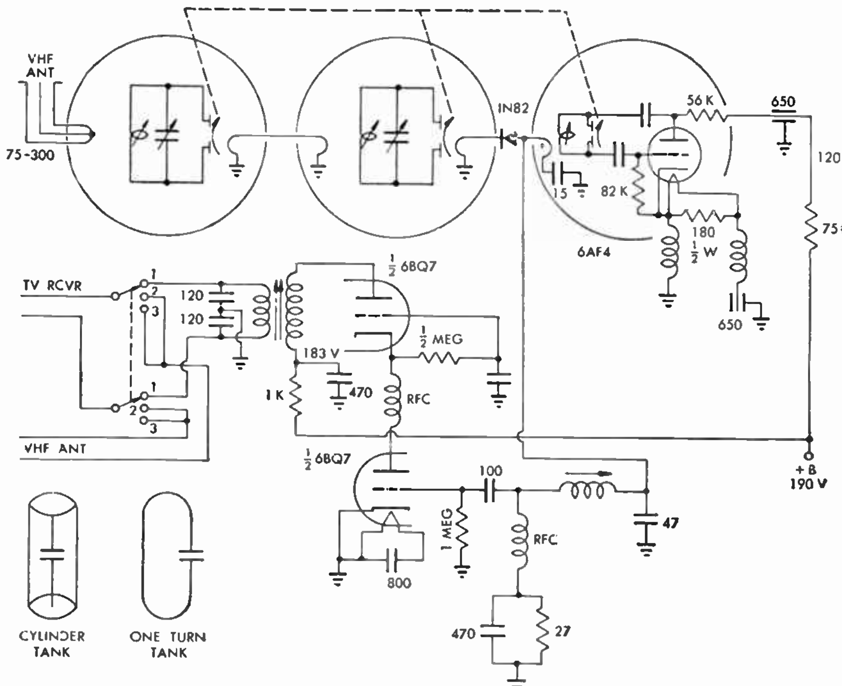


FIG. 402 Granco UHF Converter

a local oscillator cavity. A cavity-tuning element can be considered as one large single-turn inductor. For example, if we have one single turn about a fixed capacitor and we add an infinite number of joined turns to this single turn, we form a continuous but cylindrical single-turn inductor. The cavity is tuned by an adjustable capacitor that is centered in the cylinder. Tuning of each cavity is accomplished with a metallic plunger which, in relation to a loading capacitor, changes the value of the center capacitance. This plunger, moving inside of a low-loss dielectric sleeve, makes no physical contact with either the cavity or the leading capacitor. The capacitor, itself, is an internal coaxial sleeve that is broken at the center and separated by a dielectric tube with the metallic plunger riding inside of it and acting as the tuning element. Small loops couple in and out of the cavity field, and inter-cavity tracking is adjusted by proper positioning of the plunger, trimming of the loading capacitor, and mechanical means of changing the cavity electrical length.

Efficient shielding is obtained by the completely enclosed cavity, with additional shielding surrounding the three cavities as a group. Component parts of the local oscillator are built into its cavity, and local oscillation radiation is minimized and stability improved. Output of the silicon crystal mixer is applied to a low-noise cascode amplifier and then to the tuned output transformer which carries the signal to the antenna terminals of the VHF receiver.

STANDARD COIL VHF-UHF 82-CHANNEL TUNER

Standard Coil has developed an experimental VHF-UHF 82-channel tuner which uses switched coil-capacitor combinations. It is quite unique because it uses lumped coils and capacitors instead of the usual type of UHF-tuned circuit. A two-section dial is used, Fig. 403, one section counting in tens, the other in units. To set the tuner on channel *36*, for example, the tens' dial is set on 3 while the units' dial is set on 6.

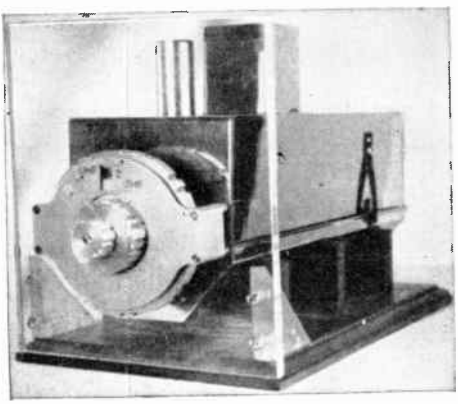


FIG. 403 Standard Coil 82-Channel Dial System

The general plan of the tuner, Figs. 404 and 405, is to employ eight specific UHF positions, each approximately 60 megacycles wide, the eight covering the entire 470- to 890-megacycle range as follows:

Tens	*Frequency (In Megacycles)*	*Channels*
1	470–510	*14–19*
2	510–570	*20–29*
3	570–630	*30–39*
4	630–690	*40–49*
5	690–750	*50–59*
6	750–810	*60–69*
7	810–870	*70–79*
8	870–890	*80–83*

Thus the UHF section, in accordance with the tens' dial-setting, opens the input over a 10-channel span of frequencies. If the tens' dial is set on 4, the signal from any channel between 630 and 690 megacycles reaches the UHF mixer. The choice of the proper one of these ten is made by the units' dial section, which sets the frequency range of the tunable cascode i-f amplifier. The cascode i-f stage can be set on any one of six megacycle ranges that include channels *7* through *13*. With the tens' dial on *4*, the channels from

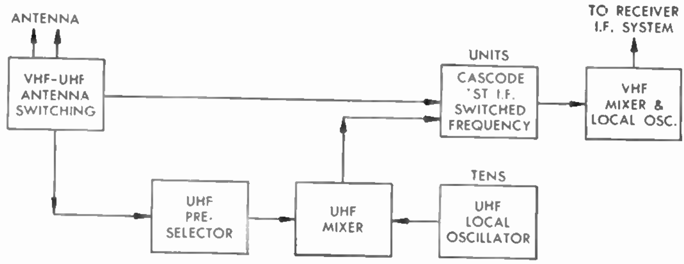


FIG. 404 Block Plan of VHF-UHF Tuner

40–49 could "fall in," each channel using a fixed UHF local oscillator frequency (one fixed frequency out of a choice of eight) as follows:

Units' Dial	*Channel Frequency (In Megacycles)*	*Local Oscillator Frequency (In Megacycles)*	*Cascode I-F Frequency (In Megacycles)*
0	626–632	470	156 162
1	632–638	"	162 168
2	638–644	"	168 174
3	644–650	"	174–180
4	650–656	"	180 186
5	656–662	"	186–192
6	662–668	"	192 198
7	668–674	"	198 204
8	674–680	"	204 210
9	680–686	"	210 216

This indicates that the UHF local oscillator need not be tunable but needs only to be switched to any one of eight pre-set fixed frequencies (plus some limited variation for alignment or fine tuning). Actual unit-channel selection is made by shifting the i-f frequency of the cascode first i-f amplifier rather than the frequency of the local oscillator.

Output of the cascode i-f is applied to a second mixer-local oscillator combination (VHF section) and thence to the conventional i-f amplifier of the receiver.

For VHF operation, the VHF antenna is attached to the input of the cascode i-f amplifier which selects channels *2* through *13* properly with the same, calibrated dial and without any additional switching required. The entire switching is automatic, and a user does not know that VHF-UHF switching occurs when the dual-dial is changed between the UHF and VHF ranges.

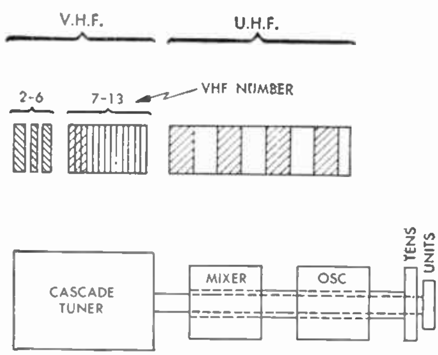


FIG. 405 Channel Selection Plan

A schematic of the UHF oscillator, Fig. 406, demonstrates the eight-position method of switching the local oscillator frequency range, using incremental

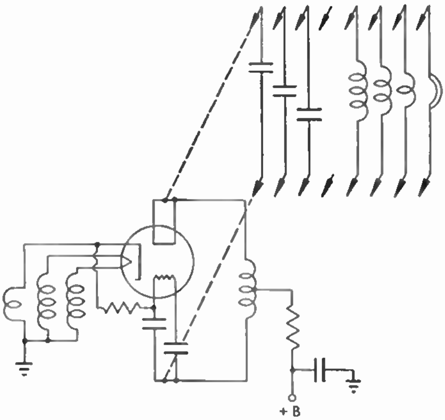


FIG. 406 UHF Local Oscillator Switching

coils and capacitors. The fixed fundamental frequency can be in the 400-megacycle range and is resonated at this frequency by two small coils in parallel (#16 wire on ¼-inch form). Higher frequencies are obtained with shunt inductors, lower with shunt capacitors. Frequency deviation has not been great, and a trimmer range of plus or minus 7 megacycles on the fundamental circuit position is sufficient to correct for the tube capacity and component dissimilarities.

186. *Test Equipment Requirements*

The alignment and testing of UHF strips and converters is not difficult. They are basically rather simple units, and very little additional test equipment of an elaborate type is necessary. Test units required will be a source of signal, a source of modulation perhaps for that signal, and finally an indicating device. Fig. 407. Initially, only one or two stations have come on the air in a given area. It is possible that a station signal and its modulation can be used as a signal source with which to complete the entire UHF-section alignment. Bandwidth is not too trying a problem on the UHF band, because of 6-megacycle bandwidth is small in comparison with a center frequency of hundreds of megacycles. Consequently, single-peaked tuned circuits have a 6-megacycle bandpass, and double-humped characteristics are not always necessary. Thus, simple peaking of resonant circuits for a maximum is often all that is necessary.

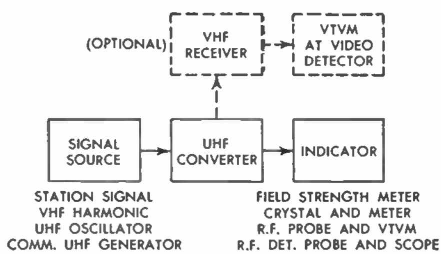


FIG. 407 UHF Test Equipment Plan

Another source of signal can be the harmonic output of the present signal generator (sweep and/or marker). VHF generators with a high fundamental range have been found to have a high-to-just-useable output on their UHF harmonics. Still a third possibility is to construct a small UHF oscillator to operate on certain assigned UHF channel or channels. The oscillator can be calibrated with a UHF converter, an accurate wave-meter, or harmonic-generating crystal-controlled oscillator. Manufacturers of test instruments will gradually market UHF-sweep signal generators with marker facilities. To summarize, the following signal sources are possibilities—UHF stations, harmonic output of present VHF generators, home-constructed UHF oscillator, and commercial UHF instrument.

It is likely that single-frequency alignment will suffice for most of the initial UHF converters. Modulation of the signal source is or is not required, as a function of the type of indicator used. For single-frequency alignment there is a number of indicator systems, depending on what is available. Inasmuch as the output of the UHF converter is in the VHF range, an r-f measuring device can be used.

An excellent indicator is the conventional field-strength meter used for VHF-signal level measurements, Fig. 407a. This type of instrument can be

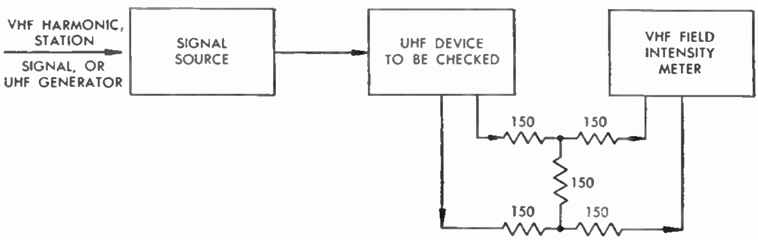


FIG. 407a VHF Field-Intensity Meter as UHF Peak Alignment Indicator

attached to the output of a UHF converter, set on the correct channel frequency, and serves as a good, sensitive, peak indicator. It also gives an approximate measure of the converter output signal voltage in microvolts.

Some other possibilities are a-c VTVM with an r-f probe or balanced crystal detector, Fig. 408, with sensitive current-meter. General Electric suggests the balanced crystal detector and a meter attached to the output of the converter or a VTVM attached to the output of the video detector in a VHF receiver. If modulation of signal source is possible, a crystal detector probe and scope can be used.

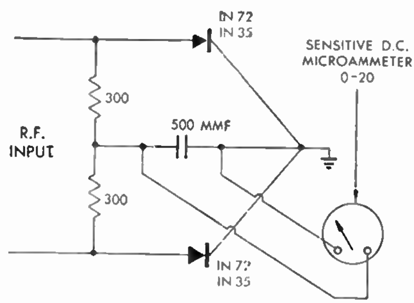


FIG. 408 Balanced Crystal Detector

A useful signal source for UHF-checks can be constructed around a single Mallory UHF-tuned circuit. A simple ultra-audion oscillator can be constructed of small high-frequency components, Fig. 409. The secrets of any UHF construction are short leads and good quality parts. Keep the plate and grid leads at a minimum by elevating the tube mount to a position in which the grid and plate terminals are very near the tuned circuit lines. A low-impedance signal is taken from the cathode via the 100-ohm termination.

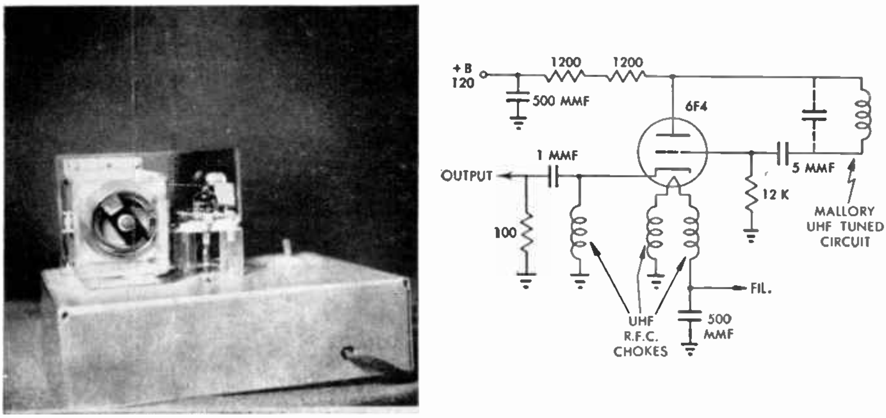


FIG. 409 UHF Test Oscillator

TRANSMISSION LINE CHECKS

In comparing transmission lines for UHF performance, it is advisable to check on at least two frequencies—one at each end of the UHF band. The great differentials in antenna impedances and input impedances of the UHF devices over the wide span of UHF frequencies indicate that lines should be compared with the same antenna and with the same UHF unit. A tuning stub (a 2-inch piece of aluminum foil wrapped around the line) is also advisable. Stretch out 100-foot sections of the lines to be compared, Fig. 410, and apply

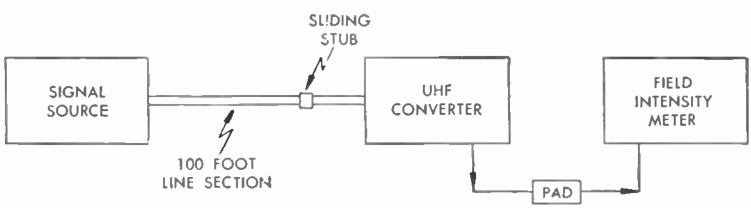


FIG. 410 Transmission-Line Check

the signal source or pick-up station signal via the antenna system. Attach one of the lines, and record the output reading on the field-intensity meter after the tuning stub has been set for maximum reading by sliding it along the transmission line. Now, substitute a 100-foot section of the second line under exactly the same conditions, and record the reading after adjusting the sliding stub for a new maximum.

A pad is inserted between the converter and field intensity meter in order to minimize the exchange of local oscillator signals between units and consequent beats that record on the meter. The pad does not eliminate these points but reduces them substantially.

All lines should be compared again after they have weathered a few weeks and should, this time, be checked in wet weather after the line has been rather well-saturated with rain. Weather has a great influence on the performance of lines—performance differentials being quite decided among lines as a function of dielectric quality and the amount of air-dielectric spacing. Two lines of the same type and with equal performance in dry weather can separate widely in their relative performance in wet weather.

ANTENNA-PERFORMANCE CHECKS

Nothing can be so confusing as antenna-comparison checks when variables are not properly satisfied. This condition applies specifically to all antenna comparisons made in the UHF range. In checking antennas, Fig. 411, again

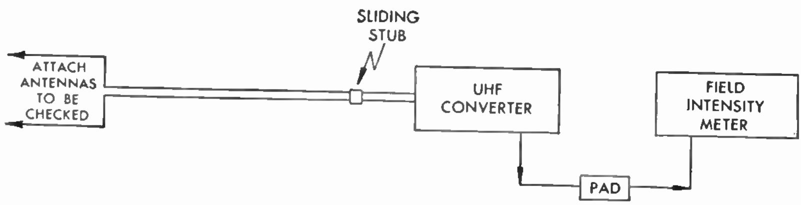


FIG. 411 Antenna-Performance Checks

use the same equipment combination for each antenna to be compared. Use a single transmission line that is fixed in position and clear of metallic surfaces. In fact, it is advisable when making comparison checks to have one conductor of the two always attached to the right side of the driven element as each antenna is positioned—the second conductor is always attached to the left side of the driven element.

Use the aluminum-foil stub to maximize the signal for each antenna and for each frequency at which antennas are to be compared. A step-by-step procedure for checking relative antenna sensitivities is as follows:

1. Interconnect the field-intensity meter and the UHF converter that is to be used for taking measurements.
2. Connect the transmission line to the converter, and position the line and equipment in a fixed position from which they will not be disturbed during the measurements.
3. Attach the aluminum-foil stub to the line near its termination at the converter. It is advisable to connect a 1½-foot section of flat 300-ohm line between the end of the UHF line and the antenna terminals of the converter and to fasten foil around it for better peaking.
4. Attach the first antenna to be checked at the antenna end of the transmission line. Record the reading after the signal has been carefully tuned in and stub positioned for maximum reading.
5. Now, quickly position the second antenna at exactly the same spot and record the reading after the stub is set for a maximum.
6. Do the same for other UHF-channel frequencies allocated in the area.

Sensitivity of an antenna to reflections can be observed by rotating it and observing the number and relative sensitivities of the minor antenna lobes. These readings can be compared with the same kind of measurements made on other antennas under check.

In UHF measurements, it is important that the field-strength meter be in good operating condition. It should be possible to peak the VHF field-intensity meter to the center of those VHF channels which are to be used for UHF reception in an area. Most meters use some form of a standard VHF tuner, and oscillator slugs should be adjusted on the various VHF channels with the fine-tuning control set at mid-range. Also, in making any relative comparisons, be certain to use the same sensitivity range-setting on the VHF field-intensity meter.

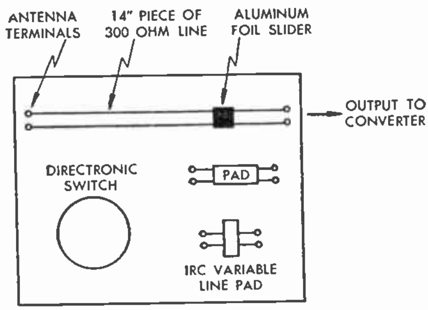


Fig. 412 UHF Test Board

We realize that in the early days of UHF operation, most of you will want to gain test experience and also make equipment-performance comparisons. However, be systematic about tests, because confused results will be your only reward if variables are not properly considered. In any antenna checks, the antennas to be compared must be mounted in exactly the same position, identical lengths of lines must be used, and the mast and line must be fastened rigidly lest positions shift in the wind as antenna changes are made. The input system and the measuring device must be identical for each unit being compared and, in most cases, should include a small tunable stub to tune-out the influence of the line and the input system in regard to the standing-wave ratios. In converter-checking too, all associated equipment should be held constant and identical and the stub used in order to maximize the signal to the input of each converter. A test layout should be planned for the fast and convenient substitution of various converters under check.

A plan of a simple field-test board, Fig. 412, should include a section (14 inches) of flat 300-ohm line pulled tight, with terminals at each end, and a small sliding stub. Also include a small attenuator with which to approximate fringe conditions for certain tests—also an IRC variable attenuator should be included and a Directronic switch for checking this type of antenna. Limited checks can be made on station signals, and if required, they can be augmented by tests performed with a small UHF-signal source.

187. *Alignment Procedures*

In gaining the best understanding of the plan of alignment, following the step-by-step procedures for a typical converter is helpful.

SUTCO ALIGNMENT PROCEDURE

Equipment required:

1. R-F signal sources covering both the VHF and UHF ranges
2. D-C microammeter
3. Detector, consisting of two 150-ohm carbon resistors in series with two *1N34* Xtals, micro-ammeter as a load, by-passed with 270-micromicrofarad returned to chassis
4. VHF field-strength meter may be substituted for items 2 and 3

VHF ALIGNMENT

Connect the 300-ohm output from the VHF signal source to terminals 1 and 2 on the booster converter and adjust to 70 megacycles. Connect the detector or the VHF field-strength meter to terminals 4 and 5 on the booster converter. *Do not* connect converter to the source of power. Rotate selector switch from channels *2* to *6*. Adjust the booster knob until some indication of output is obtained. Twist the neutralizing capacitor beneath the 6J6 tube socket until this output is minimum.

Connect the booster converter to a source of power, 117-volt 60 cycles per second.

Set the VHF signal source to 55 megacycles. Adjust to channel *2* by rotating the booster shaft to an extreme clockwise position. The yoke with attached cores will now be nearest to the chassis. Next, adjust the booster shaft counter-clockwise until the yoke has moved up by ¾ inch. Adjust the plate core (nearest to the filter capacitor) for maximum output. Move the signal source to 70 megacycles. Rotate the booster shaft until maximum output is obtained. Adjust the grid core (nearest to the front of chassis) for maximum output. The booster coils are dressed in alignment at the factory and then cemented in place and, therefore, would not normally require adjustment for the high channels (*7* to *13*).

Move the VHF signal source to 78 megacycles. Rotate the booster shaft for maximum output. Leave the booster shaft in this position, as it will be used in UHF alignment.

UHF ALIGNMENT

Rotate the selector switch to UHF and with the VTVM measure the bias voltage at the grid of the 6AF4 tube, Fig. 413. Should this voltage fall below negative 2 volts at any setting of the UHF shaft, replace the 6AF4 tube. Connect *47K* ¼-watt carbon resistor to the end of the VTVM test probe. Connect the 300-ohm output of the UHF signal to the UHF antenna terminals. Rotate the UHF shaft to an extreme counter-clockwise position. Move the UHF signal source, from 460 megacycles to 470 megacycles, until an indication of output is obtained. Adjust the oscillator trimmer capacitor located beneath the 6AF4 tube socket, until maximum output is obtained with the UHF source at 465 megacycles.

Move the UHF signal source to 475 megacycles. Rotate the UHF shaft until maximum output is obtained. Adjust the mixer and r-f trimmer (located on top of the UHF chassis) for maximum output.

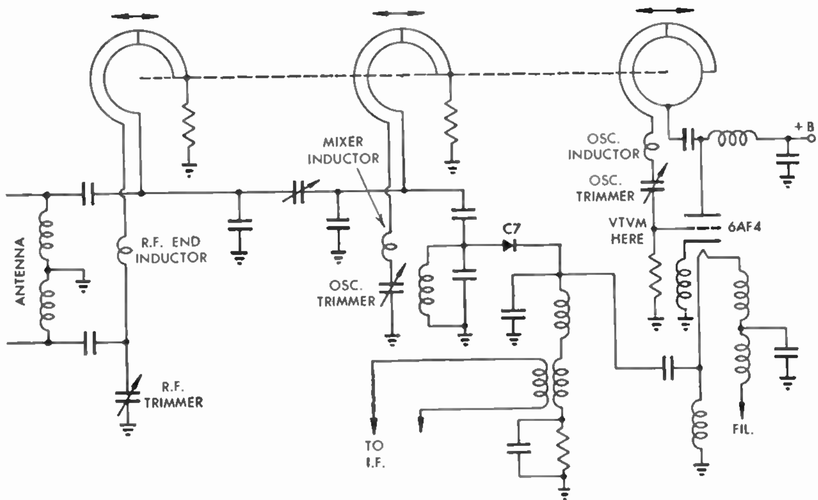


FIG. 413 Alignment of Sutco UHF Section

Move the UHF signal source to 890 megacycles. Rotate the UHF shaft to maximum clockwise position. Move the UHF signal source around 890 megacycles until an indication of output is obtained. Compress or expand the oscillator-end inductance connecting the oscillator trimmer to the tuner assembly, until maximum output is obtained at 895 megacycles.

Move the UHF signal source to 885 megacycles. Rotate the UHF shaft until maximum output is obtained.

Compress or expand the r-f and mixer-end inductance (connecting the trimmers to the tuner assembly) for maximum output.

Move the signal source to 475 megacycles, and readjust the r-f and mixer-trimmers for maximum output.

It is apparent that alignment procedure for some converters is quite similar to the method used in the alignment of AM radios. It involves alignment of a few adjustments first at one end of the band and then a second set at the opposite end of band. Finally, both ends are checked and peaked again to obtain satisfactory tracking, as there can be some interaction between adjustments. For sweep-alignment procedure of a UHF insert, refer to the discussion of alignment procedure in chapter 13.

UHF SWEEP AND MARKER GENERATION

A number of methods can be used to generate a UHF-sweep signal—rotating motor, motor drive (vibrator), or reactance tube. It is no more difficult, in fact a bit easier, to obtain the necessary frequency deviation in the UHF band, because of the much higher center frequencies. The task is to obtain amplitude linearity and to prevent the deviation system from adversely affecting the operation of the oscillator at the very high UHF frequency.

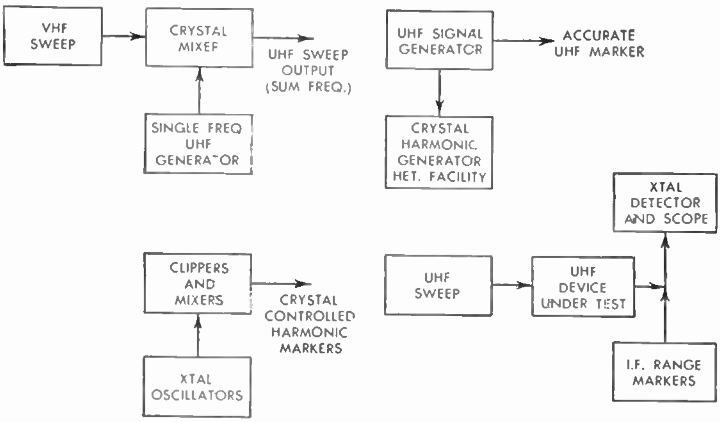


FIG. 414 Sweep Alignment and Marker Generator Systems

In the RCA UHF-sweep generator a vibrating motor placed near a modified cylinder tank circuit permits a sweep-width deviation continuously variable from zero to 45 megacycles, and amplitude variations at the output have been held down to 0.1 decibel-per-megacycle for a 45-megacycle maximum deviation. Still another method of sweep generation that may be common in UHF equipment is the use of a VHF-sweep frequency and a mixer crystal arrangement, Fig. 414. In this system, the deviation occurs in the VHF range before the sweep signal beats with a single-frequency tunable UHF oscillator. The mixing takes place in a crystal mixer; an output tank circuit, instead of tuning to the usual difference frequency, is tuned to the sum frequency and produces the same sweep deviation in the UHF band. In fact, with this method, we are using the principle of a UHF converter in reverse—we are mixing a low and a high frequency together to produce a still higher frequency output, while in the UHF converter we mix two high frequencies to produce a low-frequency output. It is possible with a UHF converter to modify the high i-f amplifier (to reverse the direction of the high i-f amplifier by having its plate supply the signal to the crystal mixer) and to obtain a UHF sweep output by applying a VHF-sweep signal to what now becomes the grid-feed to the high i-f amplifier.

There are three common methods for generating UHF markers, Fig. 414.

One system is to form, with crystal-controlled frequencies and mixing stages and clippers, a series of harmonic markers over the UHF band. A switching arrangement is required to locate certain key markers spaced widely in frequency; then by switching in other markers the desired channel markers can be located. Another method is that of a fundamental UHF oscillator which can be set on any specific UHF frequency, with its calibration being checked by crystal harmonic generators and a heterodyning frequency meter. A third method is the "back-porch" approach—putting markers on the response curve. In this system the UHF sweep is applied to the input of the converter under check, and key i-f markers are introduced on the curve at the output (high i-f frequency output) of the converter.

The latter method has one disadvantage in that the markers which appear on the response curve are true only if the local oscillator of the converter has been set properly on the channel to be received. This can be accomplished by using a station signal or a UHF generator single-frequency source.

CRYSTAL CALIBRATOR FOR UHF/VHF APPLICATION

One of the problems in the use of alignment test equipment on the UHF band is frequency accuracy. For example, if a specific UHF generator has an accuracy of 2 per cent at 800 megacycles, its output frequency could be located at any point over a span of 32 megacycles or 5 channels. The Triplett Instrument Company has developed an experimental crystal calibrator to meet this need for crystal accuracy on the UHF band, Fig. 414a.

The unit consists of two crystal-controlled oscillators set on frequencies of 60 megacycles and 6 megacycles. These oscillators can be operated singly or together. The 60-megacycle unit is an overtone circuit with the output removed from the cathode circuit and applied to the grid of the mixer output stage—the pulse nature of the cathode output means that the signal is rich in harmonics, producing strong harmonics (spaced ten channels apart) over the entire UHF band. The output is taken from the cathode of the mixer and is coupled via the marker amplitude control to the output terminal.

Likewise a UHF signal can be applied to the terminal and, therefore, to the cathode of the mixer. In the mixer tube the applied external signal can beat with one of the harmonics of the 60-megacycle crystal to develop a beat-note across resistor *R8* in the plate circuit. This signal is coupled through capacitor *C5* to a two-stage audio amplifier and is reproduced as an audible beat at the loudspeaker. Thus the unit can be used as a heterodyne frequency meter to calibrate or set an external marker generator (UHF or VHF) on the correct frequency.

The right-hand side of tube V2 is a 6-megacycle Pierce crystal oscillator that produces 6-megacycle harmonic signals at the cathode output circuit. These signals can be used to supply markers in the i-f frequency spectrum or in the television VHF band. Again, it is possible to use them as markers, or with the

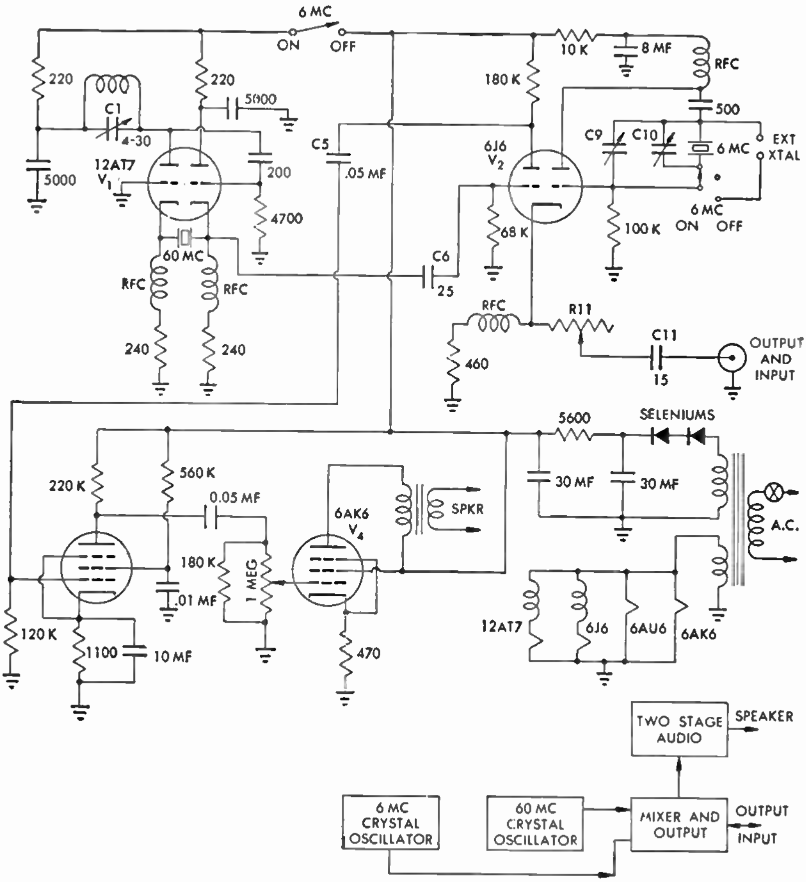


FIG. 414a Triplett Calibrator Plan

application of an external signal generator to the terminals you are able to set it on a correct frequency, using the heterodyne principle.

The two oscillators can also be operated simultaneously to permit a 6-megacycle side-band modulation of the 60-megacycle signal, producing harmonic side-band markers every 6 megacycles over the entire UHF band—*a marker for each UHF and VHF channel.* In this method of operation a beat-note is present between the 60-megacycle crystal and the tenth harmonic of the 6-megacycle crystal. These two oscillators can be brought to an inaudible beat by tuning the Pierce crystal circuit, capacitor *C9,* to a zero beat. Again, the output can be used for a marker injection of a sweep curve, or the unit can be used to heterodyne an external generator on a precise frequency.

The calibrator has the following applications:

1. It supplies 6-megacycle crystal-controlled markers over the entire VHF and UHF bands with key *60-megacycle markers* every ten channels on the UHF band and on VHF channels *2, 3, 7,* and *8.*

2. For VHF application, it supplies crystal-controlled markers on each channel as follows:

Channel	*(Megacycles)*	*Channel*	*(Megacycles)*
2	54 and 60	8	180 and 186
3	60 and 66	9	186 and 192
4	66 and 72	10	192 and 198
5	78	11	198 and 204
6	84	12	204 and 210
7	174 and 180	13	210 and 216

3. For i-f calibration it supplies 24-, 42-, and 48-megacycle markers.

4. For UHF application, it supplies a key calibration heterodyne marker every ten channels at the following points:

Channel	*(Megacycles)*	*Channel*	*(Megacycles)*
15	480	55	720
25	540	65	780
35	600	75	840
45	660		

5. An individual heterodyne marker is present on each UHF channel as follows (one megacycle above band center-frequency):

Channel	*(Megacycles)*	*Channel*	*(Megacycles)*	*Channel*	*(Megacycles)*
14	474	38	618	62	762
15	480	39	624	63	768
16	486	40	630	64	774
17	492	41	636	65	780
18	498	42	642	66	786
19	504	43	648	67	792
20	510	44	654	68	798
21	516	45	660	69	804
22	522	46	666	70	810
23	528	47	672	71	816
24	534	48	678	72	822
25	540	49	684	73	828
26	546	50	690	74	834
27	552	51	696	75	840
28	558	52	702	76	846
29	564	53	708	77	852
30	570	54	714	78	858
31	576	55	720	79	864
32	582	56	726	80	870
33	588	57	732	81	876
34	594	58	738	82	882
35	600	59	744	83	888
36	606	60	750		
37	612	61	756		

6. The above harmonics are not only useful as heterodyne signals but are often strong enough for direct application as markers on a UHF-sweep curve.

7. Harmonics of a 60-megacycle crystal can be applied directly to the input of a UHF device, thus supplying calibration signals, serving as excellent tracking signals, at spot frequencies every ten channels.

8. Weak half-frequency harmonics are also available, giving additional check points—for example, harmonics of 30 megacycles fall halfway between regular, strong 60-megacycle harmonics over the UHF range and permit marks every five channels between regular, strong 60-megacycle points. Likewise, on the VHF band fainter harmonics fall halfway between regular, strong 6-megacycle harmonics.

UHF-SWEEP GENERATOR BY MODIFICATION OF A UHF CONVERTER

It is possible with a standard VHF-sweep generator and a slightly modified UHF converter to form a UHF-sweep signal of reasonable linearity and ease of operation. In conventional application, the UHF converter-mixer crystal matches the UHF-signal frequency and the UHF local oscillator, producing a *difference frequency* in the i-f range. Likewise, it is certainly possible to crystal-mix a low and a high frequency to produce a still higher, or *sum, frequency*. Thus, if we apply a VHF-sweep signal to what is the normal VHF output of a converter, Fig. 414b, and mix that signal with the local oscillator signal, a UHF-sweep signal is available at what is the normal UHF-antenna terminals.

Although this non-modified arrangement performs to a degree, the high i-f amplifier of the converter is passing the signal from plate to grid, and there is a substantial loss. A simple modification of the i-f stage—reversal of its direction—improves the UHF-sweep output and linearity. The input circuit of the i-f amplifier should be connected to the VHF output terminals ("now" being used as VHF-sweep input terminals), and the plate output circuit of the converter i-f amplifier should supply an amplified sweep signal to the crystal mixer.

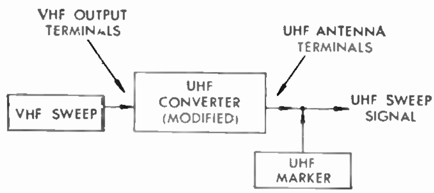


FIG. 414b Indirect UHF Sweep Generator

A Mallory UHF converter can be modified by breaking only two circuits. The lead between the r-f-c choke (at the crystal mixer output, Fig. 391) and the primary winding of the input transformer is opened, and the 300-ohm line between the switch and the VHF output terminals disconnected at the switch. The free end of the mixer choke is, at this time, connected to terminal number 5 of the output transformer, and the disconnected 300-ohm line, connected across the primary of the input transformer.

To place the UHF sweep in operation the VHF sweep must be first set on channel 5 or 6. A UHF-sweep signal is then available at the UHF antenna terminals of converter, the output center frequency being indicated approximately by calibration of the UHF converter dial. A UHF-sweep signal is available on any UHF channel by means of setting the dial on that channel (assuming the UHF converter dial has been calibrated accurately and has been tracking properly). There are a few precautions to take in setting up the alignment equipment:

1. Use short 300-ohm line to connect the sweep output to the UHF input or the UHF device to be aligned. Keep this line clear and away from other leads. Establish good grounding between units.
2. Do not over-drive the converter, a condition indicated by a flattening of the response curve as the output of the VHF sweep is increased.

3. Do not permit the marker generator to distort the response. Also be certain not to align with a false marker. There are at least two, and perhaps three, local oscillators functioning in the usual UHF alignment set-up. Opportunities for false markers and beats are obvious. A true marker gives best results as follow:

   a. When the frequency (center frequency) of a UHF sweep generator is changed (by varying the tuning control of a modified UHF converter), a *true marker will stay at the same position on the response curve* as the curve is moved along the screen.

   b. When the tuning control of a UHF device under alignment is changed, the true *marker moves along the curve* (the marker stays fixed at the same *horizontal position* on the scope screen). A true marker will respond to both checks, a false one to only one or to neither check.

UHF-SIGNAL GENERATOR

The same arrangement can be used to generate a single-frequency UHF signal as well. In this application a single VHF fixed frequency is applied at the same point as that to which the previous VHF sweep signal was applied. Thus, a single frequency for any UHF channel can be made available at the UHF-antenna terminals. In fact, this can be a video-modulated UHF signal. It is possible to pick up a strong channel 5 or 6 signal directly from the antenna and then apply it to the VHF input of a modified converter. There is available at the UHF-antenna terminals a video-modulated signal on any UHF channel for antenna or converter tests.

188. *Peaking a Weak UHF Signal*

At present it is our responsibility to use to best advantage the UHF devices now available and which operate at the signal levels that exist. UHF is hard work in weak-signal locations. An installation is a time-consuming job if the most is to be derived from the available signal.

1. The proper positioning of an antenna is more important than the type of antenna used. We cannot overemphasize the importance of probing the location horizontally and vertically, since an improvement of ten times the signal level can be obtained—a substantially greater improvement than can be obtained with a change from a simple dipole to a high-gain antenna.

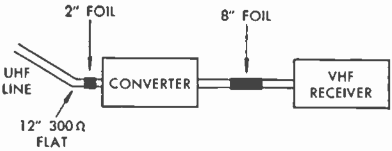


FIG. 415 Peaking of UHF Signal

2. In some locations, in valleys, and in back ridges an unusable signal is obtained when the antenna is directed toward the station. However, on occasion, a usable signal can be obtained by reflection from a nearby ridge. The reflected signal varies substantially as the state of the reflecting surface is changed by the weather conditions. Use this indirect means of signal pick-up only where necessary.

3. We have elaborated upon the proper choice of antenna type and have compared characteristics. Find the antenna type best-suited to your needs and the pertinent channels. Proper UHF line and careful installation are necessary if reliable coverage and uniform reception during changes in weather conditions are to be retained.

4. At present, converter sensitivity depends on mixer crystal sensitivity and proper local oscillator injection. Thus, substitution of the crystal and local oscillator often helps to bring up a weak-signal level.

5. Use of a stub on the UHF line was mentioned previously as a means of deriving peak signal. Some additional improvement can be attained by peaking the line between the converter and receiver, Fig. 415. We have also found it helpful to change over to a piece of 300-ohm flat line directly ahead of converter and position the stub on this, instead of on a heavier dielectric-cover UHF line.

6. In tests so far completed, the UHF-VHF coupling devices have been found satisfactory in areas where signal levels are adequately high. However, in weak-UHF and/or weak-VHF locations, the signals do not permit attainment of peak performance and the versatility of the entirely separate VHF and UHF installations. With separate installations each can be oriented and *positioned* for the peak signal. There are no exchange losses at either the antenna or converter if separate lines are run for each antenna.

Chapter 16

TRANSISTORS

189. *Basic Transistors*

The transistor has characteristics similar to those of a vacuum tube and can perform more efficiently many of the tasks of a vacuum tube. Even though its impedance parameters are somewhat different, it can be made to serve the same functions as grounded-grid, grounded-cathode, cathode-follower, and oscillator vacuum-tube circuits serve. It can be made to perform all the video and pulse functions in a television receiver.

Features of the transistor are:

1. Small physical size and light-weight permit small, compact-unit designs.
2. Efficiencies are far above those of vacuum tubes, permitting higher voltage and power gains per supply power.
3. No filament supply is required. Supply voltages need be only a few volts but can range up to a maximum of about 22½ volts with certain transistor types.
4. Estimates of transistor life reach to a point near 100,000 hours.

The transistor is a semi-conductor solid, having characteristics that lie somewhere between those of conductors and insulators. Transistors are made

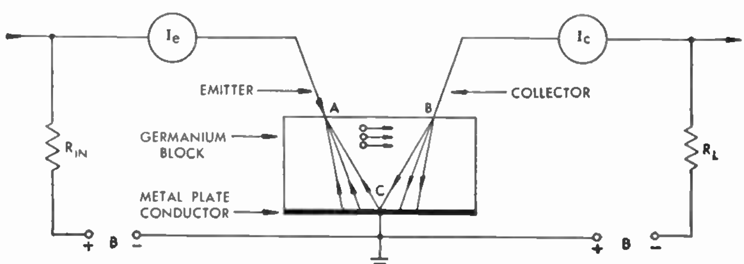


FIG. 416 Basic Transistor Circuit

from germanium, a hard, gray, but brittle, metal that cannot be mechanically worked. A very tiny amount of impurity in germanium enhances its conduction efficiency and determines its characteristics.

A basic transistor circuit, Fig. 416, has input and output circuits, just as a vacuum tube has. The usual transistor circuit has a low-impedance input circuit, a high-impedance output circuit, and is basically a current amplifier, instead of a voltage amplifier as the vacuum tube. Physically, the transistor is a small germanium block mounted on a grounded metal plate. On top of the germanium block are two cat-whiskers spaced just a few thousandths of an inch apart. When a negative potential is supplied to the right cat-whisker, called a collector, there is a current flow through the germanium block to ground. This current, I_c, is really a reverse diode current and is flowing in the high-resistance direction. When a positive potential is applied to the left cat-whisker, called an emitter, a so-called forward diode current flows from ground toward point A and the emitter. The low, forward resistance means the same level of emitter current can be made to flow with a lower applied-voltage level as compared to collector current.

The aforementioned current activity occurs when each side is operated separately. However, when both sides are operated simultaneously an increase results in current flow in the right hand or collector side of the germanium block. This flow represents a current amplification and is the basic underlying condition that causes the transistor to function as an amplifier, oscillator, or generator. It is also this underlying factor that is difficult to understand or, perhaps is merely new, as compared to vacuum-tube operations with which we are familiar. A small voltage change then at the input causes a substantial current change in the collector circuit.

190. *Semi-Conductor Theory*

A better understanding of transistor action can be obtained through a study of the behavior of atomic structures in a solid. In the crystalline structure of germanium, each atom has 32 electrons that move in a continuous and probably orbital motion about its nucleus. We represent this atom symbolically with concentric rings, as illustrated in Fig. 417. Each concentric ring is said to be complete if it contains a particular number of electrons—first ring, 2; second, 8; and third, 18. The fourth ring has only four electrons and is said to be unfilled. The four electrons of the unfilled ring are more unstable than are the electrons of the filled rings, since there is much greater probability of activity by electrons of the unfilled ring than by those of the filled rings. The former are referred to as the valence electrons, and the germanium atom can thus be represented by the second symbol as illustrated.

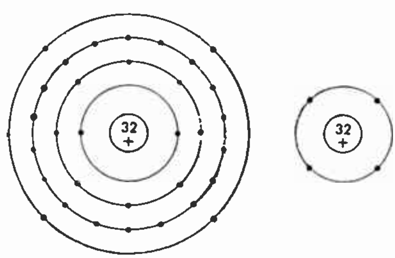


FIG. 417 Symbolic Representation of Germanium Atom

Now in the crystalline atomic structure of germanium the four outer electrons match with the four electrons of nearby atoms, Fig. 418. This germanium atom lattice is also symmetrical, and the electrons of the outer ring are said to be "valence-bound" with respect to the other atoms of the lattice. When electrons are valence-bound there are no free electrons available for conduction, and the material is a non-conductor. In the case of germanium, however, a slight impurity upsets the valence bonding, and under proper conditions there can be current flow. In fact, just a small amount of impurity, one part in millions and millions, permits germanium to become a semi-conductor. If the germanium impurity is an element having five valence electrons, such as antimony or arsenic, electron-conduction is encouraged. The impurity atom, Fig. 419, also joins in the valence bond with the germanium atom. However, when the impurity has five valence electrons there will always be an electron free from bond; this will be a free electron and act as a carrier. This type of impurity, which contributes free electrons, is referred to as a *donor*. Since the majority of carriers are negative particles or electrons, the donor is called an *n*-type semi-conductor material.

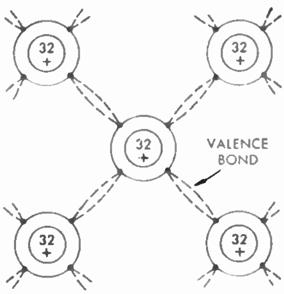
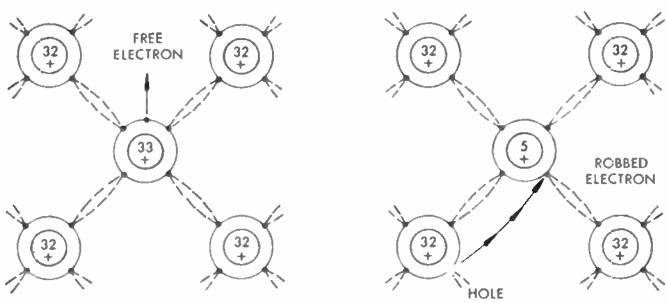


FIG. 418 Partial Arrangement of Germanium Atoms

It is also possible to have a dominating impurity, such as boron, which has three valence electrons. The boron, too, attempts to set up a valence bond with the germanium atom lattice. However, in so doing, it continues to remove one of the atoms from germanium, thus gaining the fourth with which to set up the four-atom valence condition. Thus, throughout the germanium block there is a distribution of holes in the germanium atoms, caused by removal of



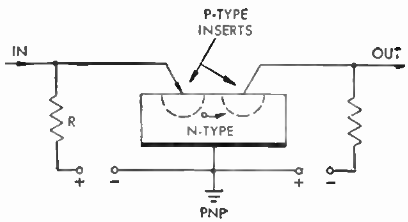
N-TYPE ELECTRON CARRIER P-TYPE HOLE CARRIER

FIG. 419 Electron and Hole-Carriers in Germanium

germanium atoms by the entrance of boron into the lattice. This deficiency of one negative charge (absence of an electron in the outer ring) causes the atom to have a positive charge because of the holes in its construction. In effect, the hole becomes a positive charge and, in effect, can become a moving positive particle in the germanium with application of proper potential. The impurity in this case is called an *acceptor*. We refer to a semi-conductor in which the majority of carriers of current are positive charges or holes as a *p*-type germanium.

191. *Transistor Theory*

There are two basic types of transistors: point contact and junction, Figs. 420 and 421. The point contact transistor consists of an n-type germanium block, metal base, and two point contacts (cat-whiskers). At the points of



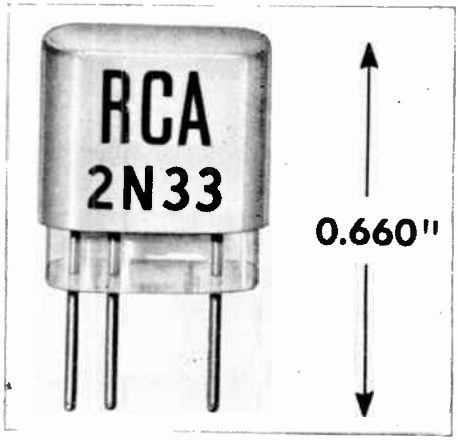


FIG. 420 N-Type Point Contact Transistor, RCA 2N33

contact—for both the emitter and collector—there is a small p-type insert, forming a p-n barrier. Thus, a point contact transistor is referred to as a p-n-p type. This p-n barrier and its unique directional characteristics permit transistor operation.

When a negative potential is applied to the collector, the holes in the p-type insert are attracted toward the collector and the electrons to the base. Since there are few holes in the n-type and few electrons in the p-type, little current can cross the barrier between the two types. As carriers between the p and n types are few, no appreciable reverse current can flow, because of the high resistance presented at the barrier.

When a positive potential is applied to the emitter, it attracts the negative electrons of the n-type while the holes of the p-type can be attracted by the base. Thus, the electron and holes are attracted past the p-n junction, and the barrier is broken down, permitting a high forward current.

However, more than the activities described above occurs when both contacts have the above-mentioned potentials applied to them simultaneously—a plus to the emitter and a minus to the collector. The holes (positive particles) diffuse through the emitter barrier and move toward the collector, where they assist in breaking down the collector barrier, a higher collector flow being permitted—the holes "want" to penetrate the barrier because of the positive-particle pull of the collector. These additional carriers cause a substantial increase in the collector current. The ratio of collector-to-emitter current change is referred to as the *alpha* or *current gain* of the transistor. A stage-voltage amplification is possible, of course, because in the collector circuit a higher current-variation exists across a higher resistance, as compared to a lower current-variation in the lower-resistance emitter circuit.

A second transistor type is the junction type, Fig. 421, which employs non-rectifying contacts, one firmly attached to each of three germanium regions.

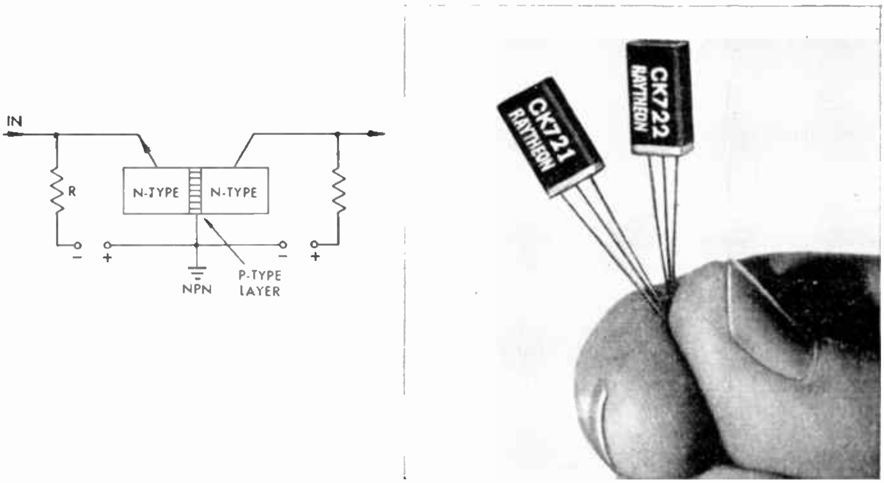


FIG. 421 Junction Transistors, Raytheon CK721 and CK722

In the center of the n-type germanium block is a thin layer of p-type germanium to which the base connection is made. For a junction transistor of an n-p-n type a negative potential is supplied to the emitter, a positive to the collector. In operation, though, the transistor activity is quite the same. The positive collector and its attraction for electrons still cannot pull the p-layer holes free and through the p-n barrier area. Thus, the collector current is lower than in a point contact type because of the much higher reverse resistance of the bulk barrier, as compared to the small close barriers of point contact.

Emitter current is higher because the negative emitter does permit some current flow through the barrier because of its pull on the p-type holes. In fact, the diffused electrons from the emitter n-area pass through both p-n barriers into the collector n-area. Thus there are carriers crossing the barrier, and collector current increases correspondingly. Although there is no current gain (alpha approaches unity), the much higher ratio of collector impedance to emitter impedance (better rectifier action) permits a higher voltage gain as compared to that of the point contact type.

In general, the junction types have higher gain, higher power output, and lower noise level as compared to a point contact unit, while point contact types operate at higher frequencies and can be made very tiny physically. Junction transistors are also available as *p-n-p* types using a *p*-type germanium block.

192. *Practical Transistor Circuits*

There are three basic connections for the transistor: grounded base, grounded emitter, and grounded collector, Fig. 422. Perhaps the most common is the grounded-base arrangement with the input signal applied to the emitter and

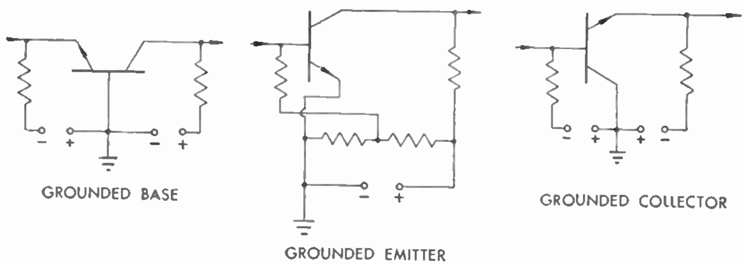


FIG. 422 Basic Transistor Circuits

with the output removed from the collector circuit. This connection has low input impedance, high output impedance, and good gain. Typical values for input and output impedances would be 1000 and 100,000 ohms, respectively.

The highest gain is obtained with the grounded-emitter circuit, which also has a low input impedance and a somewhat lower output impedance than those of the grounded-base arrangement. It does have the highest gain and requires only a single battery for most circuit designs. It is particularly adaptable to lightweight equipment requiring a minimum of battery power.

For obtaining a higher input impedance, the grounded-collector circuit is advised. The grounded-collector circuit has a lower attainable output with a low output impedance.

A typical transistor audio amplifier such as you might expect to find in the audio section of a television receiver, hearing aid, or audio amplifier is illustrated in Fig. 423. The audio input signal is applied through a transformer and volume control to the base of the transistor (a grounded-emitter circuit is used in this example to obtain maximum amplification). Output is removed at the collector junction and applied to an audio output transformer. Capacitor *C1* is a high-value capacitor in terms of the usual interstage coupling circuit, because of the much lower input impedance of a transistor as compared to a vacuum

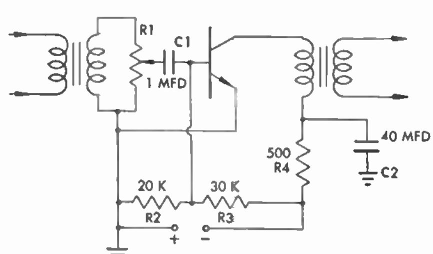


FIG. 423 Transistor Audio Amplifier

tube. The values of the voltage divider resistor *R2* and *R3* are chosen in such a way as to have the proper amount of collector current flow for the transistor's best operating condition (a typical value would be 0.5 milliampere). On occasion, only a single adjustable resistor, which is used in place of the voltage divider, permits correct setting of collector current. In planning a volume-control circuit for the audio amplifier allowance must be made for it to be inserted at a point where moving the control would not change the d-c component of the collector current.

When a multi-stage high-gain audio amplifier is planned it is at times necessary to use de-coupling (resistor *R4* and capacitor *C2*) to prevent oscillation. In the de-coupling circuit it is customary to use a very-low-value resistor in order not to influence the very low transistor current, as developed from the small supply potential. To obtain the necessary de-coupling time constant,

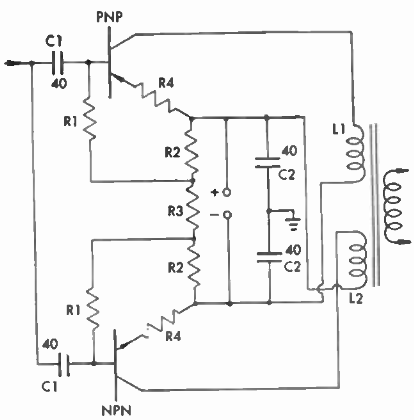


FIG. 424 Push-Pull Audio Amplifier

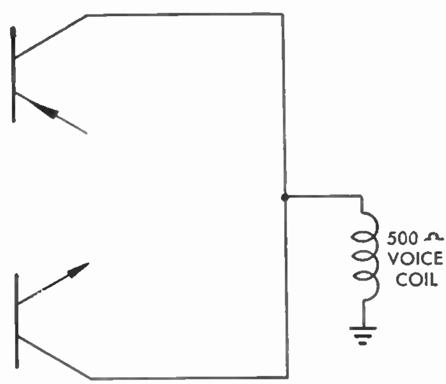


FIG. 424a Transformerless Output

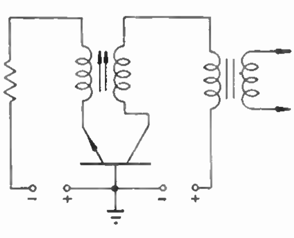
then, the value of the capacitor must be very large. Fortunately because of the low d-c potentials, these capacitors can be very small physically despite the exceedingly high value of capacity. Thus capacitors such as *C1* and *C2* are no larger—in fact smaller—than the usual low-capacity coupling capacitors used in ordinary vacuum-tube practice.

Transistors are also very adaptable to push-pull audio operation, Fig. 424. Transistors have especially inviting features for push-pull operation because of their equal but opposite symmetry. In comparing the *n-p-n* and *p-n-p* types we find that the emitter-collector flow will increase in one type and decrease in the other type, for a given change in base current in the same direction. Thus we can supply two of these units with an in-phase signal (single-ended) and obtain out-of-phase push-pull output components. The push-pull amplifier consists of two opposite junction transistors (*n-p-n* and *p-n-p*) or a point contact and a junction transistor in a grounded-emitter circuit. An in-phase signal through capacitor *C1* is applied to the base connections of both transistors. The *p-n-p* transistor has a plus potential applied to its emitter with respect to the base (voltage divider across a supply source), while a negative potential is applied to its collector (return to the negative side of supply). The junction *n-p-n* transistor in a similar divider arrangement has a negative potential supplied to its emitter and a positive to its collector. When the applied signal goes positive the base voltages of both transistors rise. In the case of the *n-p-n* transistor this means that its effective emitter voltage is more negative, the transistor resistance decreases, and the collector current rises, a drop or negative swing being produced at the top of *L2*. In contrast, the high resistance of the *p-n-p* transistor with the applied positive base voltage causes a negative swing at the top of *L1*. High resistance causes a decrease in the collector current, and *negative* collector voltage rises. Thus, tops of *L1* and *L2* swing in the same direction (push-pull operation). When the applied signal is on its negative swing, the *p-n-p* transistor has a low resistance from the emitter to collector, and the top of *L1* swings positive while the high resistance of the *n-p-n* transistor also causes a positive swing at the top of *L2*. It is interesting to observe that although the a-c signal components in the two windings are in phase, the very small d-c collector currents are flowing in opposite directions. In fact, the output circuit could be only a single inductor, such as a 500-ohm voice coil for a direct-speaker drive and no transformer, Fig. 424a.

Again, large capacity but low-voltage input capacitors must be used to preserve the low-frequency respone at the low impedance input of the transistor circuit. The voltage divider system across the supply points is adjusted to obtain correct collector current. Grounded-emitter circuit must also be properly decoupled at the signal frequency by resistor *R4* and capacitor *C2*—again a long time-constant consisting of a large value *C2* and a small value resistor *R4* that must be small enough to prevent too low an emitter d-c current flow.

TRANSISTOR OSCILLATORS

The transistor can also be used as an oscillator in the audio-frequency range and, in the present stage of development, up to frequencies well over 100 megacycles (a frequency limit that will rise higher and higher as transistor techniques are developed). A simple audio oscillator, Fig. 425, consists of



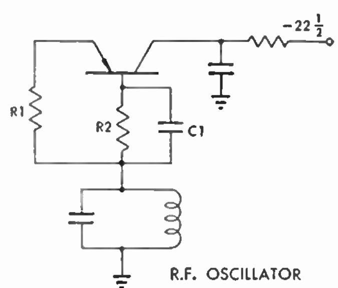


FIG. 425 Transistor Oscillators

transistor, feedback transformer, resistor, and output circuit (earphones or audio transformer). The feedback transformer supplies sufficient transfer of a small signal component between the collector output circuit and the emitter input circuit to sustain oscillation. The resistor is of proper value, as a function of transistor rating, to produce the proper collector-current flow for the junction-type transistor.

With proper transistor design, it is possible to obtain strong enough oscillations for mixer operation at very high frequencies. A typical point contact high-frequency oscillator consists of transistor, emitter current-regulating resistor *R1*, charging time-constant resistor *R2* and capacitor *C1*, collector de-coupling circuit resistor *R3* and capacitor *C2*, and resonant circuit.

When a transistor is connected in a grounded-emitter circuit it can be compared to a triode vacuum-tube circuit (except for impedances) with the emitter acting as the cathode, base as grid, and collector as plate. Thus, if we connect a tank circuit between collector and base, Fig. 425a (plate and grid),

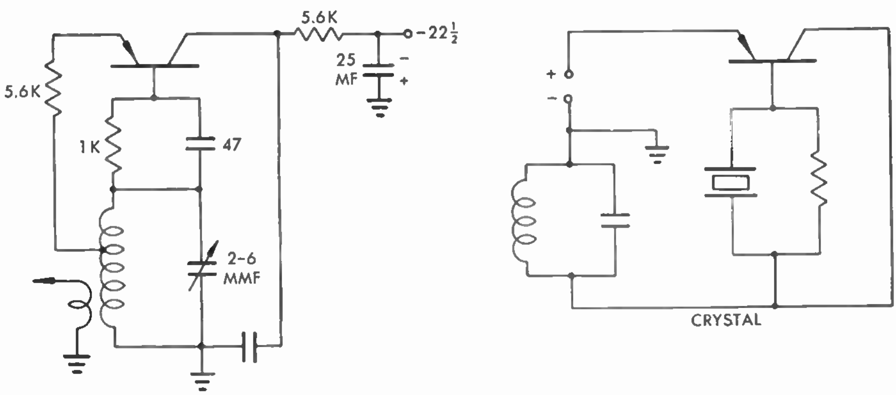


FIG. 425a High-Frequency Oscillators

and properly tap the coil with a lead from the emitter (cathode), a "Hartley-like" oscillator is formed. The tap on the coil supplies a reduced and out-of-phase base signal (as compared to tank voltage) necessary to support oscillation.

A good point-contact transistor can produce oscillations higher than 100 megacycles in this type of circuit. A transistor can also be used in a crystal oscillator circuit.

A high-frequency transistor plus proper high-frequency components in the circuit can also be used for pulse and/or video amplification or for i-f or r-f high-frequency carrier amplification, Fig. 426. Although the high-frequency response and high-frequency gain of the initial transistor circuits are limited, it is only a matter of time until transistors and their associated circuits are developed that will permit wideband high-gain amplification. In a video-amplifier stage using transistors there are a number of factors that make the design a bit different from that in standard vacuum-tube practice. First, the input impedance of the transistor is very low, and every precaution must be taken to preserve low-frequency response when coupling between stages. However, this is not too difficult a problem, because the transistor shows signs

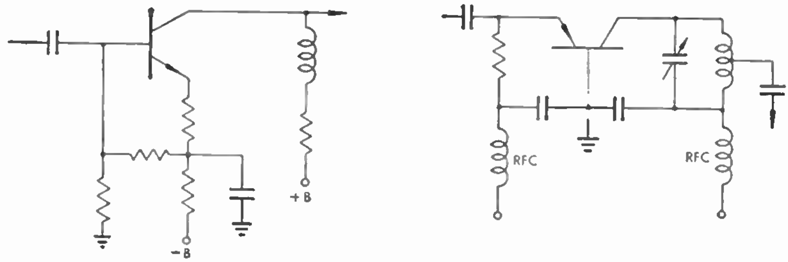


FIG. 426 High-Frequency Amplification

of lending itself well to direct-coupled amplifier chains in which low-frequency response is maintained without difficulty. High-frequency response limitations are overcome with the usual peaking circuit. The proper use of feedback with these low impedance circuits makes more uniform the gain of the transistor stage from one end of the spectrum to the other.

It is possible to construct the very simple stagger-tuned i-f strip, using transistors with special high-frequency characteristics. Such a transistor i-f amplifier consists of just a few component parts—input r-c circuit, proper de-coupling chokes and capacitors, and a tuned collector output circuit. Output is removed from a low-impedance point on the inductor and coupled to the low-impedance emitter input circuit of the succeeding stage.

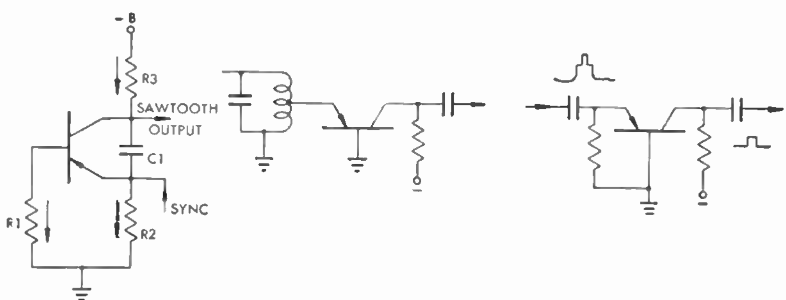


FIG. 427 Special Television Circuits

All the various television receiver circuits can be evolved with use of transistors—sawtooth oscillators, video detector, sync separators, deflection amplifiers, d-c restorers, and a-g-c systems. In a typical transistor sawtooth-generator, the collector current flowing down through resistor *R1* keeps a negative voltage on the transistor base. Likewise, the flow of current through resistor *R2*, during the interval that capacitor *C1* is charging toward minus-B, keeps a negative potential applied to the emitter. However, the current through resistor *R2* falls off exponentially as capacitor *C1* charges slowly. Finally, the negative voltage across resistor *R2* is low enough to produce an emitter potential that is positive with respect to the base. This results in transistor action and an emitter-collector flow that discharges capacitor *C1* quickly. After the discharge of capacitor *C1*, the cycle repeats and is self-sustaining, producing a sawtooth output waveform. Such an oscillator can be triggered by means of a positive pulse suppled to the emitter just before the time of desired syncronization.

A transistor can be used as a video detector or as a sync separator if a grounded base, with the emitter returned directly to the base (no supply potential), is used. In this connection it acts, in the case of the video detector, as a diode detector that is direct-coupled to the collector output circuit, taking advantage of the alpha amplification of the detected signal. With the proper application of a composite video signal of correct amplitude and the use of an input r-c circuit it is also possible to clip sync because of the diode-limiting action of the emitter base circuit. The sync component is also direct-coupled to the collector output circuit and develops a negative-polarity output pulse.

PART II

Chapter 1

COLOR TELEVISION PRINCIPLES

1. *Color Fundamentals*

Light is a variation and has a frequency, just as sound and radio-television waves. Light variations occur at extremely high frequency and stimulate the eye. The wavelength of such a frequency is very much shorter than that of the very highest frequency VHF-UHF wave and has, in fact, a higher frequency than those of our microwave spectrum. For example, the visual frequency spectrum extends from 400×10^{-7} to 700×10^{-7} centimeters. Its short wavelength and high frequency are recognized when we compare them with the 300-centimeter wavelength of a 100-megacycle frequency.

True white light (sunlight) contains all these frequencies, but each individual color has its specific wavelength, beginning with red at the longest wave-

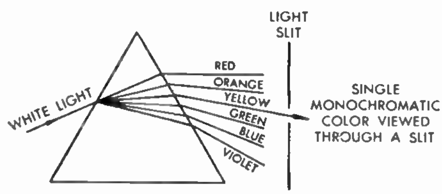


FIG. 1 Prism Formation of Color Spectrum of White Light

length and ending with violet at the shortest wavelength. White light passed through a triangular-shaped prism, Fig. 1, is bent because of the differing velocities of light in air and in glass. Thus at the entrance and exit points of the prism the various colors will be bent by different amounts because of their respective wavelengths. The low-frequency colors are bent less than the high-frequency colors. Therefore, white light in passing through a prism is spread out into a spectrum or band of colors. It is possible, by means of a prism, to view the entire spectra of color between violet and red or, through a very narrow light slit, observe each individual color of the spectra. Color wavelength is generally measured in millimicrons—one millimicron equals 10^{-7} centimeters. Some of the spectrum's basic colors and their wavelengths are tabulated in chart form:

Color	*Wavelength (In Millimicrons)*
Red	650
Red-orange	605
Orange	590
Yellow-orange	583
Yellow	578
Yellow-green	565
Green	530
Blue-green	490
Blue	460
Blue-purple	430
Purple	380

Human vision is not fully understood; the activity of eye and brain is complex, especially in the perception of color. Nevertheless, much information about vision has been gained by experiment, if not by fundamental knowledge. When an observer views the colors of the spectrum through a slit, one color may be separated from another and appear as a single, pure or monochromatic color. For example, if the slit is in position to afford a view of the color at a wavelength of 578 millimicrons, the viewer sees a pure yellow. However, it is a characteristic of human vision that it is able to mix other colors in proper relationship and match a true monochromatic color. For example, an equal mixture of red and green will also reproduce, in terms of human vision, as a yellow. Thus the eye has the ability not only to see a monochromatic yellow but also, and more important from the standpoint of color television, to see a yellow by mixing certain other colors properly. In fact, with the use of three basic, so-called *primary colors,* we can reproduce any color of the spectrum with suitable fidelity. Using the average of many normal observers, the International Commission on Illumination standardized a set of color-mixture curves (CIE), Fig. 2. These standard CIE color-mixture curves give the relative intensities of the three primary colors: red, green, and blue, required to match a unit amount of radiant light having a specified color wavelength. For example, in the case mentioned previously, the chart in Fig. 2 indicates that to produce a yellow (frequency of 578 millimicrons) it is necessary to match equal intensity levels of red and green. Refer to the intersection of the red and green curves on the chart.

The levels represent the intensities of each primary required to match the same intensity of monochromatic color at a specified wavelength. In other words, if, observing a monochromatic yellow with a certain brightness, we want to match it by use of red and green primaries, we would require red and green primaries of equal intensities, each having an intensity of 0.88 of intended monochromatic brightness. It is interesting to know that, despite the equal, measured intensities (lumens) of the red and green, the human eye itself has an apparent brightness sensitivity that causes the green to appear more than twice as bright as the red when each is viewed alone.

In fact, after the formation of a white from equal intensities of red, blue, and green, if we were to observe the component green by itself, it would appear to us twice as bright as the red and perhaps twenty times as bright as the blue. The brightness sensitivity of the normal eye as a function of color follows the shape of the green primary curve. A calibration on the right side of Fig. 2 shows the apparent sensitivity of the eye to the color spectrum, with a peak in the region of green.

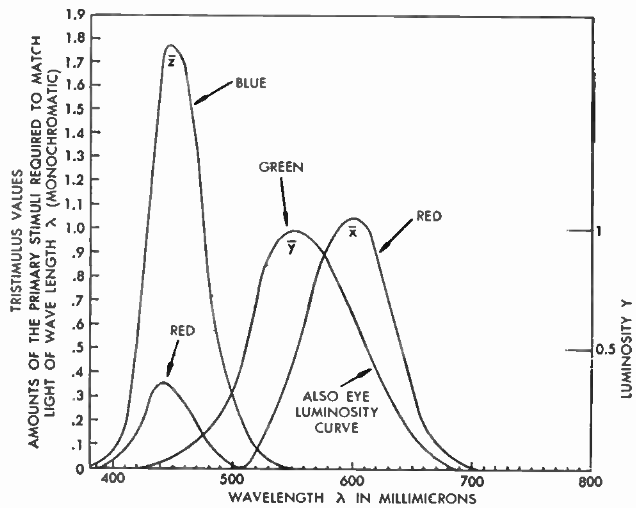


FIG. 2 Standard CIE Color-Mixture Curves

With the choice of three suitable colors (no two of which can match the third) any color can be matched by using the proper relative level of the two or three chosen primary colors. The most common primaries used are red, blue, and green, which spot a primary at each end and at the center of the color spectrum. Furthermore, with the use of equal levels of red, blue, and green it is possible to produce an apparent white. Consequently, a color television system based on the three primary colors of red, blue, and green is able to synthesize any color and also permits conveyance of whites, grays, and blacks—"grays" and "blacks" being low-brightness "whites."

COLOR ATTRIBUTES

Color is evaluated in terms of three factors:

1. *Hue*—Hue is a property of color in terms of its redness, yellowness, and so forth. In terms of a pure monochromatic color it represents wavelength. In terms of a color television system it represents one of the three

primary colors or the proper mixture of primary colors to produce a desired *secondary* color.

2. *Saturation*—Color saturation refers to the purity of the color and determines the degree of vividness (whether it is a deep or a pastel shade) of a given color. Color saturation refers to the degree of dilution of the pure color with white. A light or pastel shade of a given hue means an appreciable dilution with white light, while a very dark or saturated color of a given hue means a minimum of dilution.

3. *Brightness*—Brightness or luminance is one characteristic of color that is also associated with the normal transmission of a black-and-white picture and refers to the intensity of light reflected by the colored object.

These three aspects of color must be conveyed by a television color system in reasonable facsimile. It is not possible to attain complete saturation of every hue with the use of the three primary, or tristimulus, method of color transmission. However, full-color saturation (absence of white in a color) is seldom found in nature or in objects normally televised. Thus, the inability to transmit highly saturated colors represents no significant problem in a color television system.

CHROMATICITY CHART

A two-dimensional plot of the information contained in the mixture chart, Fig. 2, produces the standard CIE chromaticity chart, Fig. 3. The horseshoe curve, referred to as the *spectrum locus,* outlines the region of realizable or possible colors, by use of tristimulus presentation. The primaries indicated by the corners of the triangle are unreal and were assumed merely to permit the construction of a two-dimensional diagram as a mathematical transformation.

The x and y terms are referred to as trichromatic coefficients of the chromaticity diagram. Within the spectrum locus, point W represents a location where trichromatic functions x and y equal 0.3333 and correspond to white light. If we now draw a straight line from W to, let us say, the yellow wavelength on the spectrum locus, we delimit the range of yellows from very light to very deep. At the point where the line crosses the locus we have the saturated yellow, while approaching point W we have increasingly light pastel yellows (or tints) with correspondingly large percentages of white dilution. Thus the diagram indicates both the so-called dominant wavelength of the color as well as the degree of saturation or purity—purity being measured over a range from zero at point W to one (unity) at the spot where the line crosses the spectrum locus. A chromaticity diagram represents the hue of a color in terms of dominant wavelength as well as the color saturation in terms of purity or dilution with white.

For example, with the chart it is possible to decide dominant wavelength and purity for a point such as A. First, a line is drawn from W through A until it crosses the spectrum locus at 530 millimicrons. This point of intersection represents the wavelength of green and, inasmuch as point *A* is half the distance between *W* and the intersection point, the degree of saturation or purity is represented by the factor 0.5. It is not possible for a color television system to transmit the entire range or gamut of colors enclosed by the spectrum locus. In fact, it is not possible to duplicate every color with color photography or with color printing, as now employed. Rather the color television gamut rec-

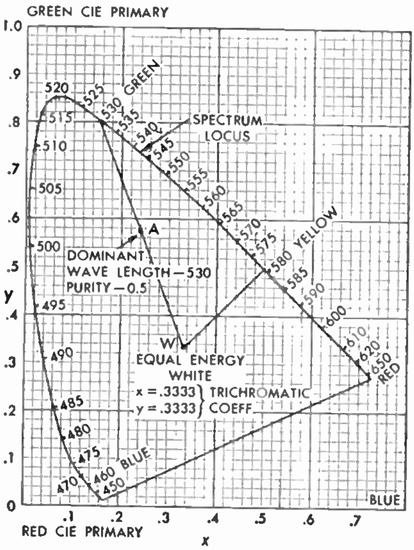


FIG. 3 CIE Chromaticity Diagram

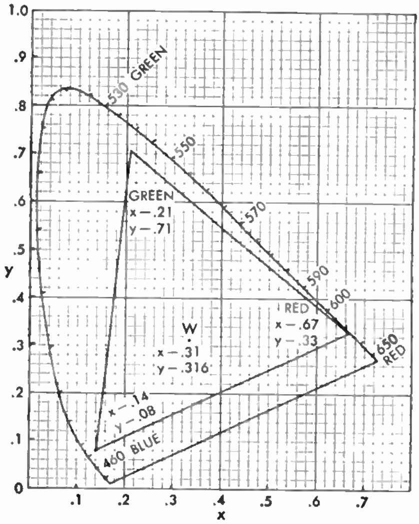


FIG. 4 Chromaticity Gamut of NTSC Color System

ommended by the National Television Standards Committee is enclosed within the triangle shown in Fig. 4. Notice that it is not possible to transmit saturated blues and greens, although it is possible to transmit reasonably saturated colors in the yellow-to-red spectrum range. The NTSC decision was made because of the tendencies to greater saturation in nature over the yellow-to-red range. It is significant to realize that this range of color rendition in television is superior to results in color printing and is comparable to those obtained in color photography.

COLLECTION OF COLOR INFORMATION

In color television the scene and objects to be televised are illuminated by white light, which contains the complete range of colors. An object illuminated by white light has the property of reflecting wavelengths represented by its color and rejecting (or absorbing) other colors (or wavelengths). Thus, a red object when illuminated by white light, Fig. 5, will reflect a red wavelength but other colors to a lesser degree or not at all. This red "information" is carried back to the lens of the color camera, and on its way passes through a red filter which prevents the entrance of other colors and illumination.

When a yellow object is illuminated by white light, it reflects dominantly a yellow wavelength (light) as well as other colors to a lesser degree. The color filters of the color camera must be arranged and chosen so that the reflected light from the yellow object creates equal intensity signals in passing through the dominantly red and dominantly green filters. Therefore, yellow light is evaluated in terms of the red and green filters of the camera. Consequently, red, blue, and green filters of the television camera must be properly chosen, and proper gain controls must be included in the corresponding color amplifiers to permit the matching of the gamut decided upon in the color television system standards. In summary, the red, green, and blue information in the

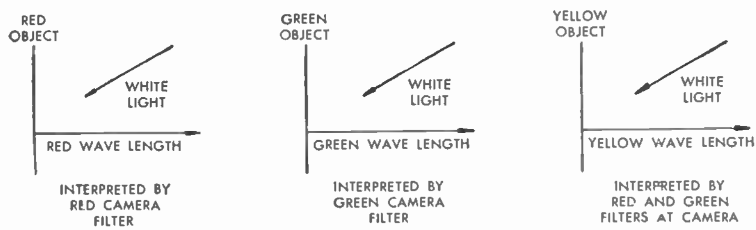


Fig. 5 Color Reflection from Objects

televised scene is evaluated by red-, blue-, and green-sensitive cameras, while the secondary colors are evaluated in terms of the impressions that their reflected wavelengths make on combinations of the red-, blue-, and green-sensitive cameras. Thus it is the responsibility of the color-pickup equipment to respond to all colors in terms of relative levels of red, blue, and green; proper responses enable colors to be reproduced at the receivers in their proper relative relationships, thus re-creating the colors televised.

Color television uses a so-called additive process of reproducing the color picture in which the images in each of the three primary colors are either projected simultaneously one over the other onto a screen or are combined on a single direct-view screen. We see the various secondary colors by an additive process of relative levels of the primaries.

2. *Methods of Color Transmission*

In a color television system it is possible to transmit the three basic colors simultaneously or to send them one color at a time in a sequential order, Fig. 6. In the simultaneous system three separate color pictures are picked up by the color camera and transmitted as signals continuously through the transmitter to the receiving location. At the receiving location the three signals are segregated and applied separately to the tricolor tube or to a group of three individual tubes, whose outputs are projected as pictures on a viewing screen. In the sequential method of color transmission three separate signals are formed by the camera and are commutated, only one signal at a time being conveyed from transmitter to receiver. A switching arrangement is used to transmit individual primary colors in a sequential order. At the receiver the three primary signals are again separated and passed through their own individual channels to the tricolor tube or color projection system. They are then focused sequentially on a single viewing screen. However, if they are applied at a fast enough rate, persistence of vision enables viewers to observe each set of three as a single, continuous, color picture.

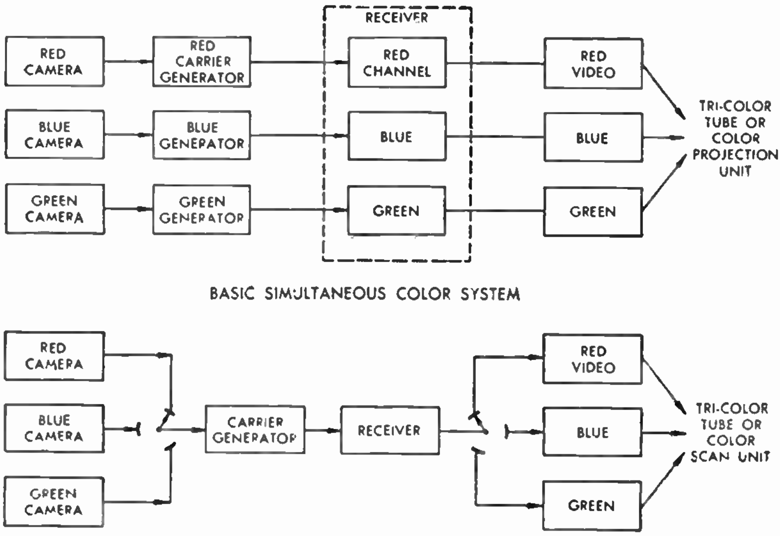


FIG. 6 Basic Methods of Color Transmission

SEQUENTIAL COLOR TRANSMISSION

There are a few basic methods of sorting primary colors sequentially (making the camera sensitive to each of the basic colors in a definite order for transmission). In such a color system, not only is each picture element scanned in relation to its intensity or luminance but also in relation to its color or chromaticity. The rate of color shift or interruption need not, although it can, be the same as the rate of element transmission. Entire groups of elements can be first transmitted in one color and then, a bit later, in still another basic color. If a sequential scanning system were used (no interlace), it would be possible to transmit three successive frames, each in one of the basic colors. This process would be repeated at a specific frame-rate per second. In any color system, however, it is necessary to have the color-interruption rate fast enough to prevent flicker because of the inability of the eye to hold the image for a slow rate of color interruption.

It is common practice in television broadcasting and even in closed-circuit television to use a two-to-one interlace. Under these conditions it is possible to color-shift each field. In fact, this method of color-interruption is known as field sequential. Color, of course, interlaces with the scanning, Fig. 7; at the end of six fields (three frames), therefore, each element has been scanned in each of the basic colors. In such a field sequential system, color changes would occur as follows:

First field:	Odd lines in red
Second field:	Even lines in blue
Third field:	Odd lines in green
Fourth field:	Even lines in red
Fifth field:	Odd lines in blue
Sixth field:	Even lines in green

Thus, in six fields each element and line has been scanned in each basic color as it should be for good color fidelity.

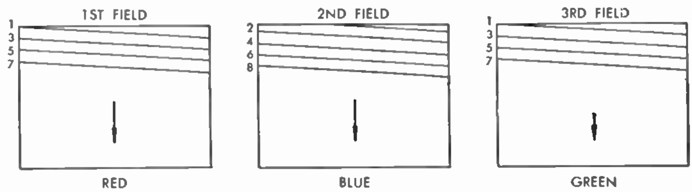


FIG. 7 Field Sequential Color Scanning

A second method of color scanning involves a system for changing colors for each individual scanning line, Fig. 8. This system also requires six fields of a two-to-one interlaced system in order to obtain a complete television color picture, as each line must be scanned in each color. Both odd and even lines must be scanned as follows:

First field:	Odd lines in order	Red, green, blue
Second field:	Even lines in order	Red, green, blue
Third field:	Odd lines in order	Green, blue, red
Fourth field:	Even lines in order	Green, blue, red
Fifth field:	Odd lines in order	Blue, red, green
Sixth field:	Even lines in order	Blue, red, green

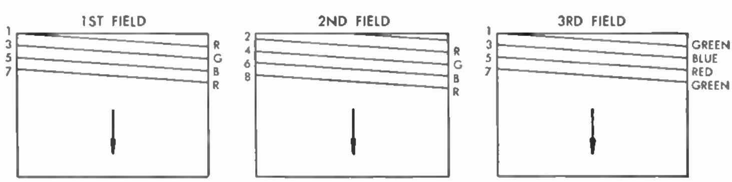


FIG. 8 Line Sequential Color Scanning

It is to be noted that the rate of color interruption is much faster in this method, being the same as the line rate of the television system. Consequently, the line sequential system (change of color each line) is so much faster than field sequential (change of color each field) that mechanical means of color interruption cannot be employed. Instead, the actual commutating, or switching between the basic colors, must be accomplished by means of fast-acting electronic circuits.

A still faster means of color shift for interruption is dot sequential, which breaks up each line into a series of color dots, Fig. 9. Thus the several hundred elements along each line are observed in the color order: red, green, blue. This same process continues right on down the raster, and dot sequence continues as follows:

First field:	Odd line elements in order	Red, green, blue
Second field:	Even line elements in order	Blue, red, green
Third field:	Odd line elements in order	Blue, red, green (Element position has been shifted—new groups fall between elements scanned previously.)
Fourth field:	Even line elements in order	Red, green, blue (Same element arrangement as in third field.)

In the dot sequential system the rate of color switching is extremely fast—switching at a rate of between 8 to 12 million times per second. Consequently, proper timing is a complex problem in this type of color interruption. Of the three sequential methods only field sequential interruption occurs at a slow enough rate to permit the use of a mechanical color-interruption system. All three systems can use electronic means of color interruption, but with the line and dot sequential methods electronic means of interruption must be used.

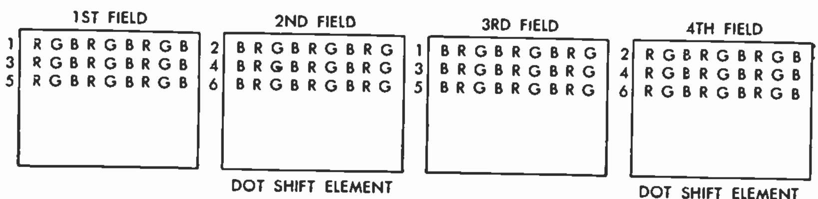


FIG. 9 Dot Sequential Color Scanning

3. *CBS Field Sequential Color System*

The color system developed by Columbia Broadcasting System was not the method finally accepted in 1953 by the Federal Communications Commission as a means of color broadcasting in the United States. The major objection to the system was its incompatibility and poorer resolution as compared to more complex color television systems. The CBS standards were not compatible, using different line and field rates than now exist in our commercial monochrome broadcasting. Consequently, it was not possible to view the field sequential color transmission as a monochrome or black-and-white picture on a standard television receiver which had no previous circuit conversion. The resolution is also somewhat less because of the fewer lines per frame and the limitation on frequency response, compared to our present standards.

Although the field sequential system was not found acceptable for commercial color broadcasting, it is a simple and inexpensive means of obtaining a high-fidelity color picture. Thus the method is likely to see wide use in closed-circuit and industrial color television systems. In its simplest form, Fig. 10, it consists of a black-and-white camera, conventional except for a rotating color filter in front of the camera. This rotating disc contains transparent color filters that permit light to pass on to the photosensitive surface of the camera tube. The intensity of the light that reaches the sensitive surface is a function

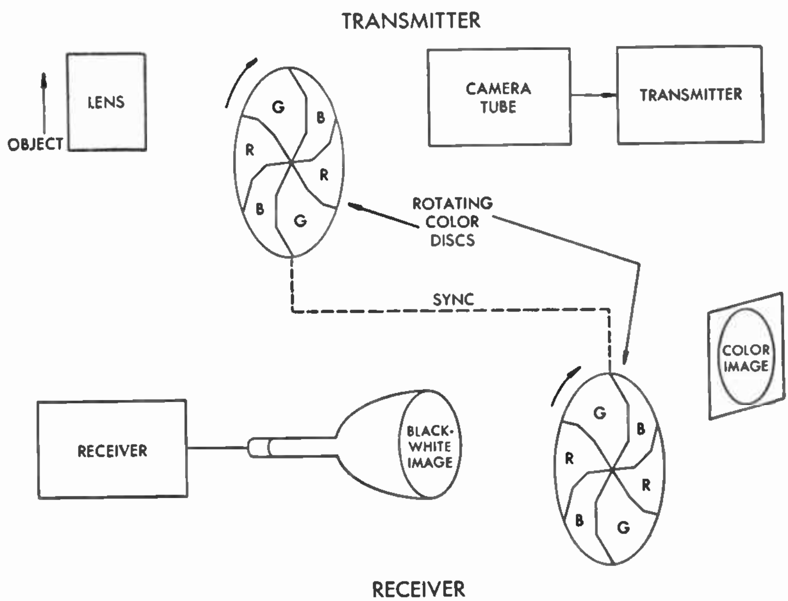


FIG. 10 Field Sequential Color System

of: 1) the color of the object, 2) the color of the part of the filter which is in front of the camera at a particular instant, and 3) the ambient lighting at the pickup point. If the object is red, red light is reflected to the camera; when the red filter moves into position, the red light will pass on to the sensitive surface of the tube forming a signal representative of the object. At the receiver, this signal will "paint" a black-and-white picture of the object on the fluorescent screen; however, at the very instant it appears, the picture will be viewed through a red filter and will appear as a red object to the audience.

Likewise, each of the basic colors contributes its own signal that is sent to the receiver sequentially and is viewed through a proper and similarly colored filter. It is a fact that three separate images are conveyed from the transmitter to the receiver—each image representing one of the basic colors. As the color filter disc rotates each color image passes quickly in review, the colors blending to form one color picture of good fidelity. This good fidelity is obtained only when the rotating disc at the receiver is synchronized with the rotating device at the transmitter. When the red filter is in position at the transmitter, it is interpreting red from the televised scene—at the same instant, at the receiver, it is necessary to be viewing the image through a red filter. Thus, it is not only a problem of sending the color information but also proper utilization of a signal component in synchronizing the mechanical rotating device at the receiver with a similar mechanical color-interruption device at the transmitter. If a yellow object is to be televised, the yellow light it reflects excites, to a degree, both the red and green filters as they appear in the light path between

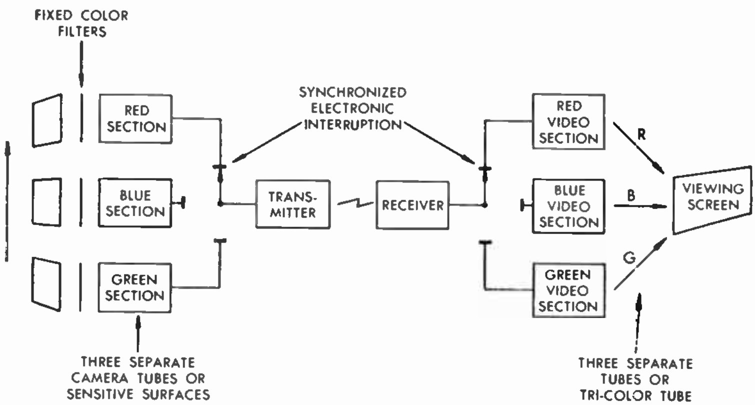


FIG. 11 Field Sequential Scanning with Electronic Interruption

object and camera tube at the transmitter. Thus both the red and green images have been excited by the yellow light, and at the receiver, a picture is present on the picture-tube screen during both the red and green intervals. An observer, viewing through the filters, will then see red and green blending to form the original yellow.

In summary, the rotating color disc contains segments of red, blue, and green filters that rotate between the lens and camera tube. When the red filter is in position the scanning beam of the camera tube generates signals that vary in amplitude in proportion to the varying intensities of the red components of the scene. The same process is repeated for subsequent blue and green filter segments. At the receiver a rotating color disc is positioned in front of the picture tube, and, if this disc is rotated at the same speed as and is in position corresponding to that of the camera disc, color images pass in review in the same order as they were accepted at the transmitter.

It is not necessary to use a rotating color disc to reproduce the color picture in a field sequential system. Instead, signals can be channeled to a set's three separate picture tubes and, through proper color filters, projected on the viewing screen, Fig. 11. A tricolor direct-view tube can be used as well in the reproduction of the color picture.

The standards suggested by CBS as compared to our black-and-white standards are tabulated in chart form.

	CBS Color	*Monochrome*
Field rate	144	60 fields
Line rate	29,160	15,750 lines
Lines per field	202½	262½ lines
Lines per frame	405	525 lines
Frame rate	72	30 frames
Interlace	2 to 1	2 to 1 interlace
Field time	6,944 microseconds	16,667 microseconds
Field blanking	347–555 microseconds	833–1333 microseconds
Field retrace	less than 347 microseconds	less than 833 microseconds
Line time	34.3 microseconds	63.5 microseconds
Line blanking	5.5–6.2 microseconds	10.2–11.4 microseconds
Line retrace	4.5 microseconds	8.3 microseconds

In addition, several more specific color standards are as follows:

1. A three-color sequence in order of red, blue, and green.
2. Six fields are needed to send a complete color picture with a color picture rate of 144/6 or 24 per second.
3. The color disc consists of six color divisions or segments (in order of red, blue, green) and is rotated at a speed of 1440 revolutions per minute. The total number of segments per minute would then be 6 times 1440 or 8640. At this rate, there are 144 segments per second (8640/60), a rate which corresponds to the field rate of the color standards (one color per field). Thus each color segment of the rotating filter represents one field of our color picture. This relationship between color segments and field rate means that the vertical pulse used to synchronize the vertical scanning rate of the receiver can also be used to time the motion of the rotating disc.

4. *Simultaneous Color Transmission*

In a simultaneous color system, all three of the primary channels are operating continuously. In a basic simultaneous system there would be three separate color cameras and three separate channels feeding the modulation to three separate carrier generators. In effect, we would be transmitting three entirely different signals, one for each color. At the receiver the three individual signals would be picked up, again channeled separately, and used to excite a tricolor tube or a three-tube color-projection unit. Thus in the basic simultaneous system the three signals remain separate from each other and are combined only at the viewing screen.

It must be emphasized that use of such a system would require approximately three times the bandwidth necessary for the transmission of a single signal. Consequently, in the initial days of color experimentation the use of a simultaneous system did not appear feasible. However, much has been learned about the characteristics of color and the necessary information that must be conveyed for reproduction of a color picture. Two most important discoveries were the facts that only one signal is needed to carry the full resolution of a picture and that color information can be conveyed on a much narrower bandwidth than that used originally without harming color resolution. It has been possible, using a simultaneous system, to transmit a high resolution color picture in the usual assigned six-megacycle channel.

This conservation in frequency spectrum has been made possible by a more complete knowledge of response of the optic nerve of the human eye to color stimuli and by a thorough investigation of the minimum amount of information that needs to be conveyed in order to allow a viewer to reconstruct a good color picture.

Some of the most significant factors have been:

1. In the transmission of a color picture it is necessary to transmit the hue and saturation as well as the brightness information to reproduce the *large area* color information in a scene.

2. For *small area color detail* in the same picture it is necessary to transmit only the brightness information along with some limited indication of color saturation.

3. In the transmission of the *very fine detail* contained in same color picture it is necessary to convey only the brightness or luminance information.

Thus, the detail in a color picture is a function of brightness or luminance, and it is necessary to transmit only this information in order to obtain adequate resolution in a color picture. It is a fact that as the small detail becomes finer and finer the eye no longer sees that detail as a specific color but only as a brightness variation. As detail is made finer and finer human vision cannot distinguish hue, and colors are perceived as grays. It is therefore quite possible to convey the color information of a picture at a much lower resolution limit than that required for the luminance information, which truly carries the fine detail in a color picture. This circumstance leads to two basic conceptions employed in our commercial television system.

A transmitted color signal must contain luminance and chromaticity information. The luminance information must be conveyed with high resolution if proper pictorial detail is to be obtained. The less detail in which chroma (hue and saturation) can be conveyed will form, when combined with the luminance information, a well-resolved color picture. It is possible, then, to transmit one of the primary signals at full bandwidth, obtaining picture detail and luminance. The other two signals can each be conveyed with a lesser resolution, permitting a saving in bandwidth. A more suitable method would be to combine the three primaries in two specific signal groups, transmitting one with detailed luminance information and the second with chromaticity information at lesser bandwidth.

To go a step further, it is conceivable that the basic chromatic information could be used to modulate a sub-carrier which could be properly positioned in the bandwidth spectrum of the luminance signal. Thus in a single 6-megacycle channel it would be possible to transmit the brightness information as well as chrominance information of the three primary colors.

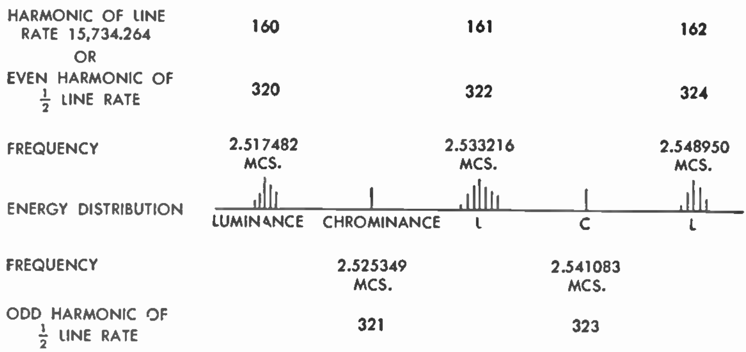


FIG. 12 Energy Distribution over Small Segment of Video Spectrum

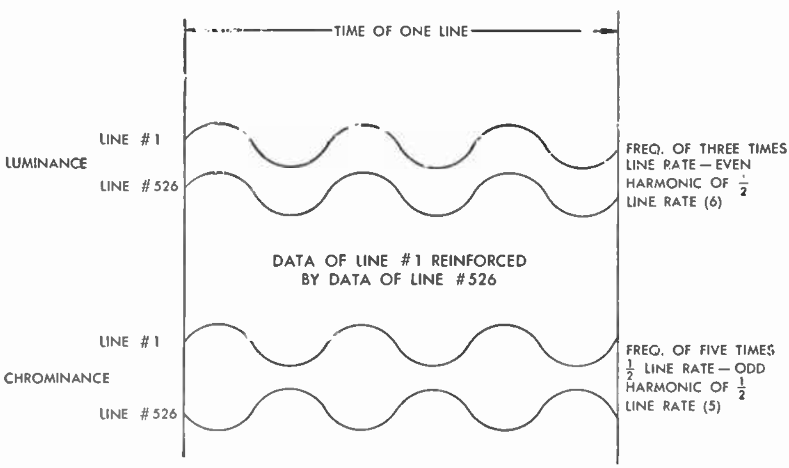


FIG. 13 Frequency Interlace Cancellation of Chrominance Crosstalk in Luminance Channel

It is possible, because of the spectrum distribution of a typical television signal, to position within the video spectrum a sub-carrier of luminance information without interference arising between the two signal groups. Spectrum analysis of a typical video signal shows that the energy and information in that signal are concentrated at frequencies which are whole harmonics of the line rate. Consequently, the video information in the television signal is grouped at harmonics of the line rate (15,750) throughout the entire video spectrum up to the 4-megacycle limit. In between these harmonics, all through the video spectrum there is no useful signal information. Thus it is conceivable that, with the proper location of a sub-carrier, additional information can be conveyed within the same frequency spectrum. Accordingly, a sub-carrier frequency, which is an *odd harmonic* of *one half the line rate* (always located halfway between two adjacent line-rate harmonics in the video spectrum) is chosen. If this sub-carrier is now modulated with video information, its own sidebands will position or "interleave" themselves between the line-rate harmonics of the initial information and thereby fill in the gaps with useful picture information, whose transmission will then require no increase of the necessary bandwidth. This process is also referred to as "frequency interlacing." In summary, the use of a more restricted bandwidth for the transmission of the chrominance information and of a sub-carrier permits not only the transmission of a color picture within a 6-megacycle channel but the same quality of picture resolution possible in a standard monochrome transmission. The chrominance information is so positioned that it does not interfere with the normal brightness and luminance data which carry picture detail and resolution.

The insertion of chrominance information at odd harmonics of one half the line rate, Fig. 12, also prevents the appearance of a dot pattern, which can be caused by interaction between the chrominance and luminance signals. Such a disturbance would make itself apparent as a dot pattern on a monochrome rendition of the color signal. Although the crosstalk does exist it is cancelled by the frequency interlace action that results from choice of a sub-carrier frequency that is an odd harmonic of one half the line rate.

The luminance information (picture detail) occurs at integral harmonics of the line rate, and consequently, if information exists at this frequency rate, Fig. 13, there would always be a number of full cycles existing during a single-line interval. For example, if the video information along line number *1* were bunched at the third harmonic of the line rate, there would be three complete cycles of this information occurring during the line time. Exactly one frame later or at line *526,* the same three cycles would exist and reinforce one another. Now, if we had chrominance information that existed on a frequency five times the half-line frequency (chrominance information occurs at odd harmonics of one half the line rate), there would be 2½ cycles of information included in the line. Exactly 525 lines later the same 2½ cycles would occur, but then they would be opposite in polarity. From this we see that a possible brightness variation along the line caused by the chrominance signal is effectively cancelled and that dot structure is far less apparent. Pattern is not removed completely in dark areas because of the cutoff characteristic of picture tubes. However, dot structure is not discernible at normal viewing distance.

Chapter 2

NTSC COLOR SYSTEM

5. *NTSC Color Transmission*

After a general over-all description of the functional plans of the NTSC Color System is presented, in conjunction with the block plan of Fig. 14, details of the system and standards will be discussed. At the transmitter a three-color camera arrangement is used to pick up the three basic color primaries. These primaries are applied first to a signal mixer or matrix unit. It might at first be thought proper to transmit all three colors at equal amplitude levels. However, it is a fact that the human eye is more sensitive to some colors than to others, and it has been found advisable to change the relative levels of the colors prior to transmission. These changes permit a method of transmission known as constant luminance, which will be discussed in more detail shortly.

There are three signals available at the output of the matrix unit. First, there is the *luminance signal,* consisting of red, blue, and green signal components in proper proportions. Used to convey picture detail and brightness variations it is referred to as the Y or luminance component. In addition to information pertaining to luminance, we must transmit the hue and saturation information or chrominance data. The second output from the matrix unit is a $B - Y$ component, or blue-chrominance component, which is first applied to a low-pass filter and then to a balanced modulator. A third component, referred to as the $R - Y$, or red-chrominance component, is applied also to a separate low-pass filter and balanced modulator. These two signal components, $B - Y$ and $R - Y$, contain the chrominance information that is to be transmitted. It must be pointed out here that it appears as though only red and blue chrominance information, in the form of $B - Y$ and $R - Y$ signals, is supplied. However, as will be discussed in greater detail shortly, it is possible to combine the $R - Y$ and $B - Y$ signal components, thereby forming a green resultant signal at the receiver. Thus, in these two signals we have the chrominance information necessary for the three primary colors.

These two chrominance signal components are used in modulating a subcarrier frequency that is located near the high-frequency end of the video spectrum of the NTSC color signal, Fig. 15. The frequency chosen has been made high enough to prevent the appearance of any interference pattern on the receiver screen but low enough to permit optimum operation when a picture is to be reproduced in monochrome. The sub-carrier frequency, approximately 3.58 megacycles, and its sidebands occupy a span of frequencies between 2 megacycles and 4½ megacycles above that of the picture carrier. The chrominance information in this span of frequencies is interspersed with the luminance information in a frequency interlace form as discussed earlier—the 3.58-megacycle sub-carrier frequency is an odd harmonic of one half the line rate.

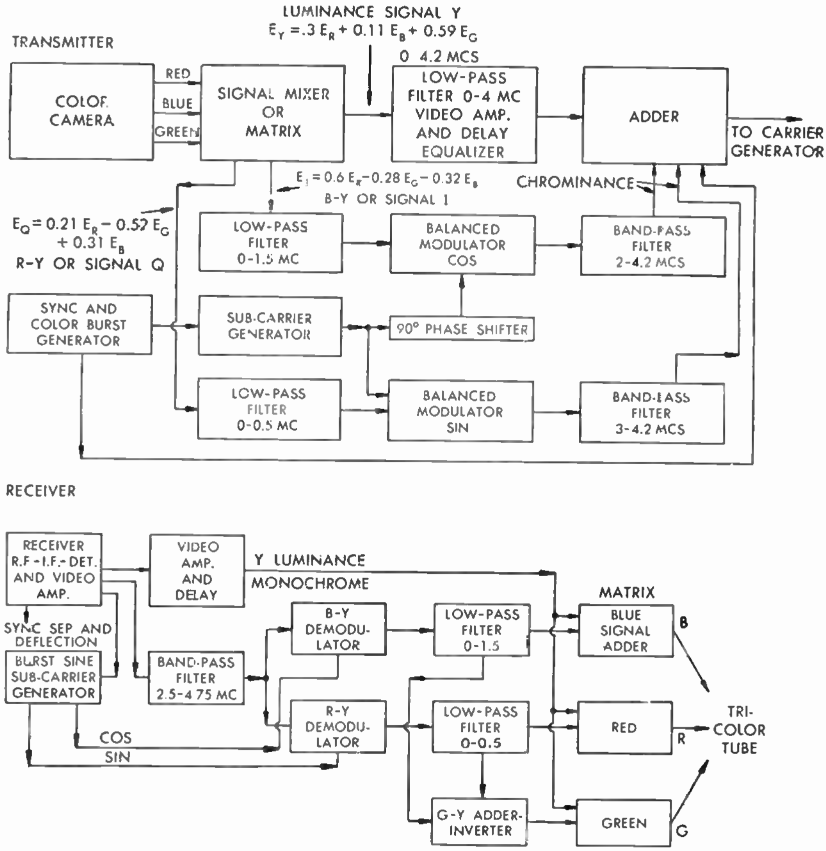


FIG. 14 Functional Block Plan of NTSC Color System

The simultaneous pair of chrominance components amplitude-modulate the sidebands of a pair of suppressed sub-carriers in quadrature, which have a common frequency near 3.58 megacycles (actually 3.579545 megacycles). In modulating one of the sub-carrier sine waves the $B - Y$ component has a bandwidth of approximately 1.5 megacycles, and its energy is concentrated on the low-sideband side of the sub-carrier signal. The $R - Y$ chrominance signal modulates the second sub-carrier sine wave (90 degrees related to the first) but has a more limited bandwidth, approximately ½ megacycle, and forms a double-sideband signal. In this double-sine wave modulation system, the color saturation is determined by the amplitude of the modulation and the hue by the phase relationship between the sub-carrier sidebands. As modulation takes place the actual sub-carrier is suppressed and is not transmitted. The sub-carrier is again inserted by a proper sine wave generator at the receiver.

Thus, in the transmitter section of the block diagram, Fig. 14, we notice the $B - Y$ component is first applied to a low-pass filter (upper limit of 1.5 megacycles), then to a balanced modulator, and finally to the bandpass filter which allows only the sideband components between 2 and 4 megacycles to reach the output. The $R - Y$ component is applied to a second low-pass filter with an

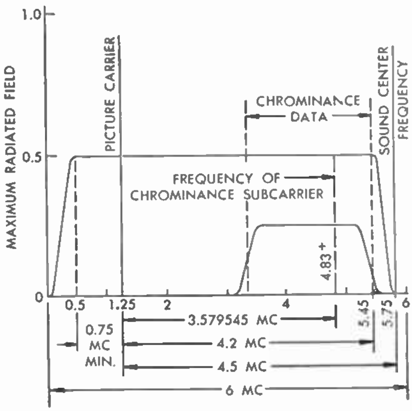


FIG. 15 Idealized Amplitude Characteristic

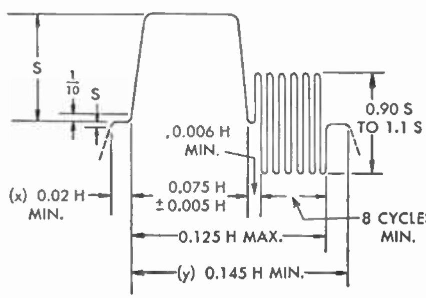


FIG. 16 Color Sync Burst on Backporch of Horizontal Sync Pulse

upper limit of only ½ megacycle and then to a balanced modulator and bandpass filter output circuit. Associated with the modulation section is a sub-carrier generator whose timing is under the control of the sync generator of the station. It forms the sub-carrier frequency that is modulated by the $R - Y$ component. This sub-carrier frequency is also passed through a 90-degree phase shifter to the balanced modulator at which it is modulated by the $B - Y$ chrominance signal.

The luminance and chrominance signals are combined in an adder or mixer stage. At this mixer system, the composite sync waveform is also introduced. This sync waveform is the same, with the exception of somewhat closer tolerances and the addition of a sub-carrier burst on the backporch of the horizontal sync pulses, Fig. 16, as that now existing for present monochrome transmission. The burst consists of a minimum of 8 cycles of the sub-carrier frequency and is used at the receiver to time properly and generate the sub-carrier frequency components that are to be used in demodulating the chrominance information. This constitutes the only synchronization data that must be added to the standard television signal to lock in the chrominance operations at the receiver.

The remainder of the transmission process continues as per conventional techniques except for the closer tolerances in regard to bandwidth and phase response.

The color signal is also received as per conventional practice which uses the superheterodyne principle, employing tuner, i-f amplifier, video detector, and video amplifier. Again alignment tolerances are more exacting and phase response is or more significance and importance than in standard monochrome

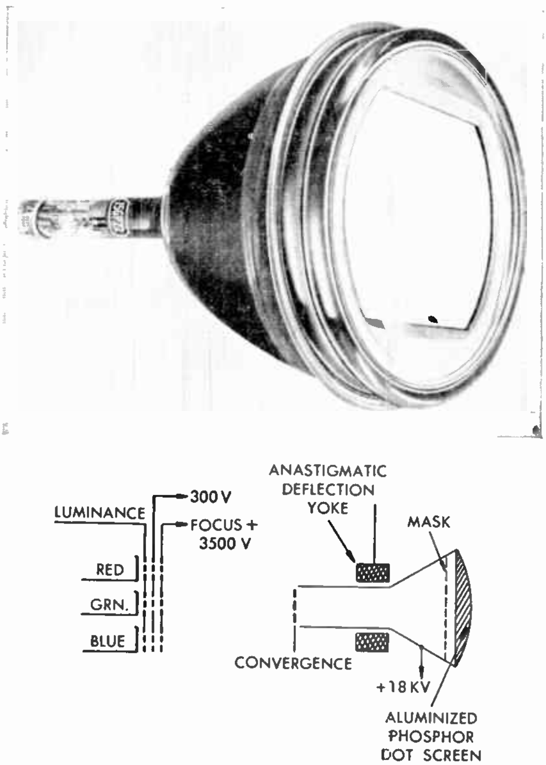


FIG. 17 General Plan of Tricolor Tube

reception. The deflection and the sync separator and amplifier systems are similar to those in conventional practice and are, therefore, not expanded on the block diagram. The luminance, or Y-signal, component is applied to a video amplifier, which has delay characteristics suitable for timing the luminance signal properly with respect to the two chrominance signals and also for adjusting the time delay of the low- and high-frequency components of the signal properly with respect to each other.

The chrominance signal components are applied through a suitable band-pass filter to the $B - Y$ and $R - Y$ demodulators. In the demodulators the chrominance information is arranged in proper order for release and is demodulated under control of an inserted sub-carrier sine wave. This sine wave is generated at the receiver under control of the sub-carrier sine wave burst which rides on the backporch of the received horizontal sync pulses. Two 90-degree related sine waves are generated and demodulate the $R - Y$ and $B - Y$ chrominance signals. The output of each demodulator is applied through a low-pass filter to a signal-adding stage, Fig. 14. The outputs of the two filters are also applied to a $G - Y$ adder and phase-inverter which form the $G - Y$ signal data from the $B - Y$ and $R - Y$ components. Into each of three separate stages (one for each primary color) the luminance or Y signal is introduced. As a result, the adder output circuit contains the same three individual primary colors which were originally picked up at the color camera.

The receiver's tricolor tube contains three separate electron-guns and a multielement fluorescent screen, Fig. 17. Each primary color excites a different electron gun, causing each of the three guns to generate a stream of electrons. Each gun's electron stream represents the luminance level of its assigned color.

Hundreds of thousands of individual red, blue, and green dots of phosphor are positioned in triangular groups of three over the entire fluorescent screen. (Each group consists of one dot of each of the three colors.) The deflection and electron-beam control systems of the tricolor tube permit each gun's electron stream to strike only its assigned color (phosphor element). Thus, each of the three guns, in association with its corresponding phosphor dots, forms a complete color picture with its assigned color. The minuteness and very close positioning of the individual color dots form three complete color pictures which, because of the very fine dot pattern, appear as a single picture to the eye. The eye is not able to distinguish each tricolor group as three individual dots; rather it "sees" them coincident. Thus a three-color tristimulus picture is registered on the fluorescent screen which permits the reproduction of both primary and secondary colors.

6. *Constant Luminance*

Let us now consider in more detail some of the individual functions and characteristics of the NTSC color system. In constant luminance transmission the color signals control the chromaticity of the reproduced image but do not influence its luminance, which is controlled only by the Y signal. First we must mention that the application of the theory of constant luminance has done much to reduce flicker and the effect of noise and interference on the quality of the color picture or monochrome version of the color picture. This method was adopted because the influence of crosstalk and interference is more apparent to the eye when they affect the brightness or luminance of the color picture and is less apparent when they alter the hue of the color picture. By proper choice of relative red, blue, and green signal levels, the luminance factor can be held constant in the chromaticity channels.

In a *symmetrical chrominance color system,* there are equal levels of red, blue, and green primaries that make up the luminance signal. Likewise, the response of the chrominance channels would be symmetrical with respect to each of the primary colors. Thus the three primary color signals (spaced at 120-degree intervals) have equal susceptibility to noises and interference entering the chrominance channels. Inasmuch as the channels are equal in response and 120 degrees apart, the presence of noise creates an equal disturbance in each primary color channel, Fig. 18, drawing A. The symmetrical chrominance

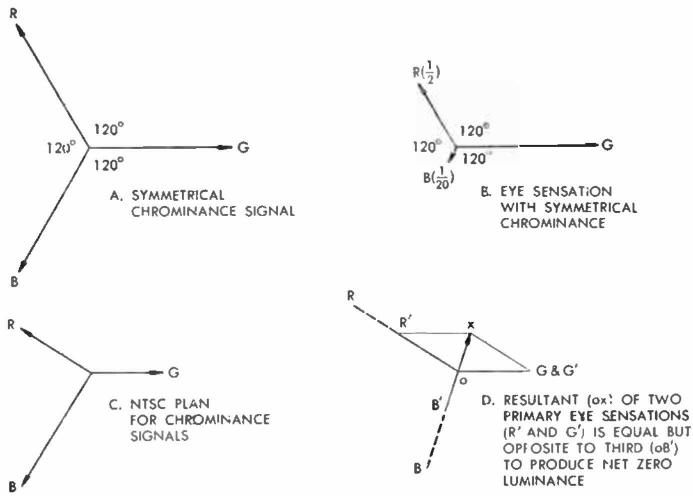


FIG. 18 Constant Luminance Plan

vector indicates that noise would affect each primary color signal equally in terms of voltage disturbance, and it would at first seem possible to have a net interference cancellation at the fluorescent screen of the color picture tube. However, the response of the human eye is not linear in terms of the brightness of the three primary colors. Rather, visual sensation is illustrated in vector B (the eye is twice as sensitive to green as to red and approximately 20 times more sensitive to green than to blue). Accordingly, in terms of net visual sensation, the symmetrical chrominance signal does create an apparent noise-pattern on the color screen because visual sensation responds differently to equal noise-voltage changes in each of the primary colors.

In the NTSC color system the type of arrangement of the relative levels of the three primary color signals produces equal visual sensation, rather than equal voltage variations at the output of the chrominance channel. Therefore, in the chrominance channel there is a higher amplification in the red channel than in the green and a still higher sensitivity in the blue than in the green. Consequently, the presence of noise in the chrominance channel creates a greater voltage change in blue than in green. However, when this differential is applied to the color tube and then viewed by human eyes, no brightness change will be observed, the change, in terms of visual perception, having been equalized. Thus differing degrees of noise variations in the chrominance channel will reproduce as equal brightness changes in terms of visual sensation, producing a net cancellation of the disturbance. Noise in the chrominance channel will not be apparent because of the constant luminance response to noise variations. Instead, noise components in the chrominance channel will cause changes in hue and saturation which are far less noticeable in terms of visual sensation, producing a perceived brightness change of zero, vector D. The apparent noise reduction with use of constant luminance is approximately 6 to 8 decibels, as compared to a symmetrical chrominance signal.

Equal brightness response by the chrominance signals is obtained not only by proper relative sensitivity to the three basic colors but also by proper choice of angles. We can control net brightness by means of the timing of signal application as well as the strength of the applied signal. Thus perception of equal brightness can be effected without making the relative voltage ratios among the various primary colors excessive.

7. *Formation of NTSC Constant Luminance Signal*

The luminance portion of the NTSC color signal contains proportions of signal from the red-, green-, and blue-sensitive cameras required to form a signal that is representative of the brightness variations of the transmitted scene. These proportions are also chosen to correspond to the spectral brightness characteristic of the eye in order that the monochrome presentation of the scene will have a true gray or neutral scale. The combination of color signals most representative of luminance as a function of visual sensation, color sensivity of the television system, and phosphor dot response is as follows:

$$E_Y \text{ or } Y = 0.30E_R + 0.59E_G + 0.11E_B$$

The color signal is developed at the camera, either by a single-color camera tube especially designed for color pickup or with the use of three, separate, image-orthicon camera tubes of the monochrome type and special color filters and mirrors, Fig. 19. A single-lens system supplies the color information to two special dichroic mirrors. These dichroic mirrors allow the passage of green light but reflect red and blue. The reflected blue information passes through a suitable filter arrangement to a conventional mirror that reflects the blue information into the so-called blue camera tube (a conventional monochrome camera tube that is responding to the luminance information of the blue light). The red information is reflected from a red dichroic mirror to a reflecting mirror and thence into the red camera tube. The green information passes through the red and the blue dichroic mirrors directly to the green camera tube.

The signal voltage output of each camera is linear with respect to the luminance information presented to its photosensitive circuit. However, the television system and, in particular, the color tube have a nonlinear characteristic in terms of brightness information. Thus it is necessary to alter or predistort the color signal at the output of the camera tube to permit an over-all linear response for the color television system. The picture tube has an approximate square-law response to brightness, and consequently an inverse power law is used in the amplifiers that follow the camera tube to obtain an over-all linear brightness response. This is referred to as "gamma correction" and occurs before the signal is applied to the matrix unit for formation into a constant luminance construction.

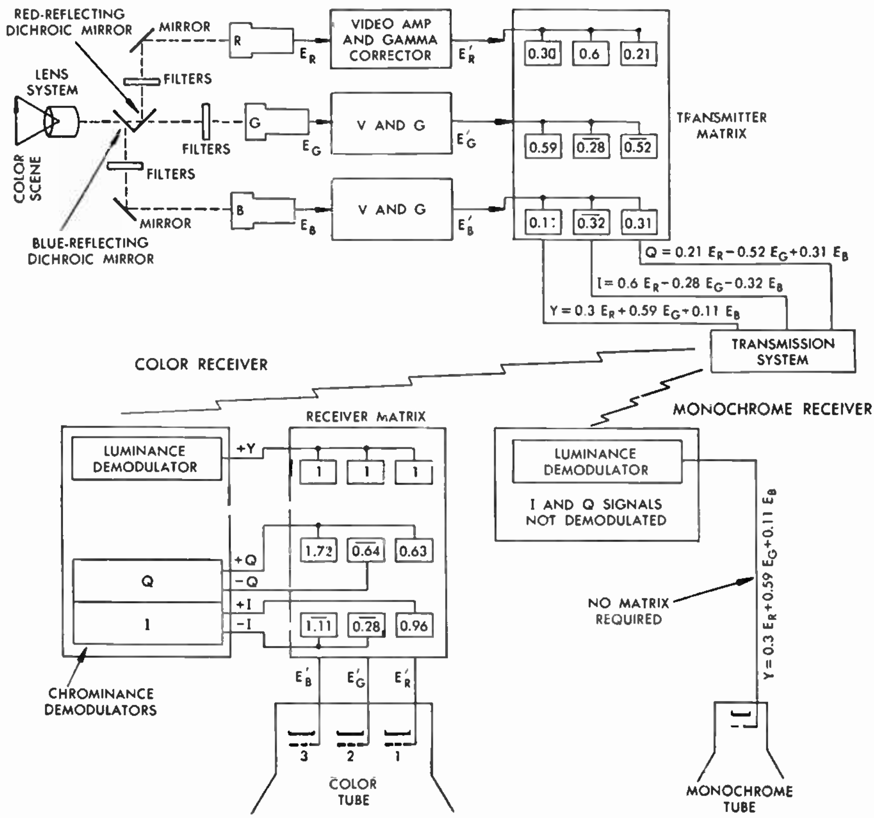


FIG. 19 Formation of NTSC Signals Showing Matrix Constants

One of the functions of the matrix unit is to combine the three color signals in the ratios of 59 per cent green, 30 per cent red, and 11 per cent blue, thus producing the constant luminance signal indicated by the previous formula. Inasmuch as this signal represents luminance and not chrominance information, it is actually a monochrome signal and can be applied without any change directly to the grid of a monochrome picture tube. This is the compatibility feature that permits a color picture to be reproduced in monochrome on the screen of a conventional black-and-white picture tube. Furthermore, the signal percentages contributed by the primary colors develop proper signal voltage variations at the grid of the monochrome picture tube. These variations produce brightness differences on the screen which correspond to the special brightness characteristic of the eye, reproducing a true, gray-scale image of the original color scene.

The chrominance information is contained in the two color-difference signals or so-called "*I*" and "*Q*" signals. The relative levels of the primary color signals in the *I* and *Q* chrominance components are determined by the necessity of retaining a constant luminance signal and the proper location of the *I* and *Q* color axis in the color gamut of the system. The *I* and *Q* voltages can be represented as color difference signals or in terms of their relationship with respect to the primary color voltages from the camera as follows:

$$\begin{aligned} E_Q \text{ or } Q &= 0.41(E_B - E_Y) + 0.48(E_R - E_Y) \\ &= 0.21E_R - 0.52E_G + 0.31E_B \end{aligned}$$

$$\begin{aligned} E_I \text{ or } I &= -0.27(E_B - E_Y) + 0.74(E_R - E_Y) \\ I &= 0.60E_R - 0.28E_G - 0.32E_B \end{aligned}$$

If we position the expressions for luminance and two chrominance signals vertically, they become three simultaneous equations—so related that when a white or neutral shade is to be transmitted (luminance information only, or $E_Y = 1$) the two chrominance components reduce to zero:

$$\begin{aligned} Y &= 0.30E_R + 0.59E_G + 0.11E_B \\ Q &= 0.21E_R - 0.52E_G + 0.31E_B \\ I &= 0.6E_R - 0.28E_G - 0.32E_B \end{aligned}$$

Thus, if the camera output is producing equal voltages of each color, as it would in the transmission of white, the substitution of one (1) in each equation would produce a luminance summation of one, while the summation of the two chrominance voltages would equal zero:

$$\begin{aligned} Y &= 0.30E_R + 0.59E_G + 0.11E_B \\ Y &= 0.30 + 0.59 + 0.11 = 1 \end{aligned}$$

$$\begin{aligned} Q &= 0.21E_R - 0.52E_G + 0.31E_B \\ Q &= 0.21 - 0.52 + 0.31 = 0 \end{aligned}$$

$$\begin{aligned} I &= 0.6E_R - 0.28E_G - 0.32E_B \\ I &= 0.6 - 0.28 - 0.32 = 0 \end{aligned}$$

when $$E_R = E_G = E_B = 1$$

If we are to transmit a yellow, the green and red voltages would be represented by one; the blue voltage would be zero because equal levels of red and green are required to transmit a yellow. The luminance formula would have a sum of 0.89 while the *I* and *Q* signals would be 0.32 and −0.31 respectively:

$$Y = 0.30E_R + 0.59E_G + 0.11E_B$$
$$Y = 0.30 + 0.59 + 0 = 0.89$$

$$Q = 0.21E_R - 0.52E_G + 0.31E_B$$
$$Q = 0.21 - 0.52 + 0 = -0.31$$

$$I = 0.6E_R - 0.28E_G - 0.32E_B$$
$$I = 0.6 - 0.28 - 0 = 0.32$$

when $E_R = E_G = 1$

and $E_B = 0$

The summation of various other colors in terms of the three basic signals as well as of the color difference component is tabulated in chart form below:

Color	E_R	E_G	E_B	Y	I	Q	$E_R - E_Y$	$E_G - E_Y$	$E_B - E_Y$	E_C	Angle
Neutral	1	1	1	1.00	0	0	0	0	0	0	0°
Yellow	1	1	0	0.89	0.32	−0.31	0.11	−0.11	−0.89	0.44	13°
Red	1	0	0	0.30	0.60	0.21	0.70	−0.30	−0.30	0.63	77°
Magenta	1	0	1	0.41	0.28	0.52	0.59	−0.41	0.59	0.59	119°
Blue	0	0	1	0.11	−0.32	0.31	−0.11	−0.11	0.89	0.44	193°
Cyan	0	1	1	0.70	−0.60	−0.21	−0.70	0.30	0.30	0.63	257°
Green	0	1	0	0.59	0.32	−0.31	−0.59	0.41	−0.59	0.59	299°

It is a function of the transmitting matrix to form three color signals from the voltage constants indicated in the chart.

The values of the chrominance signal constants also locate the two color axes they represent on the color gamut, Fig. 20. In this illustration the color gamut is again represented by a triangle with the wide-band, or Q, axis and the narrow-band, or I, axis both passing through illuminant (W) at respective angles that are a function of the color difference each signal conveys. For example, the wide-band axis extends from orange on the right side of the triangle to cyan (blue-green) on the left side. Points along this line thus represent color differences between orange and cyan. The eye is better able to distinguish the color differences along this axis in terms of color detail. Consequently, it is advisable to convey this color range over a wider band of frequencies, and therefore, the I channel has been assigned a bandwidth of 1.5 megacycles. The second chrominance signal, referred to as the narrow-band axis, extends from purple to green; along this axis the eye has very little resolving power in terms of color detail. Consequently, for the narrow-band or Q chrominance signal the bandwidth is only ½ megacycle.

The transmission system must convey three color signals to the receiver to permit presentation of a color picture (Fig. 19). The luminance signal carries the brightness detail, while the two chrominance signals convey the hue and saturation to the receiver. Inasmuch as the constant luminance principle is used, a matrix unit is also required at the receiver when a color picture is to be presented. If only monochrome pictures were to be observed on the color television screen, the luminance signal itself would provide all the information required. However, in the presentation of the color picture the luminance and chrominance signals combine—after the chrominance signals have been passed through the proper matrixing unit, whose function is to restore the color signal to the same respective blue, green, and red primary voltage levels that existed at the output of the gamma corrector at the camera. The output of the matrix unit is in the form of color difference signals that, when combined with the luminance signal, produce the original primary-color signal voltage levels.

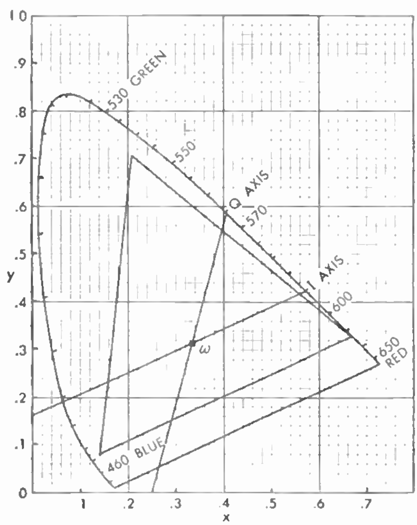


FIG. 20 Axes of I and Q Signals

When a gray or neutral tone is to be conveyed over the color system, the equal primary voltage levels at the camera produce the constant luminance Y signal as well as the chrominance I and Q components. However, because of the standards chosen in the formulation of the chrominance signals, the outputs of the chrominance channels reduce to zero for neutral colors, and only a monochrome presentation appears on the color-tube screen (equal excitation of the phosphor dots). Likewise, if a monochrome presentation of a color signal is to be made on a color picture-tube screen, it can be produced by turning off the chrominance signals and allowing the luminance signal to provide equal brightness excitation of the phosphor dots.

8. *Matrix Units*

It is the function of the matrix unit at the transmitter to proportion the luminance and chrominance signals. Proper proportioning is required in obtaining constant luminance transmission as well as in attaining faithful color reproduction at the receiver. Thus a matrix unit is necessary at the transmitter in order to form the proper relationship between primaries in luminance and chrominance channels; a second matrix unit is required at the receiver for the recombining of luminance and chrominance signals at the color picture tube. This recombining yields true color rendition.

The activity that takes place in the receiver can best be explained by assuming initially a symmetrical chrominance signal. If it is desired to transmit a group of saturated color bars positioned from left to right across the screen, Fig. 21, a symmetrical chrominance system, which would respond equally to all three primary colors, could be used. For example, saturated green, red, and blue would each have an effective amplitude level of 0.33, and a summation of the primaries of equal intensity would equal *one* and represent white as indicated in waveform *B*. A saturated yellow would be at the 0.66 level because it would be the summation of saturated red and green of equal intensity. Thus the luminance signal, as indicated in waveform *B*, represents one line of information (as indicated between horizontal sync pulses), and proper timing of the color sequence would be controlled by the sine wave's color burst on the back-porch of each horizontal sync pulse. The actual color information, as mentioned previously, is added as modulation components on the sub-carrier sine wave.

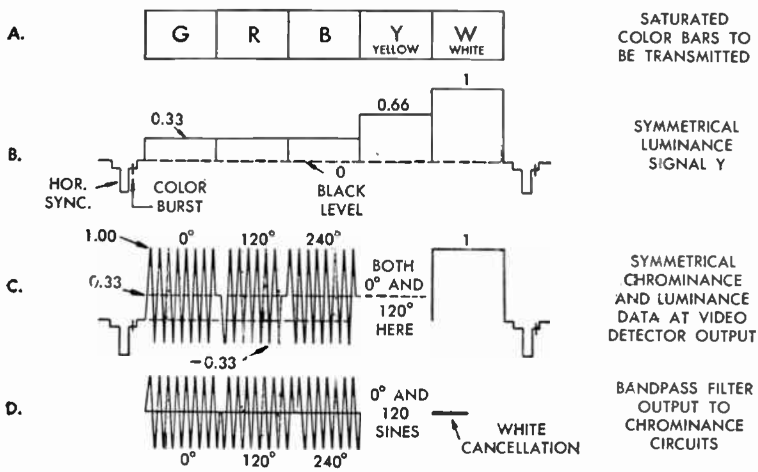


FIG. 21 Luminance and Chrominance Signal Formations (Symmetrical)

At the video detector of the television receiver the combined symmetrical chrominance and luminance information is as illustrated in waveform *C*. Sine-wave levels for each of the three basic colors are equal. These colors can be interpreted separately because of their 0-, 120-, and 240-degree relations. The yellow signal contains both 0- and 120-degree sub-carrier signals. No chrominance signal represents white (white-to-black gradations represent luminance information only). At the output of the bandpass filter leading to the chrominance demodulators the signal is as indicated in waveform *D*. The signal has equal amplitude levels for the three primary colors, and two individual chrominance signals of equal amplitude level represent the yellow information. Three primaries of equal amplitude produce cancellation during the white interval when no chrominance information is present.

The outputs of the color demodulators are the three top waveforms of Fig. 22 and are referred to as the $B - Y$, $R - Y$, and $G - Y$ chrominance signals. For the $B - Y$ signal, or blue primary, the green and red levels are below the central axis, and the blue is the positive signal output with a level of

0.66. When the chrominance signal, $B - Y$, is added to the luminance signal, Y (waveform B of Fig. 21), the resultant is waveform E, Fig. 22. Notice that blue has reached the saturated level of one and that the white bar is also at a saturated level of one because its component is present in the luminance signal. Green, red, and yellow information cancel in terms of the $B - Y$ chrominance contribution. However, the $R - Y$ chrominance signal matches with the luminance signal, producing red at the saturated level of one and, in a similar manner, the $G - Y$ component, when added to the luminance signal, produces the green bar at a saturated level of one.

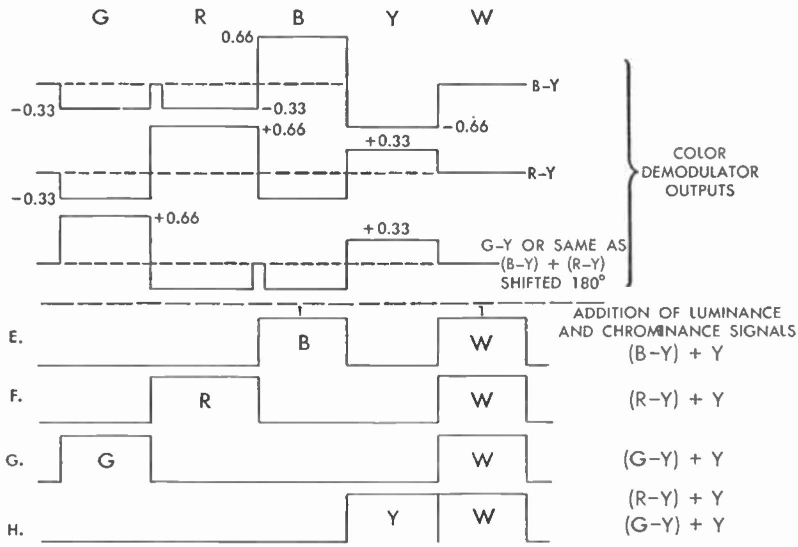


FIG. 22 Addition of Luminance and Chrominance Signals

Notice that when the $B - Y$ component is added to the $R - Y$ chrominance signal and the resultant signal inverted, the $G - Y$ signal is formed. Thus in the color system it is not necessary to have a $G - Y$ demodulator and channel because we can form exactly the same signal in the $G - Y$ adder and inverter, Fig. 14.

The formation of the yellow bar is interesting because it results from the combination of two primaries. At one and the same instant the $R - Y$ signal, added to the luminance signal, produces a saturated red signal; simultaneously, the $G - Y$ signal adds with the luminance signal, producing a saturated green. The overlapping red and green colors reproduce as a yellow visual sensation.

In a *constant luminance system* as opposed to a symmetrical chrominance system, the same net results are obtained, although the addition cannot be expressed simply in terms of waveform addition and subtraction because of the complex influence of the matrix unit. In the matrix unit at the transmitter the three primary colors are apportioned in a manner to produce a constant luminance signal. For example, signal Y consists of $0.3E_R + 0.11E_B + 0.59E_G$, or a total of unity, and represents white (luminance information). Eventually at the tricolor tube the three colors are to be reproduced in true proportion (equal levels of red, blue, and green producing white).

The transmitting matrix also forms the *I* and *Q* chrominance components for the standardized proportions indicated. At the receiver the luminance signal is detected and applied, as is, to the matrix unit that drives the color picture tube. The two chrominance components are applied through the band-pass filter to their separate demodulators and low-pass filter output circuits. The outputs of the low-pass filters are also supplied to the receiving matrix. Secondary outputs are taken from the low-pass filters and applied to the $G - Y$ adder and signal-forming circuit; output of the circuit, in turn, is applied to the green section of the matrix unit. Gains of the receiving matrix are proportioned in such a way as to produce the constant luminance resultant as well as to re-establish proper primary color levels (trichromatic coefficients required to reproduce the chromaticity information with proper hue and saturation). The luminance signal, of course, enters the matrix unit at the same levels developed at the transmitting matrix output. This luminance signal, Fig. 23, must add to the three chrominance signals in order to reproduce proper color levels. For example, in the transmission of saturated green, red, and blue bars, the $B - Y$ chrominance signal, when added to the luminance signal, results in a blue of one, or saturated blue, green and red being cancelled. Likewise, the $R - Y$ signal, when added to the luminance signal, produces a saturated red with a net cancellation of green and blue signal components.

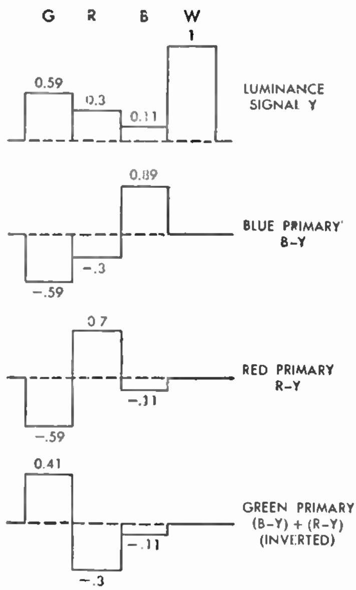


FIG. 23 Signals for Saturated Color Bars—NTSC Constant Luminance

Although the formation of the $G - Y$ signal cannot be illustrated in a simple waveform drawing, because of the complex gain function of the matrix unit, the net $G - Y$ signal would be a saturated green and cancellation of red and blue components.

The simple mathematics of the matrices demonstrates most effectively the action of the transmitter matrix in the formation of the constant luminance and the two chrominance signals and of the receiver matrix in preparing the signals for presentation to the color tube. At the transmitter, Fig. 19, the matrix gains are arranged to form the luminance and two chrominance signals from the gamma-corrected primary-color voltages. Thus the figures in the matrix block represent the gain of a particular matrix section with relation to the red, blue, and green signal voltages. For example, in the formation of the luminance, or Y, signal the green-signal output would be higher than the red, and the red-signal voltage higher than the blue despite the application of equal red, blue, and green signal voltages to the matrix. The green section would have a gain of 0.59, while the red and blue sections would have gains of 0.3 and 0.11, respectively. The luminance signal does not contain equal amplitude signals of each of the three primary-color voltages, although equal amplitude color voltages are previously applied to the input of the matrix. This matrix section forms the constant luminance signal that conveys the brightness information of the color scene to the tricolor tube.

In the same manner the proper choice of matrix gain permits formation of the two chrominance signals, which also have unequal color signal components despite the application of equal color signal voltages to the inputs of the individual matrix sections. The matrix gains are chosen to produce an output voltage of zero in each of the two chrominance channels whenever the matrix input voltages are equal. The matrix gains in the chrominance channel must also have the proper constants to permit re-establishment of the basic color signal voltages at the receiver when the two chrominance signals are once again combined with the luminance signal. The minus signs on some of the matrix gain constants represent a phase-shift of 180 degrees or a polarity reversal.

In the reproduction of a monochrome version of the color picture or the reproduction of a neutral tone through the color system it is necessary that the luminance signal only be applied to the color tube. The constant luminance Y signal ensures proper brightness-excitation of the color dots on the screen. Likewise, the constants of the luminance signal also ensure a true gray-scale rendition of the color picture on the screen of a conventional monochrome receiver, because the relative brightness variations correspond to the manner in which visual sensation responds to the brightness of those colors.

To reproduce a color picture, however, it is first necessary to establish the three primary colors at the same relative levels at which they existed at the output of the gamma-correction system. Consequently, a second matrix unit is necessary at the receiver. The gains of the receiving matrix are as indicated in Fig. 19.

The gains of the matrix unit to the luminance signal are "unity" in all three of the matrix sections, as is necessary to reproduce the three primary color signals representing red, green, and blue. In the formation of the blue-signal voltage, the luminance and the I and Q signals must be combined. The luminance signal is introduced to a gain of unity, the Q signal a matrix gain of 1.72, and the I signal a matrix gain of 1.11. The minus sign in front of the I matrix refers to a phase-shift of 180 degrees. The action of the matrix can be mathematically demonstrated by the completion of the equation for one of the color channels. In the example, when the luminance and two chrominance signals (I and Q) are applied to the blue section of the matrix unit and the equation is solved, the voltage remaining represents the blue voltage only. The green and red components of the I and Q signals have been cancelled out.

$$E_3 = +Y + 1.72Q - 1.11I$$
$$E_3 = 0.3E_R + 0.59E_G + 0.11E_B + 1.72(0.21E_R - 0.52E_G + 0.31E_B) - 1.11(0.2E_R - 0.28E_G - 0.32E_B)$$
$$E_3 = 0.3E_R + 0.59E_G + 0.11E_B + 0.36E_R - 0.9E_G + 0.53E_B - 0.66E_R + 0.31E_G + 0.36E_B$$
$$E_3 = E_B$$

Likewise, in the red or green channels the final signal voltage remaining represents either red or green respectively, while the other two colors have cancelled out.

$$E_2 = +Y - 0.64Q - 0.28I$$
$$E_2 = E_G$$
$$E_1 = +Y + 0.63Q + 0.96I$$
$$E_1 = E_R$$

A few basic matrix systems are illustrated in Fig. 24. In the first system the plates or cathodes of three tubes are common, and each tube receives excitation by voltage of one of the basic color signals. For example, if we were

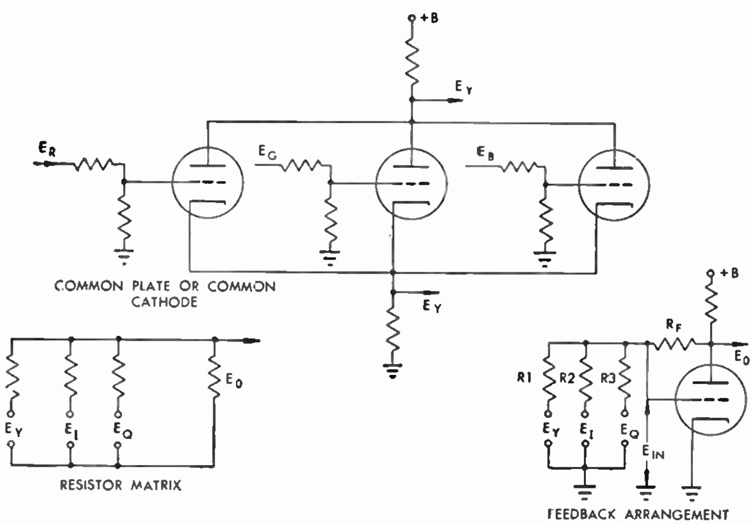


FIG. 24 Typical Matrix Circuits

to form the luminance signal at the transmitter, the grid would be excited separately by the red, blue, and green primary voltages. A combined luminance signal would be available at either the plate or cathode circuit. The proper division of the primary signal voltages could be accomplished by a resistive voltage divider in the grid circuit of each tube. If equal voltages were to excite each grid, as in the transmission of a neutral tone, the voltage dividers would supply 59 per cent of unity voltage to the grid of the green tube, 30 per cent of unity voltage to the grid of the red tube, and 11 per cent of unity voltage to the grid of the blue tube. Thus the constant luminance signal (according to NTSC constants) would be present at either the plate or cathode circuit of the matrix section.

When the plate circuit output is employed, it is possible to obtain some gain and a higher output voltage while the cathode or cathode-follower output system permits a low output impedance but no gain.

It is also possible to use a simple resistor network, as indicated in the second drawing of Fig. 24, if the three applied voltages are derived from a low-impedance source. The values of the series resistors are chosen to provide the proper proportion of applied voltages across the summation resistor.

A vacuum-tube circuit can be used in a feedback arrangement to obtain a matrix action by another method. Again the proper proportion of each primary voltage is applied to the grid of the tube as a result of choosing the relative values of the series resistors $R1$, $R2$, and $R3$ properly. In this arrangement the value of the feedback resistor R_F determines the amplification of the matrix unit. Changing of this value varies the gain of the matrix section but does not disturb the proportional distribution of the three primary voltages applied to the grid. Consequently, in a system employing three such stages, it would be possible to control the over-all amplitude of each individual signal formed by the matrix without disturbing the proportionate part of each of the primary voltages that makes up any one signal of the group of three formed by the complete matrix system.

9. *Filters in the Color System*

In a color system a most important concern is to prevent crosstalk and interference between luminance and chrominance channels and between the two chrominance components. The various bandpass and low-pass filters throughout color circuits minimize this type of interference. The planning of a color system with a minimum of interaction starts with the choice of the sub-carrier frequency. A high enough frequency must be chosen to prevent flicker and color crawl. Crawl is most pronounced when large color areas are interrupted, as is necessary with the use of low-frequency color sub-carrier. Thus the color sub-carrier frequency is located at the high end of the video spectrum and, as discussed previously, a frequency which is an odd harmonic of one half the line rate is chosen for it. Use of this frequency results in the interspersing of the chrominance information at the dead frequency points between the luminance detail in the picture.

Proper choice of necessary bandwidth is also important in minimizing crosstalk. For example, the transmission of color information at a much lower bandwidth than that for the luminance data not only conserves bandwidth but is helpful in separating the luminance and chrominance data. Thus the NTSC color system utilizes wide-band transmission of the luminance signal only because of the signal's responsibility to carry picture detail in the form of brightness variation. There are two chrominance components, one with a narrower bandwidth than the other. First, there is a reasonably wide wide-band signal which carries color signal information dominantly in the orange-red to blue-green hues and a second and narrow-band color-mixture signal which distinguishes colors in the green-to-purple range. In this arrangement of the chrominance signal, Fig. 25, there is less likelihood of mutual interference of the two color signals. The wide-band color mixture can be in the form of single sideband transmission, while the narrow component produces double-sideband emission. If there were two single-sideband signals the possibility for crosstalk would be greater. The use of one single-sideband and one double-sideband arrangement minimizes the interaction between the two chrominance-modulating signals and also permits more color detail in the preferred range of hues. Each of the two chrominance components is also effectively isolated by the use of quadrature sub-carrier signals. The first chrominance signal amplitude modulates one sub-carrier sine wave, while a second sub-carrier sine wave that has been shifted 90 degrees is modulated by the second chrominance signal.

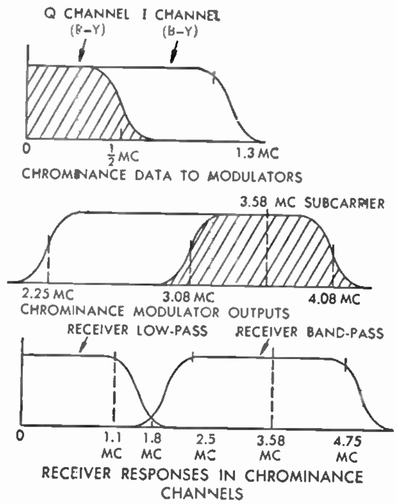


FIG. 25 Bandwidths in the Chrominance Channels

At the transmitter, Fig. 14, the chrominance outputs of the matrix unit are supplied to the balanced modulators through low-pass filters. The *I* chrominance component is carried to a low-pass filter which has a 0- to 1.5-megacycle range and the *Q* chrominance signal through a low-pass filter with a ½-megacycle upper limit. Thus only the low-frequency information reaches the balanced modulator, and the high-frequency luminance signal components are blocked from the modulator. The balanced modulators modulate the two sub-carrier sine waves with the chrominance information. The output of the wide-band chrominance modulator is then passed to a bandpass filter, which permits the transfer of the sub-carrier components in the 2- to 4.2-megacycle range. This bandpass filter arrangement has dual functions, one, preventing chrominance harmonic-modulation components below 2 megacycles from entering the luminance signal and two, attenuating higher-order harmonics of the chrominance signal. It can be understood that the luminance and chrominance information are segregated most effectively. First, the high-frequency components of luminance information are prevented from reaching the chrominance modulators by the input low-pass filter, while interaction between chrominance and luminance is minimized by the bandpass filters at the output of the balanced modulator. Likewise, interaction between the chrominance signals is minimized by the system of modulation and the formation of two sub-carrier sine waves in quadrature.

Filters are employed at the receiver in order to retain proper isolation of signals. The luminance signal is amplified at full bandwidth and is applied to the receiving matrix unit. The chrominance signals are removed from the complete video signal at the output of the video detector by application first to a bandpass filter and then to the demodulator. The bandpass filter passes the sub-carrier frequency and harmonic range, rejecting the low-frequency luminance components below 2½ megacycles. The demodulator, under control of the separate receiver-generated sub-carrier sine wave, demodulates the information present on the received chrominance signals. The demodulated *R-Y* and *B-Y* chrominance signals are passed through low-pass filters before they are applied to the output matrix. This low-pass filter, Fig. 25, prevents the feed-through of high-frequency luminance information into the chrominance channel. Thus the combination of the bandpass filter and low-pass output filters prevents the passing of both low- and high-frequency components of luminance information; consequently, a true chrominance output, representing color saturation and hue, is obtainable from the filters.

In summary, the various filters used throughout the color system prevent crosstalk between luminance and chrominance channels. However, it must be stressed that transients in either the luminance or chrominance channel can also create picture disturbances, particularly at points of rapid change from one color to another or from one brightness level to another. This means that the video response throughout the color system must be more linear and freer of transients than that of similar amplifiers in monochrome service.

10. *Sub-Carrier Chrominance-Modulation System*

In the NTSC-approved color system the chrominance information is present on a sub-carrier frequency, the phase of which determines the hue of a color while the amplitude determines the color purity or saturation. In the preceding section the influence of the matrix unit and of various filters was discussed. The matrix unit proportions the color signal voltages properly while the various filters accent the desired signals and prevent crosstalk between separate signal components. It is the function of the chrominance modulators to apply the *I* and *Q* signals to the sub-carrier frequency and of the demodulator to remove this information at the receiver, prior to the formation of the basic primary color voltages.

Thus in the study of the modulation system it is first necessary to consider the choice of the chrominance sub-carrier frequency.

1. The chrominance sub-carrier must be chosen to be an odd harmonic of one half of the line rate to obtain proper "frequency interleaving" and prevent crosstalk between chrominance and luminance channels. As discussed in an earlier section, the sub-carrier components automatically reverse in polarity between the successive scans of each line, and consequently the patterns which correspond to the interference between luminance and chrominance channels cancel out through persistence of vision.

2. A sub-carrier frequency must be chosen such as will minimize interference between the two chrominance components. The proper choice of sub-carrier frequency, use of a narrow-band axis and a wide-band axis, proper filters, and use of one vestigial sideband chrominance channel minimize the possibility of crosstalk between chrominance components. Over the frequency range of ± 0.6 megacycle, which corresponds to the symmetrical double-sideband range of the Q signal, no crosstalk can enter into the I channel, provided that suitable filters are present in the Q channel to prevent any side-band components higher than 0.6 megacycle. The I signal channel also has symmetrical sidebands over the same frequency range, and therefore, because of the quadrature-phase relationship of the two sub-carrier sine waves, there is no crosstalk into the Q channel. However, over the range from 0.6 megacycle to 2 megacycles the I channel incorporates vestigial sideband transmission, and there is a possibility of crosstalk. Again, crosstalk between I and Q channals is minimized even in this range because of the special filtering circuit in the Q channel that rejects frequencies above 0.6 megacycle which would be within the range of possible interference from the I channel.

3. The sub-carrier frequency is positioned at the high end of the video spectrum and thereby minimizes interference with the luminance components because any pattern that might exist from interference will then be as fine as possible. Interference will also be minimized by the limited attenuation in the receiver at the high end of the video passband. The sub-carrier, however, must be low enough to permit the transmission of the upper sidebands of the sub-carrier chrominance signals that convey some color detail.

4. The sub-carrier frequency must be spaced properly with respect to the sound-carrier frequency, because insufficient attenuation of the sound carrier in a receiver can sometimes cause a beat between the sound carrier and chrominance sub-carrier. Again, the principle of frequency interleaving minimizes this disturbance. It is necessary to choose a possible interference frequency that occurs at an odd multiple of one half the line rate.

Thus in the NTSC choice of carrier frequencies the separation between picture and sound carriers remains at 4.5 megacycles. However, the exact sound-carrier frequency was made to be the 286 harmonic of the line rate. Thus the horizontal frequency becomes

$$\text{line frequency} = \frac{4.5 \times 10^6 \text{ cycles}}{286} = 15{,}734.26 \text{ cycles}$$

Consequently, the field frequency, instead of being 60 cycles, becomes

$$\text{field frequency} = \frac{15{,}734.26}{262\frac{1}{2}} = 59.94 \text{ cycles}$$

The sub-carrier frequency for the chrominance information must, of course, be an odd harmonic of one half the line rate. In the NTSC standards the harmonic chosen was the 455, which produces a color sub-carrier frequency of

$$\text{sub-carrier frequency} = 455 \times \frac{15{,}734.26}{2} = 3.579545 \text{ megacycles}$$

Now if we subtract this sub-carrier frequency from the sound-carrier frequency, the resultant beat-interference frequency will be the 117 harmonic of one half the line rate:

$$4.5 \text{ megacycles} - 3.579545 \text{ megacycles} = 0.920455 \text{ megacycle}$$

or

$$\frac{920455}{\frac{1}{2} \times 15{,}734.26} = 117 \text{ harmonic}$$

Therefore, frequency interleaving will exist, and the disturbance will be less apparent.

A functional block diagram of the chrominance modulation system, Fig. 26,

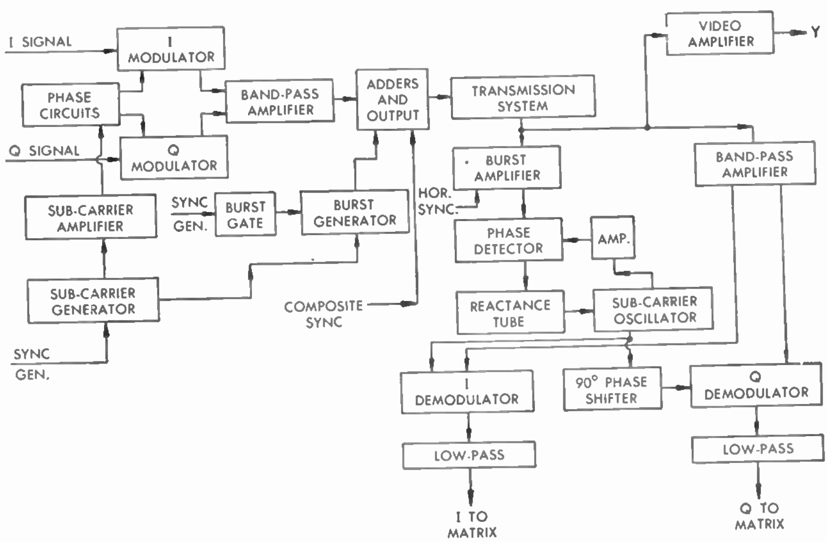


Fig. 26 Functional Block Plan of Chrominance Modulation System

demonstrates the basic circuits required in the transmission and synchronization of the chrominance channels. At the transmitter the *I* and *Q* chrominance signals from the matrix unit are applied to the balanced *I* and *Q* modulators.

It is necessary to introduce into the modulation system a sub-carrier sine wave. This sub-carrier sine wave—under control of the sync generator—forms the modulated sub-carrier frequency. It is applied through an amplifier and a phase-shifting circuit to insure a 90-degree relationship between the *I* and *Q* sub-carriers. The output of each modulator is simply the sideband components of modulation that result from the presence of the sub-carrier sine-wave frequency and modulation. The video modulation itself and the actual sub-carrier frequency are suppressed and do not exist at the modulator output. The amplitude of the output of each modulator is a function of the amplitude of the chrominance video information applied as modulation.

The two outputs of the *I* and *Q* modulators combine in a succeeding bandpass amplifier. The resultant chrominance signal has a phase determined by the relative amplitude and polarity of the two sub-carrier frequencies that form the signal. The phase angle, which can exist anywhere between zero and 360 degrees, determines the hue of the color information. The amplitude of the resultant chrominance signal is determined by the amplitudes of the two sub-carrier components that form the resultant, individual amplitudes which were determined by the original video *I* and *Q* modulation. The amplitude of this resultant chrominance signal determines the purity or saturation of the color information.

To demodulate the chrominance information accurately it is also necessary to transmit a synchronizing signal that will permit the proper timing of a sub-carrier frequency generator at the receiver. The receiver sub-carrier generator forms a sub-carrier signal which substitutes for the suppressed carrier at the transmitter. The suppression of the sub-carrier is necessary in the formation of the sideband components at the transmitter because of their eventual use of the sidebands in producing a phase-modulated resultant.

The sub-carrier sine wave, through the action of a burst gate, is added to the backporch of the horizontal sync information. The burst gate permits the addition of the sub-carrier sine wave at proper intervals which enable the sine wave to ride in correct position on the horizontal sync backporch. At the receiver the sub-carrier burst on the horizontal sync backporch is keyed off by a pulse (from the horizontal deflection circuits) and, after amplification, is applied to a phase detector. The receiver contains a sub-carrier sine-wave generator (usually a crystal-controlled oscillator) that, in conjunction with a phase detector and reactance tube, evaluates the incoming sine-wave burst in order to form a sub-carrier frequency of proper phase to permit its reinsertion as the sub-carrier of the chrominance information.

The *I* and *Q* resultant chrominance signals are applied through a bandpass amplifier to the *I* and *Q* demodulators at the receiver. To these same demodulators we now apply the locally generated sub-carrier (the sub-carrier insertion for the *Q* demodulator must be shifted 90 degrees in phase). The demodulators at the receiver are often referred to as product demodulators or synchronous detectors. The term "synchronous detection" is used because proper operation requires insertion of a locally generated sub-carrier that has previously been synchronized by information from the transmitter. The resultant plate current is the product of the chrominance signal and the inserted sub-carrier signal.

In the demodulator the combined I and Q signals are compared with the injected sub-carriers and produce difference-frequency signals in the output, the phase and amplitude of which are determined by the phase relationship between the sub-carrier sine wave and the individual I and Q sub-carrier components (the Q demodulator sub-carrier sine wave has been shifted 90 degrees with respect to the I demodulator sub-carrier). The demodulator output contains the original I and Q video information. A low-pass filter output circuit insures the suppression of the higher order of sideband components of video information as well as of signal components in the sub-carrier frequency range. The I and Q signals at the output of the low-pass filters pass into the matrixing circuit.

11. *Balanced Modulator and Demodulator Circuits*

In the chrominance sub-carrier modulation system (Fig. 27), a balanced modulator is used at the transmitter and a synchronous demodulator at the receiver. The balanced modulator is so connected that the actual sub-carrier frequency and the I and Q video information cancel in the modulator output. In the modulator the two signals are introduced and heterodyne to produce the actual applied frequencies, the sum and difference frequencies, and harmonic components. The circuit is arranged to permit the development of only the sub-carrier sideband frequencies in the output circuit. For example, the $+I$ chrominance signal is applied to one grid and a $-I$ chrominance signal to the other grid, and, inasmuch as the plates of the two tubes are in parallel, the actual chrominance signal cancels in the output circuit. Likewise, the feedback between the third grids and the plates permits some feed-through of the sub-carrier sine wave into the output circuit, but the signal is so phased that cancellation again exists in the plate output circuit. Only the sideband components, therefore, are present in the plate circuit.

The harmonic components are eliminated by proper choice of the output resonant circuit and by the transfer of the signal through a low-pass amplifier which attenuates any harmonics of the sub-carrier frequency.

A similar arrangement is used to form the Q chrominance signal with the exception that the sub-carrier sine wave applied is phased 90 degrees with respect to the sub-carrier applied to the I modulator. Inasmuch as the carriers have been suppressed, only the 90-degree related side-band components exist at the input of the low-pass amplifier. However, both signals combine in the low-pass amplifier to form a resultant sub-carrier sine wave—the amplitude of this resultant is a function of the amplitude of the individual I and Q signals, while the phase of the resultant signal, which determines the hue of the color information, is a function of the relative amplitude between the I and Q signals.

The actual phase relationship between the *I* and *Q* signals is demonstrated in the vector diagram (Fig. 27), in which the sub-carrier sine-wave phase that forms the burst placed on the horizontal blanking pulse acts as a reference. The *Q* chrominance signal is related 180 degrees plus 33 degrees with respect to the reference burst. The *I* chrominance signal is in quadrature with the *Q* chrominance signal. A resultant plotted between the *I* and *Q* signals, which is

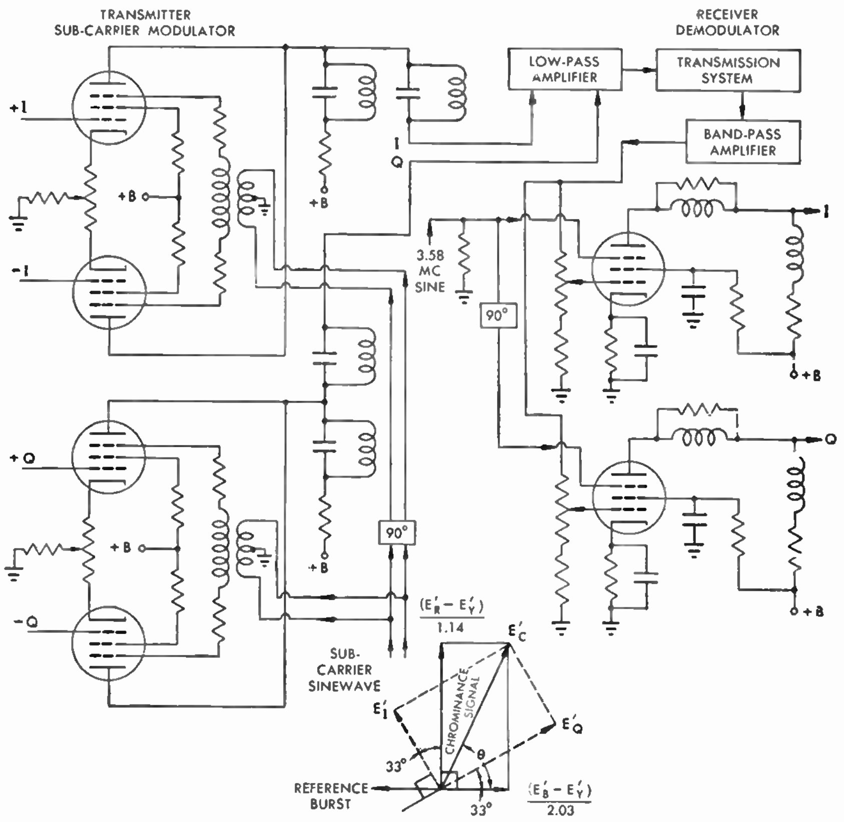


FIG. 27 Sub-Carrier Modulation System

indicated by the vector E_C of the NTSC diagram, indicates the sideband resultant that is present in the low-pass amplifier and that is used as the chrominance-modulating signal of the composite color television signal.

An explanation of the choice of vector angles will help you better understand the functioning of the chrominance portion of the color television system. First, recall the plan of a color difference signal where the *Y*, or luminance, signal contains only the brightness information of all three of the primary colors in a form of $K_1E_R + K_2E_G + K_3E_B$. Also recognize that with the luminance information removed from the three primary colors the chrominance signal remains in a form of $R - Y$, $B - Y$, and $G - Y$. Remember, too, that

$G - Y$ can be formed by the summation and inversion of $(R - Y) + (B - Y)$ and, consequently, the chrominance information can be conveved by only two color-difference signals.

Just for explanation, let us assume that the color gamut can be represented by the equilateral triangle of Fig. 28. The red and blue color-difference signals would be respectively represented by one line drawn from the "blue" corner through the illuminant or "white" point to the center of the opposite side and a second line drawn from "red" through W to the center of the opposite side. The $R - Y$ axis displays hues from red through white to red's complement cyan; the $B - Y$ axis displays blue through white to blue's complement yellow. Along the $B - Y$ axis the $R - Y$ equation reduces to zero. This does not mean that red is absent from a yellow hue (yellow represented by equal quantities of red and green), but rather that the red and green which form the yellow are a part of the $B - Y$ equation and not a part of the $R - Y$ equation which goes to zero. Likewise, along the $R - Y$ axis, the $B - Y$ equation goes to zero. Both equations go to zero at point W which represents white light (chrominance equations go to zero, and only luminance data remain).

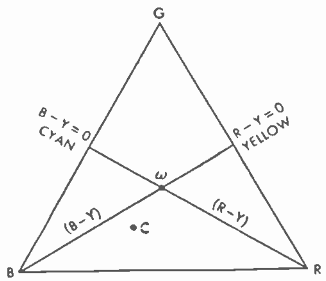


FIG. 28 Position of Axes on Color Triangle

Any other hue within the triangle can be represented by combinations of $R - Y$ and $B - Y$ components. For example, point C would be a resultant of the $B - Y$ axis moved down to C, using corner B as a pivot and axis $R - Y$ moved down to the same point, using corner R as a pivot. Thus any hue in the triangle (area of realizable colors) can be represented by the two color-difference signals.

The actual position of the $R - Y$ and $B - Y$ color-difference axes is illustrated in Fig. 29. Although the NTSC gamut does not form an exactly equilateral triangle and the illuminant W is not in the exact center of the triangle, the same color-difference signals can be obtained by properly choosing the constants associated with each primary color. The two color-difference axes represent the angles for the two narrow-band chrominance components. It is to be noted that the resultant chrominance signal which can be formed in the low-pass amplifier at the transmitter could also be formed from various other combinations of two signals to produce the same resultant.

As illustrated in Fig. 27, the chrominance resultant E_C could be formed by the vector addition of the two color-difference signals $R - Y$ and $B - Y$. However, two vectors such as E_I and E_Q when added vectorially will also produce the same resultant E_C. In the NTSC color system there is not only a wide-band chrominance signal but also a narrow-band, and by proper choice of angle we can favor those hues which are more easily discernible by the eyes and, in which, accordingly, more color detail can be seen. Thus the

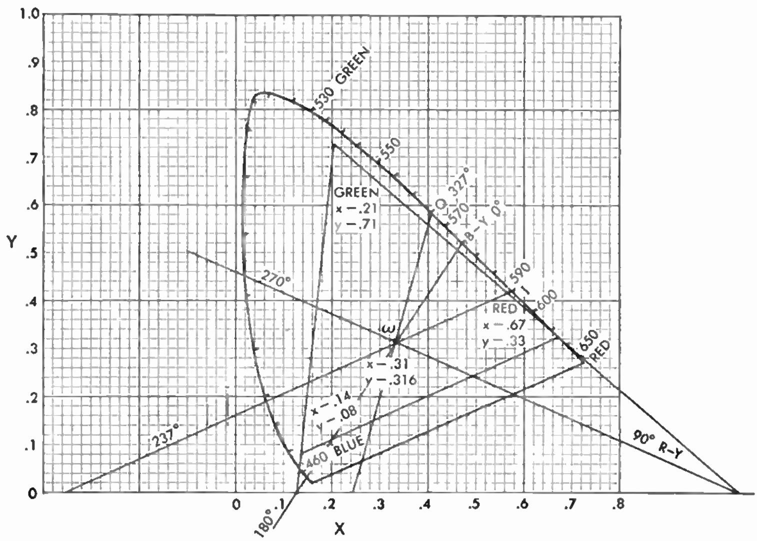


FIG. 29 Axes and Angles on NTSC Color Gamut

NTSC I and Q axes are shifted in phase with relation to the color-difference signals, as indicated in Figs. 27 and 29. Actually, the phase of the Q and I signals has been shifted 33 degrees with respect to the color-difference signals mentioned. Thus the I vector, in terms of its position on the color gamut, conveys a color range to whose detail the eye responds readily, while the Q signal, although still related at 90 degrees with respect to the I signal, sets off a range of colors to whose detail the eye responds less readily. As mentioned previously, the I component is transmitted with a greater bandwidth than that of the Q component and permits the transmission of a more detailed color picture, as compared to that resulting from the limited bandwidth obtained from use of the color-difference axes.

It is significant that with the transmission on the I and Q axes the same resultant (as above discussed) is obtained in the low-pass amplifier of the transmitter. By properly controlling the phase of the inserted sub-carrier at the receiver, the receiver is made to respond to I and Q signal components and to derive a picture with full-color detail. Likewise, it is possible to phase the inserted sub-carrier at the receiver in such a way that it will respond at the angles of the two color-difference signals. By use of a narrow-band chrominance signal, it is possible to obtain a color picture of lesser detail but with a substantial simplification of circuits in the chrominance channels of the receiver.

The phases of the various vectors indicated in Figs. 27, 29, and 30, are all in relation to the phase of the $B - Y$ color-difference axis and the phase of the sub-carrier reference burst. In most vector presentations of the color signals the positive side of the $B - Y$ axis is represented by zero degrees, and the actual phase of the color burst (synchronizing sub-carrier sine wave) is represented by wt (out-of-phase by 180 degrees). The Q axis is plus 33 degrees with respect to the $B - Y$ axis, or at an absolute 327 degrees, while the I signal is 33 degrees in advance of the $R - Y$ axis or at an angle of 237 degrees, Fig. 29. The formation of the resultant chrominance signal for some typical colors is illustrated in Fig. 30. The formation of a red resultant chrominance

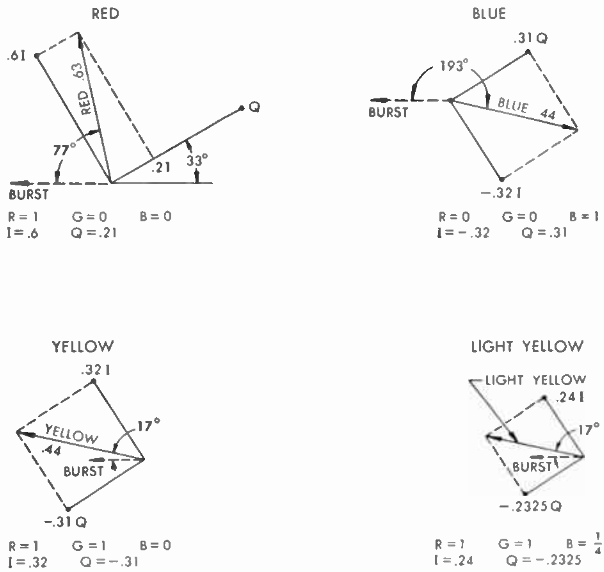


FIG. 30 Typical Hues and Angles

signal is illustrated in the first drawing. In the transmission of red a value of unity is assigned to the primary color of red, and since there is no green or blue present in the red information, both green and blue are assigned a value of zero. Now if we substitute these values in the matrix equations, we obtain the following results:

$$I = 0.6R - 0.28G - 0.32B$$
$$I = 0.6$$

$$Q = 0.21R - 0.52G + 0.31B$$
$$Q = 0.21$$

If we next draw the I and Q vectors having an I length of 0.6 and a Q length of 0.21, we can plot the resultant. The actual angles of the I and Q axes are constant. Notice that the resultant has a length of 0.63 and is lagging the burst by an angle of 77 degrees. Notice how these vectors correspond to the values presented in chart form, section 7.

In the formation of a blue the values of the primaries are: blue primary, one; red, zero; and green, zero. When we solve the chrominance matrix equations, we find an I value of -0.32 and a Q value of 0.31. Again, when we plot the I and Q vectors, remembering that the I axis is now in a different quadrant (but still at the same angle) because of its opposite polarity, we obtain a blue resultant with a length of 0.44 and an angle of 193 degrees. Comparison of these two initial vector solutions indicates clearly that the relative amplitude of the I and Q signals determines the angle of the resultant sub-carrier wave, which angle represents the hue of the color information.

In the formation of a yellow which is a resultant of equal levels of red and green, the following values are assigned: red equal to 1, green equal to 1, and blue equal to zero. The solution of this combination indicates a resultant yellow with a value of 0.44 at an angle of 17 degrees with respect to the burst.

$$I = 0.6R - 0.28G - 0.32B$$
$$I = 0.32$$

$$Q = 0.21R - 0.52G + 0.31B$$
$$Q = -0.31$$

All three examples represent saturated colors with a purity of one. Assume that we are now to transmit a lighter yellow or a yellow combined with some white; let us equate its purity to a value of 0.75, this yellow therefore not being saturated. For this example, the values of red and green can each again be assigned a value of one, while the presence of white in the resultant is indicated by a value of ¼ for the blue primary (white containing red, green, and blue). Solving the chrominance equations we obtain the following values:

$$I = 0.6R - 0.28G - 0.32B$$
$$I = 0.24$$

$$Q = 0.21R - 0.52G + 0.31B$$
$$Q = -0.2325$$

Notice that the vector resultant has the same angle with relation to the burst as the saturated yellow did. However, the length of the vector, or amplitude, is reduced, as indicated, by a shorter vector length for the resultant. This demonstrates that the saturation of the particular color being transmitted is determined by the amplitude of the resultant chrominance sub-carrier. Thus it has been proven that the hue of a color is determined by the phase of the chrominance sub-carrier, while the amplitude determines the saturation or purity of the color.

The resultant phase and amplitude of the chrominance sub-carrier for any color can be obtained by proper substitution in the chrominance formulas of the matrixing system.

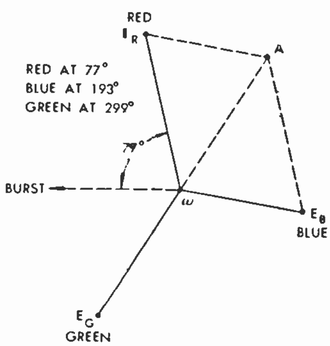


FIG. 31 Chrominance Channel Zero Resultant for Neutral Tones

Zero output from the chrominance channels during the transmission of neutral tones can also be demonstrated in vector form, Fig. 31. In the diagram the three primary colors of red, blue, and green are shown with relation to their respective amplitudes and phase angles. The addition of the red and blue vectors results in the line *AW*. Vector *AW*, of course, is equal in amplitude and opposite in phase, as compared to the green vector; consequently, the vector addition of the three primary colors reduces to zero. Therefore, there is no output from the chrominance channel during the transmission of neutral shades, the luminance signal only being active in producing the neutral tones on the color picture-tube screen.

12. *Receiver Synchronous Demodulator*

The resultant chrominance signal, after it is conveyed over the transmission system, is demodulated at the receiver by insertion of a local sub-carrier sine wave. As mentioned previously, this 3.58-megacycle sine wave is formed at the receiver, but its formation is under control of the sub-carrier burst that rides on the backporch of the horizontal sync pulse. This synchronism is necessary because the sub-carrier sine wave must be inserted at the proper angle with respect to the received chrominance signal. For example, it is necessary to phase the sub-carrier sine wave to match the phase of the original *Q* signal at the transmitter and to phase the quadrature sub-carrier sine wave so that it matches the phase of the original *I* signal. Only when local sub-carriers are inserted at the phases mentioned do we obtain true demodulation of the received chrominance sub-carrier signal applied to the receiver demodulator, Fig. 27. At the synchronous demodulator the inserted sub-carrier sine wave in the *I* channel must be in phase with the received *I* chrominance signal. Likewise, the sub-carrier sine wave is passed through a 90-degree phase-shifter so that its phase, when applied to the *Q* demodulator, is also the same as that of the received *Q* chrominance component. In other words, the original phase relation (at transmitter) between sub-carrier and the *I* and *Q* sub-carrier sidebands must be re-established.

The received chrominance signal is applied to the grids of both demodulators simultaneously. It is the relative phase of the inserted sub-carrier sine wave with respect to this resultant chrominance signal that demodulates the *I* and *Q* signals without interaction. When the resultant chrominance signal is applied to the grid of the *I* demodulator, the inserted sub-carrier wave is in phase with the *I* signal component, demodulating the *I* signal components. At this same demodulator the sub-carrier sine wave is in quadrature with the *Q* signal components and causes a resultant cancellation of the *Q* signal sideband components. With phase modulation and a suppressed carrier system, the insertion of a sub-carrier sine wave 90 degrees out-of-phase with the original sub-carrier phase results in a cancellation of the sidebands of that signal, while, if the sub-carrier is inserted in phase with the original carrier, demodulation will produce the original signal information. In the case of the Q channel the inserted sub-carrier is in phase with the Q signal components, and the demodulated Q signal appears at the output. In this stage, because of the 90-degree phase-shifter, the inserted sub-carrier is in quadrature with the I signal and results in I sideband cancellation. In summary, the I and Q signal components are demodulated and segregated at the receiver because of the fundamental relationship between carrier and sidebands in a phase-modulation system. When the inserted sine wave is in phase with the original sub-carrier, the sideband components of the resultant signal are additive and are demodulated, whereas the insertion of a sine wave in quadrature with the original sub-carrier signal results in a cancellation of the resultant sidebands.

Still another demodulation plan is possible. For example, the inserted sine wave, instead of being placed in phase with the I and Q original signal phases, can be inserted in phase with the color-difference angles, vector diagram of Fig. 27. With this phase the demodulated output of the synchronous detector would be the $B - Y$ and $R - Y$ color-difference signals, instead of the I and Q signal components. This angle operation is used in a less expensive color television receiver that does not take advantage, for economy reasons, of the wider band of color detail possible with the I and Q demodulation angles. It is significant that the station is transmitting a signal with I and Q reception capability, but by proper choice of sub-carrier sine-wave phase at the receiver it is also possible to demodulate the transmitted chrominance signal in terms of the $R - Y$ and $B - Y$ axes.

The output circuits of the demodulators are in the form of a low-pass filter in order to permit removal of any sub-carrier sine-wave components as well as high-frequency luminance information that might pass through the demodulator system. Likewise, the Q channel itself is made to have a lesser bandwidth than the I channel, thus preventing the possibility of crosstalk from the higher-frequency I channel sideband components in the Q channel. This precaution is necessary because there is only one sideband present in the I signal spectrum over its vestigial sideband frequency range. Thus there is not complete cancellation of the I signal components over that range in the Q channel demodulator.

13. *Color Synchronization System*

The functions of the color synchronization system consisting of six basic stages, are removal of the burst information from the backporch of the horizontal blanking pulse and using it to control the generation of the inserted local sub-carriers (Fig. 32). The sync and blanking information from the video amplifier of the color receiver is first applied to the input of a burst amplifier. This burst amplifier is gated by a pulse from the horizontal deflection circuit which permits the amplifier to conduct only during the short backporch interval at which time the synchronizing sub-carrier frequency is present. This sub-

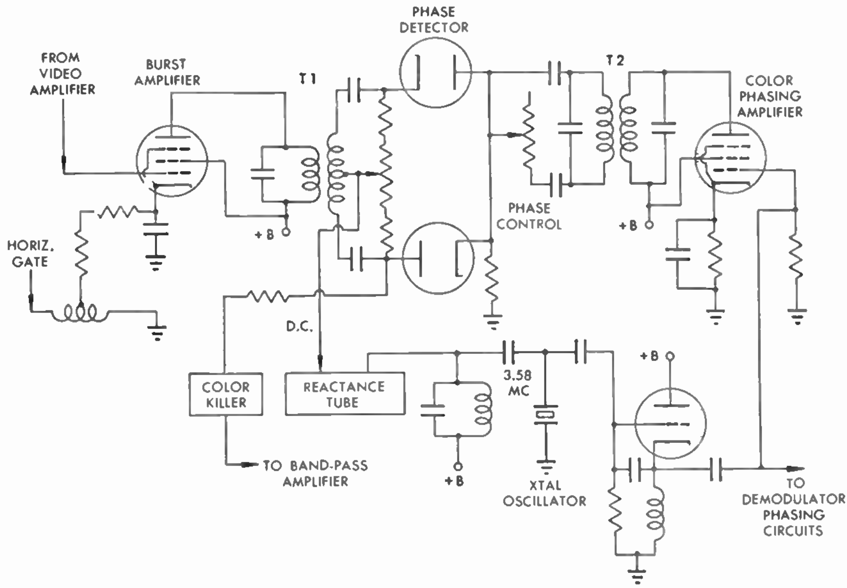


FIG. 32 Color Synchronization System

carrier burst consists of approximately 9 cycles of the original sub-carrier information at the transmitter. The sub-carrier burst is increased in amplitude by the burst amplifier and the close-coupled resonant transformer between the burst amplifier and the phase detector. The gating of the burst amplifier prevents spurious video and noise impulses from affecting the synchronization system.

The phase detector compares the frequency and phase of this burst of the sub-carrier sine-wave signal with a locally generated sub-carrier signal. Any departure in phase between the locally generated sub-carrier and the burst frequency (arriving at line rate) produces a d-c component which operates a reactance tube that is connected to the local sub-carrier generator, namely, a crystal-controlled oscillator. In effect, the phase detector, reactance tube, and crystal oscillator function as an integrator circuit, extending the time of the arriving burst and producing a continuous sub-carrier signal of the same frequency and phase as those of the arriving burst of information. This flywheel synchronization system is not subject to interference from spurious signals and is able to maintain the phase of the generated sub-carrier within five degrees with no great difficulty.

The cathode output of the crystal oscillator is applied to the demodulator-phasing circuit for formation of the quadrature 90-degree sub-carrier signals required for demodulation of either the I and Q axes or the color-difference axes. Likewise, the output of the crystal oscillator is applied to a color-phasing amplifier that supplies the sub-carrier frequency for comparison in the phase detector. In the phase detector the burst frequency is applied through T_1 to one side of the phase-comparison circuit, and the locally generated sub-carrier sine wave is applied through tuned transformer $T2$ to the other side of the phase detector. A phase control that determines the resistance-to-reactance ratio in the output circuit of transformer $T2$ can control the phase of the sub-carrier sine wave applied to the phase detector. True color hues are obtained by means of adjustment of this control when a color bar pattern or a color picture on the tricolor picture tube is observed. Once this phase relationship is established properly, any departure in phase between the locally generated sub-carrier and the incoming sub-carrier burst causes the d-c potential applied to the reactance tube to shift, either in the plus or minus direction, away from zero (as a function of the direction of the phase change), causing the crystal oscillator to follow and always to establish the zero potential that occurs when the two sub-carrier frequencies are in phase.

When a monochrome picture is being received and there is no burst present on the received horizontal blanking pulse, a special color-killer circuit in the color synchronization system biases off the bandpass amplifier through which the chrominance signals reach the color demodulator. Consequently, spurious signals are blocked from the chrominance channels during the reception of a monochrome picture, and no output issues from the chrominance channel to disturb the monochrome rendition. The color-killer circuit receives its d-c potential from one side of the phase detector (a potential that is present only when the color burst is being received) in order to bias the bandpass amplifier properly when chrominance information is being received.

Chapter 3

COLOR TUBES

14. *Shadow-Mask Tricolor Tubes*

The three outputs of the receiving matrix are applied to the respective red, blue, and green guns of the RCA tricolor tube. The tricolor tube consists of three basic sections (Fig. 33). The fluorescent screen and shadow mask permit the reproduction of the three primary-color pictures as the thousands of individual color dots are properly excited through the shadow mask. The second section of the tube consists of the three sources of electrons, one source for each primary color. This section is generally composed of three individual electron guns mounted in an equilateral triangle and spaced 120 degrees apart. After the individual beams are formed, they must converge in order to enter the holes of the shadow mask. Likewise, the entire group of three beams must be moved across and down the screen in accordance with our standard scanning pattern. Therefore, the tricolor tube must contain not only a deflection system but a convergence system as well with which to bring the three beams together and cause them to pass through the holes in the mask and strike the proper color dots.

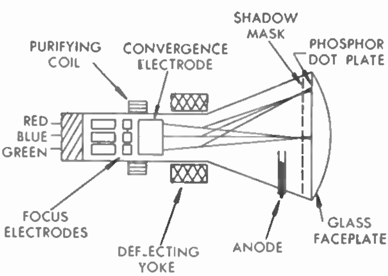


FIG. 33 Components of Color Tube

The fluorescent screen of the picture tube consists of hundreds of thousands of individual color-emitting phospor dots (Fig. 34), positioned in groups of three in tight equilateral triangles over the entire surface of the fluorescent screen. Each tiny triangle with tangent phosphor dots positioned at the corners constitutes a primary trio of red, blue, and green. In a typical color tube there

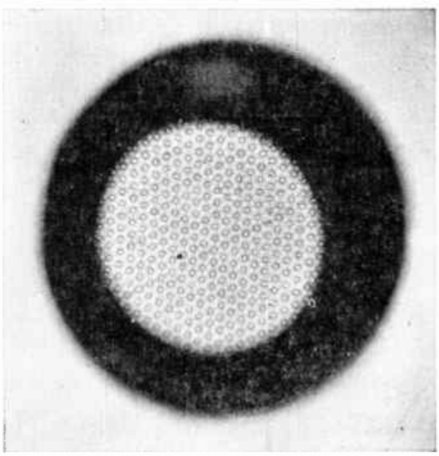


FIG. 34 Magnified Segment of Color-Dot Pattern on Screen

would be a total of 585,000 individual color-sensitive dots or 195,000 color trios. The great many color dots and their physical nearness insure that the individual primary-color images appear coincident and, insofar as visual sensation is concerned, are able to reproduce the gamut of colors conveyed by the transmission system.

In addition to the color dots there must be an aperture mask (Fig. 35), containing 195,000 tiny apertures, which permits the three separate electron

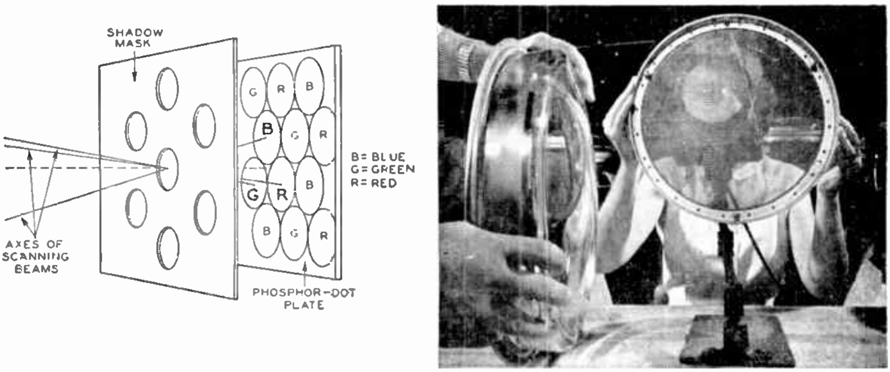


FIG. 35 Shadow Mask of Tricolor Tube

beams to pass through and strike the proper color dots. The mask prevents indiscriminate excitation of color dots in a random fashion by the three beams and permits the electron flow from each beam to strike only its respective primary color dots. The angle of convergence of the three beams at each aperture in the mask causes each of them to strike the proper color dot of each trio. As the beam moves on to the next aperture the shadow created by the mask prevents random excitation of color dots by the electron beams. Thus only at the apertures are the phosphor dots excited, and the dots are excited properly because of the converging angle of arrival of the three beams at each aperture. In a typical color tube, the aperture diameter in the shadow mask would be only $\frac{9}{1000}$ inch, with spacing between aperture centers of approximately $\frac{23}{1000}$ inch. The mask, itself, is positioned approximately $\frac{3}{8}$ inch in front of the phosphor dot plate. The three beams converge toward the center of each aperture at an angle of 1° 14′. This shadow-mask color-dot arrangement permits a horizontal resolution of between 325 and 400 lines and a vertical resolution of between 350 and 450 lines. It is interesting to note that although the spacing between apertures is greater than the aperture diameter, the arrival of the three beams and their passage through the aperture at a diverging angle (between mask and fluorescent screen) insures that approximately 90 per cent of the fluorescent screen is struck by the three scanning beams. Needless to say, production tolerances must be very exacting to allow the necessarily critical geometry of the mask aperture and phosphor trio arrangement to function properly (Fig. 36).

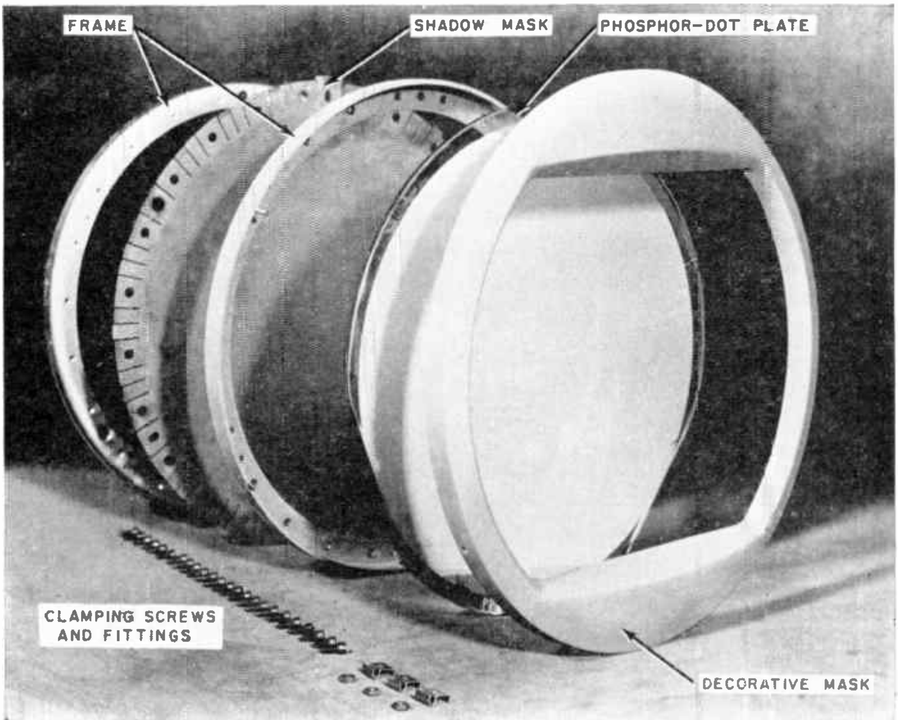


FIG. 36 RCA Tricolor Tube Viewing Screen Assembly

The electron gun of the picture tube (Fig. 37) consists of three individual guns mounted in a triangle, each gun comprising an indirectly heated cathode, a control grid, a grid No. 2 for acceleration, and grid No. 3 which functions as a focusing electrode in conjunction with grid No. 4. A beam-masking aperture, also associated with grid No. 3, removes the stray electrons around the outer periphery of the beam. Grid No. 4 is a two-section arrangement, consisting of an individual small-diameter grid at the exit of each individual gun. Each small grid connects to a large-diameter cylinder or cup.

Next in the tube there is a conductive coating on the inner surface of the neck; this acts as an anode and forms an electrostatic lens between itself and the common grid No. 4. Thus two lenses are present—one lens for bringing each individual beam to focus and a second outer lens which brings the three beams to a converging point at the plane of the aperture mask. It is possible, by means of these lenses, to control separately the beam focus and the beam convergence. The cylindrical cup of grid No. 4 is somewhat longer than its diameter and therefore minimizes interaction between the two focusing sections.

Since some variation in the convergence position of each beam can exist (because of mechanical construction tolerances or influence of stray fields), there must be a means of positioning each beam individually. This is accomplished by means of three external permanent magnets mounted on adjustable rods near the guns. An external purity coil is also employed to form a magnetic field that acts on the position of all three beams simultaneously. The coil is adjusted for best color purity when the axis of each beam coincides with the center of its respective color dots.

Proper focus and convergence are of utmost importance in obtaining a well-defined beam over the entire surface of the screen and in producing proper convergence at each aperture over the entire shadow mask. In fact, the greater distance between the flat screen and the electron gun at the outer edges of the raster, as compared to the distance at the center, can cause improper convergence and out-of-focus beams. It is possible to prevent this condition by modulating grid No. 4 with scanning-rate signals in order to retain proper beam focus over the entire screen. A mu-metal cone is provided to prevent stray fields from influencing the trio of scanning beams.

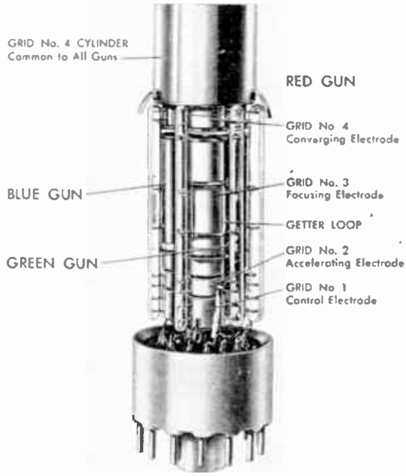


FIG. 37 Electron Gun of Tricolor Tube

The deflection system of the tricolor tube has some of the same problems as a monochrome picture tube, but disturbances appear magnified because of the larger effective size of the composite of three individual beams. In standard monochrome practice for large picture tubes, an essentially uniform, magnetic field with some occasional cosine shaping is used to obtain correct focus over the entire screen with a reasonably rectangular raster. This type of yoke was not found adequate for the tricolor tube. Instead, special segmental uniform windings form the deflection yoke in such a way as to obtain a linear and uniform magnetic field. This yoke construction, however, results in some pincushion distortion that can be corrected by other means. Despite the uniformity of the deflection field, the varying distance between the gun assembly and the fluorescent screen requires the use of special scanning-rate signals that modulate the convergence electrode grid No. 4. This modulation, which changes the focusing characteristics of the converging lens as a function of the position of the scanning beam, permits optimum convergence of the scanning beam over the entire mask. We realize the three beams travel a greater distance to reach one of the apertures at the outside of the mask than to reach a centrally located aperture. Thus it is necessary to change the difference of potential between the two electrodes that form the convergence lens and thus permit convergence at varying distances.

Likewise, the pinpoint focus of each scanning beam is a function of the distance to the screen and of the difference of potentials between grid No. 4 and the individual No. 3 grids. Again, with the application of horizontal- and vertical-rate signals, the lens is brought to focus at varying distances—in accordance with the changing distance between the gun and the screen as the scanning beam follows the standard scanning pattern. To obtain a more uniform field when the beam is deflecting to the very edges of the raster, special compensating flares or tabs (nonmagnetic but highly conductive material) are mounted near the ends of the deflection coils at the position where the coils turn up and where the neck of the picture tube progresses into the cone. This arrangement protects the wide-angle deflected beams from the distorted and large magnetic field at the end of the winding.

The CBS-Colortron tube, Fig. 38, is a similar type of tricolor tube but has a unique method of forming the phosphor screen and uses a different type of

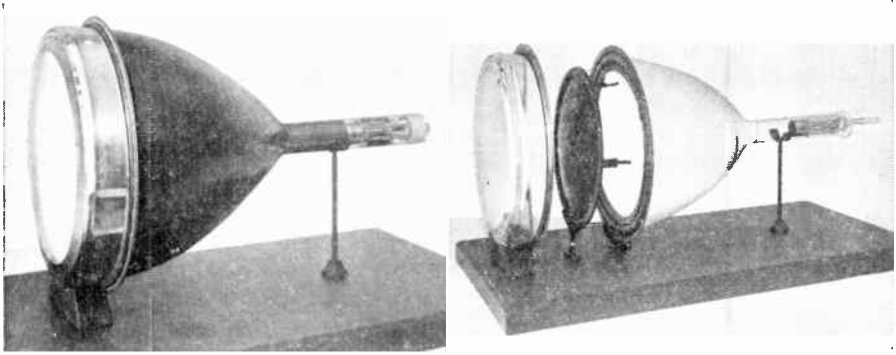


FIG. 38 CBS Colortron

aperture mask. A photographic process is used to apply the phosphor screen directly to the faceplate of the picture tube. Furthermore, the aperture mask that belongs to a particular tube acts like a photographic negative through which the screen is "deposited" on the faceplate. Consequently, the individual phosphor dots and the apertures in the mask are produced in perfect registry.

The mask itself is curved and is self-supporting with springs that hold it in place. The curvature of the mask, of course, corresponds to the curvature of the tube faceplate upon which the phosphor dots have been deposited. Consequently, as the three beams sweep across the curved surface, they travel very nearly the same distance at the edges of the picture as they do at the center. Consequently, high-amplitude convergence signals are not required, and convergence criticalness is reduced.

The relative placement of the various external components of the color tubes is illustrated in Fig. 39 along with a typical basing diagram. Typical operating conditions for a color tube are tabulated below in chart form. Notice that the maximum anode voltage required is approximately 20,000 volts and that almost 9000 volts are required by the convergence electrode. Inasmuch as such high-anode voltages are required and the regulation is critical, the high-voltage supply for a color picture tube is a rather elaborate unit; this will be discussed in connection with typical receiver circuits.

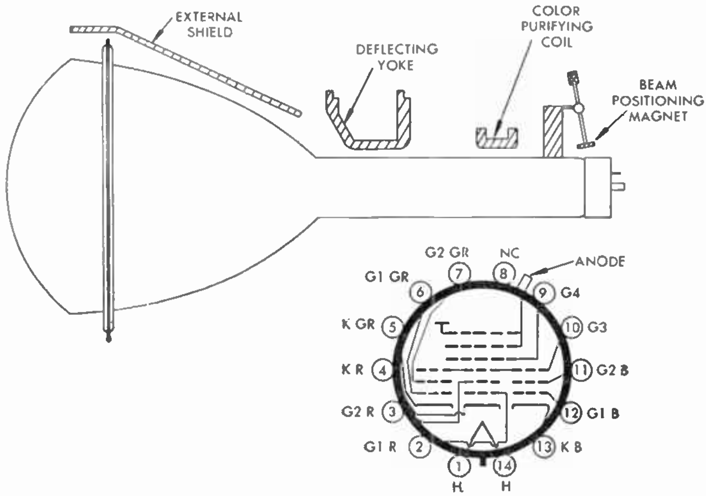


FIG. 39 Components and Basing of Colortron

COLOR-TUBE DATA CHART

GENERAL CHARACTERISTICS

Electrical Data

Heater for unipotential cathode, each gun	
Voltage	6.3 volts
Current	1.8 amperes
Focusing method	Electrostatic
Convergence method	Electrostatic
Deflection method	Magnetic
Deflection angle (approximate)	45 degrees
Electron guns, three	Red, blue, green
Direct interelectrode capacitances	
Grid No. 1 of any gun to all other electrodes except No. 1 grids of other two guns	7.5 micromicrofarads
Three cathodes externally tied together to all other electrodes	17.5 micromicrofarads
Grid No. 3 (all three No. 3 grids tied together internally) to all other electrodes	12.0 micromicrofarads
Grid No. 4 (common to all three guns) to all other electrodes	7.0 micromicrofarads
Screen	Metal-backed, tricolor, phosphor-dot type

Phosphor-dot arrangement	Approximately 250,000 triangular groups, each containing a blue, red, and green dot (a total of 750,000 dots)

Mechanical Data

Approximate weight	18½ pounds
Maximum over-all length	26⅛ inches
Maximum diameter	15¾ inches
Screen diameter	11½ inches
Bulb contact	Metal-flange seal
Base	Small-shell bidecal 14-pin
Mounting position	Any

Maximum Ratings—Design-Center Values

Anode voltage	20,000 volts d-c
Grid No. 4 (convergence) voltage	11,000 volts
Grid No. 3 (focus) voltage	5,000 volts
Grid No. 2 (accelerating) voltage, each gun	500 volts d-c
Grid No. 1 (control) voltage, each gun	
Negative-bias value	200 volts d-c
Positive-bias value	0 volts d-c
Positive-peak value	2 volts
Peak heater-cathode voltage, each gun:	
Heater negative with respect to cathode:	
During warm-up period not to exceed 15 seconds	410 volts
After warm-up	180 volts
Heater positive with respect to cathode	15 volts

TYPICAL OPERATING CONDITIONS

Anode voltage	20,000 volts
Grid No. 4 (convergence) voltage	9,300 volts
Grid No. 3 (focus) voltage	3,100 volts
With grid No. 2 voltage of 200 volts, each gun	
Grid No. 1 voltage, each gun	−45 to −100 volts
With grid No. 1 voltage of −75 volts, each gun	
Grid No. 2 voltage, each gun	240 volts
Maximum grid No. 4 current	5 microamperes
Maximum peak grid No. 3 current	400 microamperes

CIRCUIT VALUES

Maximum grid No. 1 circuit resistance, each gun	1.5 megohms
Dynamic-convergence voltage	500 volts
Dynamic voltage	150 volts

15. *Single-Gun Color Tube*

There are a number of attractive advantages available with the operation of a single-gun color tube. A single-gun tube that incorporates self-registering design permits automatic registration and eliminates the critical convergence problem. In addition, the use of a single gun eliminates the three-gun balancing problem required in the standard three-gun tube to reproduce a stable and uniform gray scale.

One such tube is the Chromatron, Fig. 40; it employs color strips instead of color dots. The information is presented to the strips in sequential order, and therefore, in a receiver employing the single-gun chromatron tube, special circuits are incorporated to change over the NTSC simultaneous color signals to a fast-switching sequential color signal. Although this adds some new circuits to the color television receiver a number of the circuits normally associated with the three-gun type can be eliminated.

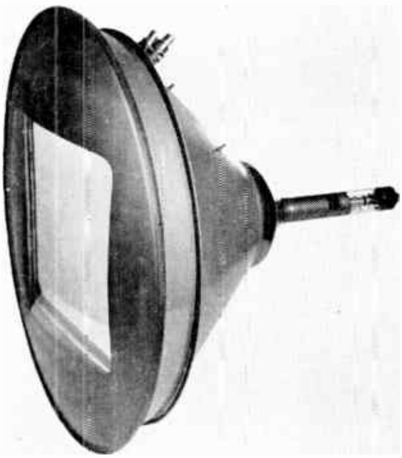


FIG. 40 Single-Gun Chromatron

The color screen of the chromatron consists of a grid of closely spaced wires on a frame which supports an image plate on which the phosphors are deposited in parallel strips, Fig. 41. The phosphor strips are deposited horizontally on the image plate by means of a screen-printing technique. After deposition the phosphors are aluminum-backed. Between the image plate and the gun of the tube and positioned near to the image plate is a wire grid whose wires run parallel to the phosphor strips. The wires are positioned in front of the red and blue phosphor strips only, with none in front of the green strips. All of the wires positioned in front of the red phosphor strips are electrically tied together and brought out to a single terminal. Likewise, the wires in the path to the blue phosphor strips are tied together and brought out to another terminal. A third terminal connects to the aluminum backing on the phosphor screen.

The difference of potential between the wire grids and the aluminum coating develops a focusing potential that causes the arriving scanning beam to converge and focus on the phosphor strips. When there is zero potential between the red and blue terminals of the wire grid (drawing *C* of Fig. 41), the electrons traveling down the length of the tube from the gun move perpendicularly to the image plate and are then sharply focused by the series of electrostatic lenses onto the green strip between each red and blue strip. When a plus

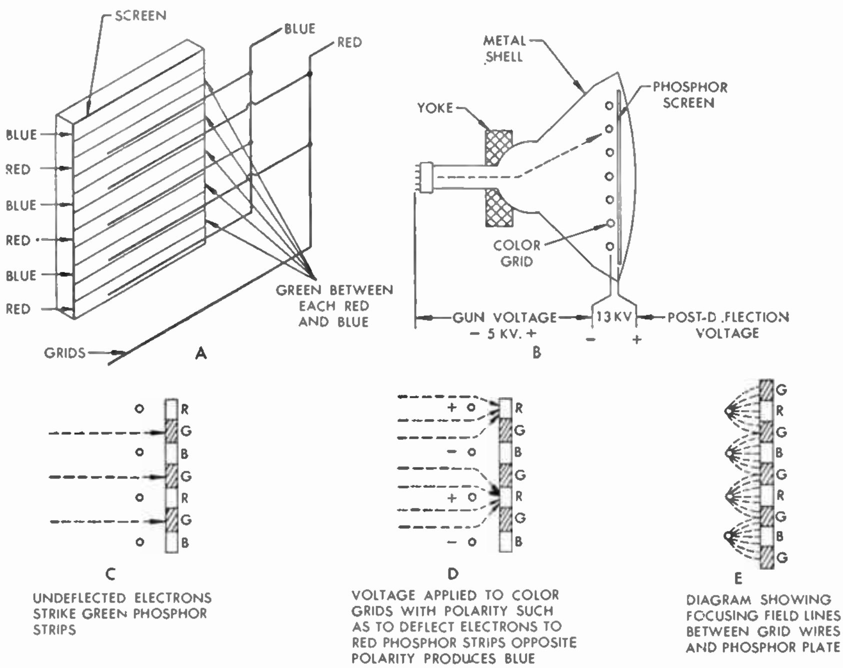


FIG. 41 Plan of Chromatron

potential is applied to the wire grids in front of the red strips and a negative potential to the wires in front of the blue strips, the electron beam is deflected and converged onto the red phosphor strips. Likewise, if the plus potential is applied to the blue grids and the negative to the red grids, the scanning beam is converged onto the blue phosphor strips. Thus it can be seen that by proper modulation of the blue and the green wire grids it is possible to direct the beam to the red, green, or blue phosphor strips. If some type of sequential-keying signal is applied to the grids, the beam can be directed to the proper color of phosphor strips in time sequence. Consequently, if the NTSC signal is properly translated to a fast-switching sequential signal, it can be made to key and present a color picture on the chromatron tube.

The actual phosphor strips are only 10 to 15 mils wide, and approximately 720 of them can be positioned from top to bottom of a raster within a picture height of slightly under 11 inches. Thus the apparent line structure of the phosphor strips is very fine and at ordinary viewing distances, in many cases, invisible. A conventional deflection yoke and assembly can be used for deflecting the chromatron picture tube.

During operation, the beam of the chromatron moves across and down the screen and traces the conventional scanning raster. However, as the beam moves along one line of information, it is also deflected vertically in sinusoidal fashion (Fig. 42), in order to allow it to scan the individual phosphor strips in proper order. The rate at which the beam is deflected in this sinusoidal manner can be at the NTSC color sub-carrier frequency. Thus the sine wave—properly synchronized and phased with respect to the color sub-carrier burst—when applied to the red and blue terminals of the chromatron grid, will cause the beam to scan (at the sub-carrier rate) in sequential order the red, green, and blue phosphor strips as the beam moves along one line of color information. The picture information is applied to the control grid circuit after it has been changed to a sequential presentation and therefore modulates the scanning beam in accordance with the brightness information for each color as the beam is being directed to the proper phosphor strip by the sine wave applied to the control grid. The rate at which the picture information is "gated on" is the third harmonic of sub-carrier frequency—three color samples per sine-wave deflection cycle to form a dot-sequential display. If the sub-carrier sine wave applied to the color grid is reversed each frame, a dot interlace display is formed, as illustrated in the second drawing. Details on the decoding of the NTSC color signal for display on a chromatron tube will be discussed in the receiver section.

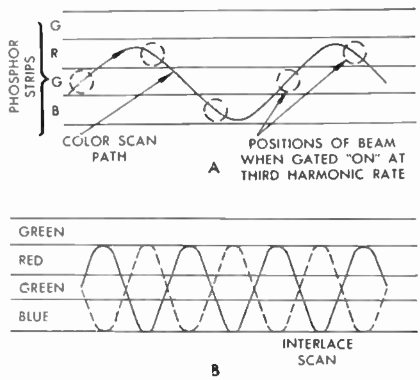


FIG. 42 Gating the Chromatron

It is also possible to use a single gun with a phosphor-dot type of picture tube. In this arrangement the scanning beam would have to be rotated in a circular manner, Fig. 43, as it moves left to right across the picture-tube screen. Instead of having three individual scanning beams that strike the correct phosphor dot in accordance with their angle of arrival at the particular masking aperture, a single scanning beam is revolved in a tight circle as it approaches the aperture. It then enters the aperture at the correct angle to strike the proper dot at a given instant. The frequency at which the beam is rotated in its tight circle can again be at the sub-carrier frequency. Thus the center of each individual phosphor dot is illuminated

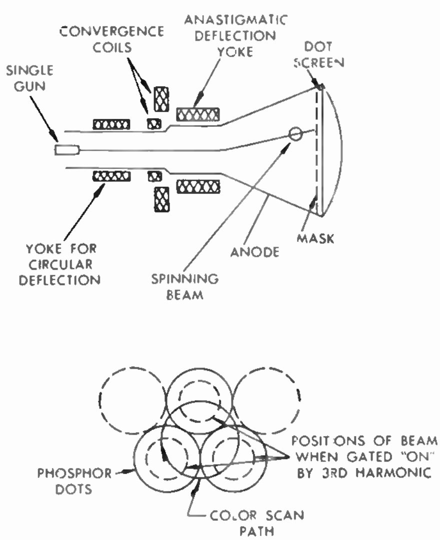


FIG. 43 Single-Gun Display for Color-Dot Screen

once during a cycle of the sub-carrier burst frequency. In this same period of time the brightness information for each individual color dot of the primary three is made to modulate the scanning beam of the single-gun color tube. Again the actual sampling or gating of colors occurs at a third harmonic rate.

Chapter 4

COLOR RECEIVER CIRCUITS

16. *Typical Color Television Receiver*

In this section a complete television receiver will be discussed in detail, from antenna to color reproducer. Those sections common to both monochrome and color receivers will be discussed with relation to additional requirements imposed by the color signal. The color sections of the receiver will be discussed at length.

ANTENNA SYSTEM

As in monochrome service, four antenna characteristics influence the reception of the color signal—gain and standing-wave ratio, field pattern, bandwidth, and stray resonant conditions. In fringe-area reception antenna gain is an important consideration if a stable and reasonably snow-free picture is to be received. The influence of noise in the picture will have the same effect as it does on monochrome reception and also two additional effects, namely, hue changes that result from noise plus the influence of noise on color synchronization. Gain will be a significant characteristic because the receiver itself will have a wide bandwidth, and the technique of reducing receiver bandwidth in favor of receiver sensitivity cannot be employed because of the adverse influence such an approach will have on chrominance fidelity and stability.

The usual broadband type of VHF antenna is satisfactory for the reception of a color signal provided that it has been installed properly and has no resonant dips on channels to be received. The bandwidth of the antenna must be broad enough to present equal sensitivity to the picture carrier and to the chrominance sub-carrier segments of the color spectrum. In the use of a narrow-band Yagi on the low-band channels a rapid drop-off in gain at the ends of the channel often is present and can influence color reception adversely.

It is possible that an antenna could have a variation of from 2 to 3 decibels in gain over a desired channel without affecting the color picture greatly. However, if the difference in gain accumulates through a number of circuits, the total can be quite detrimental. For example, if there is a loss in the antenna, additional loss because of standing wave conditions on the transmission line, or loss in tuner or in i-f strip in the chrominance spectrum, the actual color rendition deteriorates seriously. Consequently, it is advisable to keep all segments of the receiver with as uniform a bandwidth as possible for best reception of color signals.

Still another condition that arises occasionally is that of narrow dips in the response of an antenna system, as caused by spurious resonant conditions in the antenna system—from insulator mounts, cross-arm, or presence of mast. On the UHF band especially, the presence of nearby metallic objects that form resonant circuits which suck out the normal response of the antenna over a narrow band of frequencies can possibly attenuate important segments of the received spectrum.

Antenna orientation is a significant factor, too, because of the influence of multi-path signals that not only cause echoes or reflections in the picture but, at the same time, disturb the hue of the color presentation. Thus the normal disturbances which are due to faulty antenna systems in monochrome reception also have the same effect on color pictures and, in addition, the influence that such disturbances have on the chrominance information. In monochrome service, improper orientation can cause close-spaced echoes or ringing, or sync instability; in color service, the same defects are encountered and also the influence of improper orientation on the chrominance information.

Standing-wave conditions on the transmission line can also affect color reception adversely. The most serious point of mismatch, of course, is between the transmission line and receiver, where a disturbing reflection can originate. A mismatch at the antenna, provided it is not accompanied with a mismatch at the receiver as well, will result in attenuation or a lower-gain antenna system but will not set up disturbing reflections on the line. However, mismatch at the receiver accompanied by a mismatch at the antenna can set up disturbing standing-wave conditions on the line. In fact, a standing-wave condition in the antenna-transmission line system can be frequency selective and cause one end of a desired channel to pass with little attenuation into the receiver while the opposite end can be attenuated severely. This condition can reverse itself as a function of the total length of the transmission line between the antenna and the receiver input. Needless to say, if a standing-wave condition is present and the over-all length of line is such that a standing-wave minimum occurs at the receiver input on the chrominance frequency, the chrominance information could possibly be eliminated from the received signal. This condition can be checked by observing on the oscilloscope screen the horizontal blanking period at the video detector of a color television receiver. As a stub is moved along the transmission line or as standing-wave conditions are altered, the actual amplitude of the sub-carrier burst riding on the backporch can be seen to vary in amplitude.

Again the possibility of variable standing-wave conditions or suck-outs in the antenna system is of greater concern on the UHF band at the moment because of the less-consistent, uniform impedance-matching and the influence of nearby metallic objects (as well as stand-off insulators and lightning arrestors) on peak efficiency UHF reception.

It is a fact that for peak color reception there is a trend toward an antenna system with better front-to-back ratio and better response pattern accompanied with more uniformity in bandwidth which will make the antenna-installation problem less difficult and minimize the influence of reflections and interference pickup on the color picture.

TUNER AND I-F RESPONSE OF COLOR RECEIVER

The characteristics of tuners that influence color reception are sensitivity, stability, and response, just as in monochrome service. Again the tolerances will be stricter because of the presence of the chrominance carrier at the high end of the tuner response. It is true again that merely a limited attenuation of the chrominance components does not influence the color picture too seriously. However, the accumulation of losses in various sections of the receiving system produces a detrimental influence on color presentation.

Tuner sensitivity and noise factor are more important in the reception of the color signal because of the inherently greater bandwidth of the color receiver and the fact that the i-f or tuner cannot be peaked in fringe-area reception of the color picture. Likewise, the nearness of the associated sound carrier to the color sub-carrier produces an interference beat between the two components at a frequency of appoximately 0.9 megacycle. To minimize this disturbance the fine-tuning control must be adjusted very carefully to give a minimum interference pattern on the color picture. Proper setting of the fine-tuning control positions the sound carrier correctly on the over-all i-f response, so that the 4.5-megacycle traps and tuned circuits function at peak efficiency in removing the sound information from the luminance and chrominance channels. Thus the technique of employing some form of automatic frequency control for the local oscillator appears advantageous.

The presence of an additional carrier in the video spectrum of the station also foretells the possibility of an increase in adjacent channel interference for two reasons. First of all, the inherent bandwidth of the receiver is broader in order to accommodate properly the high-frequency chrominance end of the bandpass. Likewise, the influence of adjacent channel interference can be more detrimental because of the presence of the chrominance sub-carrier near the adjacent channel picture carrier. There is also the possibility of a higher receiver sensitivity to the interfering picture carrier because of the increased bandwidth of the receiver.

I-F SYSTEM

The economy-style, narrow-band i-f system is not practical for peak color performance because of the necessity for holding up the bandwidth at the high chrominance frequencies. Thus the i-f system will have a wider bandpass and more stages. With the greater bandwidth there is additional responsibility imposed on the i-f system to provide maximum rejection by adjacent channel traps and proper setting of the sound carrier on the i-f response to minimize intercarrier interference or interference with the chrominance information.

The wide bandwidth and minimum interference requirements of the color television system also indicate that more critical alignment is required. In particular, the sweep alignment equipment will have to be linear, both amplitude- and frequency-wise, to permit precise alignment. Likewise, an accurate marker system is an absolute necessity for obtaining peak performance from a color receiver. For example, the response curve of the i-f (Fig. 44) system indicates that the bandwidth of a typical color receiver must be flat out to 41.65 megacycles to accommodate the chrominance information properly. Just 0.4 megacycle lower in frequency, the response of the i-f system must drop away almost to nothing at the sound-carrier frequency of 41.25 megacycles. This indicates to us that the marker accuracy will have to be within a small fraction of a megacycle to permit accurate alignment. Note that the position of the chrominance sub-carrier at 42.17 megacycles is only 0.9 megacycle above the sound-carrier frequency. Thus to prevent interference, these two frequencies must be precisely positioned on the over-all response curve, and the need for an accurate alignment procedure is emphasized.

Frequency response is important, not so much from the standpoint of amplitude nonlinearity as from that of the nonlinear time delay it causes. In the color system the rapid fall-off of the frequency response between the end of the video spectrum and the sound carrier (4 to 4½ megacycle range) produces a detrimental phase-shift at the high end of the video spectrum, while the use of vestigial sideband transmission and the positioning of the picture carrier at 50-per cent response results in a nonlinear time delay at the low end of the video spectrum. Circuits are included at the transmitter and receiver to compensate for this nonlinear time delay at the low end of the video spectrum. However, proper correction results only when the i-f bandpass and frequency response of the various sections of the color receiver meet specifications. For example, the proper positioning of the picture carrier at 50 per cent on the response curve is important if perfect reproduction is to be obtained. Thus the proper setting of the fine-tuning control, which can change the relative response at the picture-carrier frequency, is important in not only reducing the beat interference with the sound carrier but also in permitting proper registration of the various frequency sections of the luminance and chrominance information.

Improper delay characteristics in the color system can produce a number of defects—color fringing or spreading of color information at a point of sharp transition in picture information, imperfect registration of chrominance information on top of the luminance information (the red of an actor's lips displaced slightly with respect to the lips), and interaction between chrominance signals because of improper relative delay characteristics between the two chrominance channels.

17. *Color Receiver Circuit*

Typical circuits of an RCA Color Receiver are presented in the front-end-paper schematic, which shows the color section of the receiver starting at the video detector. The various sections of the receiver are outlined in heavy blocks, starting with the picture detector and composite video bandpass amplifier. The output of this video amplifier, which is the luminance signal *Y*, is applied to a luminance amplifier and then as a positive luminance signal to the color matrix output circuit. The chrominance information is passed through the bandpass amplifier, *V118-B*, to the demodulation circuits. Likewise, the burst pulse at the plate circuit of the first video amplifier is applied to the color synchronization circuit group. In this group the inserted carrier is generated and meets with the chrominance signal from the bandpass amplifier in the demodulation system. The *I* and *Q* signal components at the output of the demodulator meet with the luminance signal *Y* at the color matrixing system. The three primary-color signals are present at the output of the matrix unit and are applied to their respective grids of the tricolor tube.

The horizontal deflection circuit consists of a conventional synchroguide circuit with a direct-drive horizontal output circuit. In this section a special high-voltage formation-circuit is used to build up the transient pulse to a peak amplitude that can form a regulated 20,000 volts for the tricolor tube. Likewise, special horizontal- and vertical-convergence signals are formed for application to the convergence electrode and focus electrodes. The vertical deflection circuit consists of a vertical blocking oscillator and a vertical output stage. A suitable output is also removed for proper excitation of the convergence amplifier that is a part of the horizontal deflection block.

LUMINANCE SIGNAL PATH

Starting at the top left of the schematic diagram, we see that the video i-f signal is applied through a series inserted parallel-resonant trap *T113* which is tuned to the sound-carrier frequency at 41¼ megacycles to block the sound-signal components from the picture video-detector, *CR102*. The actual sound i-f signal is removed from the plate of the preceding i-f stage and applied to the sound channel. The presence of the sound trap before the picture detector minimizes the beat interference (920 kilocycles) that exists between the sound i-f carrier and the chrominance sub-carrier.

This section of the receiver contains the video detector, first video amplifier, and the bandpass amplifier of the chrominance section. A 1N60 crystal is used as the video detector, developing a negative-sync composite television signal across the diode load resistor *R190*. A variety of outputs is taken from the first video amplifier—composite video excitation for the sync circuit of the receiver, luminance output signal containing the *Y* signal, take-off point for application of burst to the color synchronization section, and chrominance signal for application to the demodulator circuit.

The luminance signal is developed across the first video-amplifier plate load resistor *R199* and is applied to the luminance amplifier. Associated with the plate output circuit is a resonant transformer *T114*, that is tuned to the sub-carrier frequency range and therefore presents a high impedance to the burst frequency which signal is emphasized and applied to the color-synchronizing circuits. At the same time its high series impedance to this same frequency range attenuates the burst frequency components that might enter

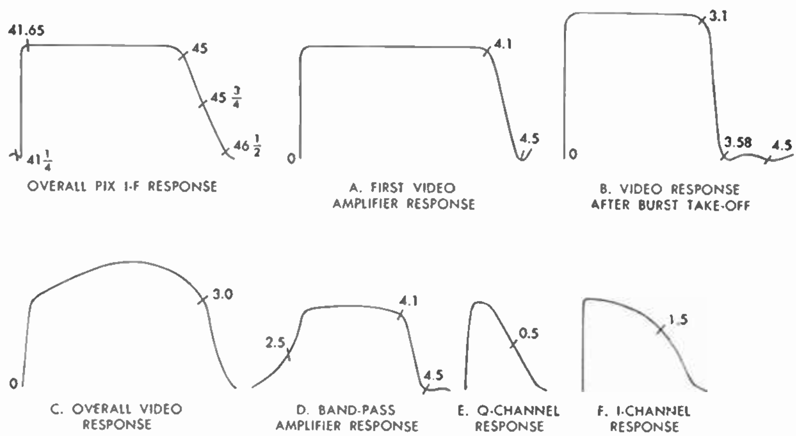


FIG. 44 Alignment Response Curves

the luminance channel. The activity in this video amplifier can be best demonstrated with typical response curves, Fig. 44. The over-all response of this first video stage is illustrated by response *A*, indicating a flat response out to 4.1 megacycles but a sharp drop-off at the sound-carrier frequency of 4.5 megacycles. After the burst take-off point the high-frequency end of the response curve drops (indicated by response *B*), showing the influence of the burst take-off transformer which is in the signal path to the plate load resistor *R199.*

A parallel resonant circuit is present in the cathode circuit of this first video amplifier and acts as a degenerative cathode trap that reduces to a very minimum the 4.5-megacycle response, as is indicated on response curve *A*. The cathode output is taken from across resistor *R196* and, through the small value capacitor *C184,* is applied to the grid of the bandpass amplifier. The bandpass amplifier, *V118-B*, amplifies only the chrominance channel frequency range, as is indicated by response *D*, showing maximum gain over the frequency range from 2.5 to about 4.2 megacycles. Thus the video response characteristics are such that a minimum amount of chrominance information enters the luminance channel and a minimum amount of luminance information enters the chrominance section. The bandpass amplifier response is, controlled by the short time-constant input circuit which de-emphasizes the low-frequency end of the video spectrum (small capacitor *C184* and low-value picture control *R198-A*) and the bandpass filter in the plate circuit which emphasizes only the desired frequency range. The chrominance information is taken off the chrominance control *R301* and applied to the balanced demodulators.

Two controls are associated with this section of a typical color television receiver—picture-contrast control and chrominance control. The picture contrast control is ganged with a similar control in the luminance amplifier. When the receiver has been adjusted properly, varying the contrast control will cause the same change in the luminance signal component as in the chrominance components and will keep the proper proportions of both signals. The chroma control, which, along with the contrast control, is a front-panel adjustment, determines the proper color saturation of the reproduced color picture. This control is adjusted until the hues appear most natural to the viewer. It is to be anticipated that viewers will differ in regard to what they consider as normal, and it can be expected that eventually some means of automatic chroma control or some standard method of adjustment of chroma will be evolved.

The luminance signal is next applied through a delay line *DL112* to the second video amplifier of the color receiver. The second part of the ganged contrast control follows the delay line and controls the luminance signal amplitude. This control is adjusted for a normal monochrome presentation (in terms of contrast and brightness), with the chrominance control set to a minimum. When the contrast control has been adjusted for a normal monochrome picture, the chroma control is turned up to obtain normal color presentation. The luminance signal at the output of the second video amplifier is applied through a series-peaking coil *L113* to the actual plate load presented by the matrix resistor network at the input of the matrix section. This input circuit has a lower impedance than that contributed by the plate load resistor *R208*. Thus the actual video load placed on the luminance amplifier is presented by the resistive input circuit of the matrix unit.

The delay line at the input of the luminance amplifier delays the video information approximately one microsecond. A delay system is necessary because of the three separate signals that pass through the television receiver to the color tube. Inasmuch as the bandwidth of the three channels is not identical, it is necessary to delay the information in the channel with the greatest bandwidth, presenting the information to the color tube in proper time coincidence with the chrominance information and preventing displacement of chrominance and luminance information.

The delay line can be a so-called artificial delay line made from coils and capacitors, or from a filter section, or it can be an actual piece of transmission line of proper length to produce a delay of approximately one microsecond. For example, in a typical receiver an 18-foot length of RG65 A/U cable can be used to obtain the required delay for the frequency response that is necessary.

CHROMINANCE SIGNAL PATH

The chrominance information is taken from the chrominance control at the output of the bandpass amplifier and applied to the grids of the *Q* and *I* demodulator tubes. These demodulators, referred to as "synchronous detectors," by means of a sub-carrier heterodyning principle produce the original *I* and *Q* signal components at their outputs. To produce these it is necessary that sub-carrier sine-wave signals, 90 degrees related, be inserted on the number three grids of the 6AS6 tubes. In the *I* demodulator the *I* signal and the subcarrier sine wave add vectorially to produce the *I* output signal. In this stage a vestigial sideband signal is present with a bandwidth extending from 1.5 megacycles below carrier to a ½-megacycle limit on the high-frequency side of the sub-carrier. This bandwidth is controlled by the output filter which must also have good rejection characteristics at frequencies above 1.5 megacycles with which to prevent the leakage of luminance information through the chrominance channel. Inasmuch as the sub-carrier sine wave applied to the *I* demodulator is in quadrature with the *Q* signal sub-carrier, there is cancellation of the *Q* signal sidebands, and no *Q* signal output exists in the *I* chrominance channel.

The *I* signal is of negative polarity at the output of the synchronous detector and is increased in amplitude by the *I* amplifier, tube V134-A, being formed as a positive *I* signal at the plate circuit of the amplifier. The positive *I* signal is applied through capacitor *C280* to the red matrixing circuit, where it is combined with the luminance and $+Q$ chrominance component for proper matrixing and reproduction of the red primary signal. Inasmuch as a $-I$ chrominance signal is necessary for proper matrixing in the green and blue channels, a phase-inverter follows the *I* amplifier. The phase-inverter produces a negative *I* chrominance signal in its plate circuit when the positive *I* signal is applied to its grid. A feedback arrangement through resistor *R356* is employed to equalize the amplitude of the plus and minus *I* signals—the negative *I* signal developed across resistor *R357* and the positive across resistor *R353*.

The *Q* signal applied to the top demodulator is in phase with the sub-carrier sine wave applied to its number three grid (Pin. No. 7) and results in the vectorial addition of the inserted sub-carrier and *Q* signal to produce the original *Q* signal modulation information in its plate circuit. In the *Q* channel the 90-degree relationship between inserted sub-carrier and *I* chrominance signal results in a cancellation of the *I* chrominance component, except in the vestigial sideband range between 0.5 and 1.5 megacycles where only a single *I* sideband is present and there is incomplete cancellation. However, the *I* sideband components between 0.5 and 1.5 megacycles are attenuated by the inherent bandwidth of the *Q* channel which extends only to 0.5 megacycle. This low-pass filter consists of the three inductors and capacitor *C270*.

The responses of the *I* and *Q* channels are illustrated in Fig. 44. The time delay of the chrominance channel and filters must also be considered in the design of the filter arrangement so that the *I* and *Q* chrominance signal components will both be properly timed with respect to the luminance information. The *I* channel signal must be delayed a fractional part of a miscrosecond because of its less bandwidth. Inasmuch as the *Q* channel has a lesser bandwidth than the *I* channel and amplifies double sideband components (instead of single sideband levels as in the *I* channel), it can have a higher gain, and a *Q* amplifier is not required. Thus the negative *Q* signal is applied to a simple phase-splitter, which develops a minus *Q* signal from cathode to ground for application to the green matrix unit, while a positive *Q* signal is developed in its plate circuit for application to the blue and red matrixing resistors. A gain control is located in the cathode circuit of the *I* demodulator in order to permit proper relative gain adjustment between *I*, *Q*, and luminance signal components at the input of the matrix unit. The *I* gain control permits proper setting of the *I* channel signal with respect to the *Q* channel signal, and therefore, no gain control need be associated with the *Q* channel.

The three separate color channels of the matrix unit are identical with the one exception that no gain control is associated with the red channel—the red channel serves as a reference for adjustment of the green and blue channel gains. Each section of the matrix unit consists of an adder, an output stage, and a d-c restorer for separate excitation of the three guns of the color tube.

In the green channel the plus *Y*, minus *I*, and minus *Q* signals are combined at the matrixing resistors, *R210, R211*, and *R212*. Through capacitor *C282* and the feedback gain control for green, *R216-A*, the signal is applied to the grid of the green adder. At this point only the green video signal remains, and it is increased in amplitude by tube V135 and applied as a negative video signal to the grid of the green gun. Inasmuch as three separate signals are developed in the matrix unit for separate excitation of the guns, it is necessary to employ a d-c restorer to retain a true average-brightness level and thereby prevent the changes in average brightness of the transmitted scene from affecting the color balance (at high- and low-brightness levels), once the color background and matrix gain controls have been adjusted properly.

Inasmuch as high-frequency video, as well as chrominance information, is passing through the matrix unit, the associated stages must have a good high-frequency response. Consequently, low-frequency degeneration and feedback are employed to maintain the proper over-all response. At high frequencies proper filtering exists in the cathode circuit to produce peak amplification, while the low-value cathode capacitor *C289* and the degenerative feedback through resistor *R214* reduce amplification at lower frequencies, thereby maintaining a more uniform over-all response. No gain control is associated with the red channel, because it requires a greater excitation to drive the red phosphor to a given light output as compared to the green and blue phosphors. Thus the gain controls are associated with the green and blue channels and are adjusted properly with respect to the maximum gain red channel to produce the proper color signal voltage level at the grid of the tricolor tube.

In the d-c restorer circuits there are three separate brightness controls for each restorer circuit, permitting the brightness levels of the three guns to be set in proper relation to each other, and d-c restorer action will retain relative brightness levels with changes in picture brightness.

The bottom control *R124-A*, referred to as the master brilliance, sets the d-c component of grid bias applied to all of the guns of the tricolor tube, while the green and blue background controls regulate the d-c bias to the green and blue guns individually. Thus with these three controls the over-all brightness, as well as the individual relative brightness levels, can be adjusted correctly. The accelerating grids are also supplied potentials that are individually controlled to assist further in obtaining proper over-all brightness, proper relative brightness, and correct color balance. In general, it can be said that the screen controls are adjusted for proper white while the background controls are adjusted for proper rendition of the high lights and low lights of a typical color-bar chart or picture. When the green and blue gain controls are properly adjusted and a monochrome picture is observed, there should be no color contamination over the entire brightness range.

COLOR SYNCHRONIZATION

In the color synchronization block the sub-carrier color burst is first removed from the horizontal backporch and is then used to control the frequency and phase of a locally generated sub-carrier sine wave. The sub-carrier burst is first applied to a so-called burst amplifier *V129-A*, which amplifies and keys the burst information. The burst amplifier is keyed on by a negative fly-back pulse applied to its cathode from the horizontal output transformer *T117*. Thus the burst amplifier will conduct only during the horizontal blanking period and is held at cutoff by the positive cathode voltage during the interval between the fly-back pulses. This same fly-back pulse of negative polarity is also applied through capacitor *C267* to the screen grid of the bandpass amplifier above, tube *V118*-B. Here it prevents the horizontal blanking and sync pulses from entering the chrominance channels. Their absence in the chrominance channels prevents the operation of the d-c restorers on the burst information instead of the desired horizontal sync pulse arriving through the luminance channel.

The phase discriminator transformer *T122* in the plate circuit of the burst amplifier is resonant to 3.58 megacycles and produces the reference sine wave that is applied to the phase detector (diode-operating tubes with plates acting as shields). These balanced out-of-phase sub-carrier sine waves are then compared with the locally generated sub-carrier signal that enters the phase detector circuit from the color-phasing amplifier through the resonant transformer *T123*. A balance control *R287* permits adjustment of the phase detector so that a zero potential appears on the reactance input capacitor *C239* when the burst frequency and locally generated sub-carrier are on frequency and in phase.

A triode crystal oscillator, tube *V131-B*, with a cathode output is used to generate the local sub-carrier sine wave, permitting better isolation from the reactance tube. With the input circuit of the reactance tube grounded, the plate inductor *L300* of the reactance tube is used to tune the crystal circuit for oscillations. The reactance tube functions as a capacitive type and varies the frequency of the crystal oscillator in accordance with the d-c potential developed across capacitor *C239* by the phase-correcting voltage from the phase detector.

Two outputs are taken from the crystal oscillator—one is the sub-carrier output that will be used in the demodulator and the second is coupled through a capacitive voltage divider *C258* and *C256* to the grid of the color-phasing amplifier. The color-phasing amplifier is used for isolation purposes and to amplify the sub-carrier frequency before its application to the phase detector. At the same time the resonant output transformer and a special phasing control are used to adjust the phase relationship properly among the locally generated sub-carrier, the received sub-carrier burst, and the proper phasing of the demodulator sine wave at the synchronous detector. True color fidelity (most often judged by the most natural flesh-tones) is obtained by means of adjustment of the phase control, resistor *R327*, while the color picture is being observed. Once the proper color rendition is attained, the color synchronization system will maintain this established phase relation between the sub-carrier and the received sub-carrier burst.

A special color-killer circuit, tube *V119-A*, is also a part of the color synchronization system and cuts off the bandpass amplifier during the reception of a monochrome signal. Thus video information does not pass through the bandpass amplifier and into the chrominance channel during monochrome reception. The color-killer tube is normally cut off during the reception of a color signal because the presence of the sub-carrier burst keeps a negative potential at the junction of resistor *R288* and capacitor *C317*, placing a negative charge on capacitor *C318* and keeping the color-killer tube biased beyond cutoff. When a monochrome signal is being received and no color burst is present, the color-killer tube is triggered into conduction by a positive horizontal pulse developed during fly-back time across a special winding of the horizontal output transformer. This positive pulse at the plate causes the tube to conduct and charge negatively the time-constant circuit which consists of capacitor *C252* and resistor *R296*. The negative charge is also applied through a de-coupling network, resistor *R197* and capacitor *C183*, to the grid of the bandpass amplifier, holding it at cutoff during the reception of a monochrome signal.

The 3.58-megacycle, locally generated sine wave for the *I* channel is removed from the cathode take-off transformer *T124* and applied to the *I* demodulator. A 3.58-megacycle component is also applied to the grid of a quadrature amplifier, referred to as the *QCW* amplifier, which has a resonant double-tuned transformer in its plate circuit. Consequently, it functions as a 90-degree phase-shifter, producing a quadrature continuous-wave signal for application to the *Q* demodulator. The locally generated sub-carriers that are used to insert the demodulating sub-carrier are often called the *ICW* and *QCW* signals. It is important to realize that the relative phase between the two locally generated sine waves is established in this circuit but that the actual phase relationships between these signals and the sub-carrier burst is determined by the phase control *R327* at the output of the color-phasing amplifier.

DEFLECTION CONVERGENCE CIRCUITS

The deflection system of the color television receiver is conventional in terms of the sawtooth generating circuits and output stages. However, the horizontal output transformer and the associated circuits required for proper stabilization of the high voltage and for proper convergence of the beams from the three electron guns become more complex. Thus in the color-deflection system it is not only a matter of obtaining a single in-focus beam that scans the fluorescent screen linearly but of obtaining three individual beams that must each scan the screen linearly and in proper position with respect to the other two. As a result, the degree of linearity required is more critical if we are to have true color rendition on all segments of the screen.

In the RCA receiver a conventional synchro-guide circuit, tube *V127*, is used to drive the 6CD6 horizontal output stage. This is a direct-drive output circuit, and transformer *T117* is a special transformer that is required in order to obtain suitable deflection, proper voltage boost, the proper step-up ratio for developing the high voltage, and a special winding for other receiver functions (such as operation of burst amplifier, color-killer, etc.).

The direct-drive transformer supplies horizontal deflection energy to the deflection yoke at terminals 1 and 3. The damper diode, tube V125, is designed to withstand the high peak voltage present and acts as a suitable damper of the fly-back transients. The width control *L119* is, in effect, in parallel across the deflection yoke and controls the picture width without having any great influence on the value of the high voltage. Although the linearity and damper circuit appear conventional, it is necessary to introduce centering current to the deflection yoke without shunting the yoke to a-c ground. To provide this isolation the actual linearity control is of a bi-filar construction, and the actual B+ (present at bottom side of capacitor *C339*) enters the centering and deflection yoke circuits through one winding of the bi-filar coil of the horizontal linearity control. This arrangement presents a high impedance between the yoke and plus-B (plus-B is at a-c ground potential) but, at the same time, permits a centering current to be developed from the voltage across the capacitor *C208*. The centering current is regulated by potentiometer *R249*, and the centering current can be controlled in amplitude and direction of flow as a function of the position of the arm with respect to the resistor center tap (center tap connects to the horizontal deflection coils).

The transient pulse at the top of the direct-drive transformer is applied to a voltage-doubler circuit, consisting of capacitors *C204*, *C205*, and *C203* with a high-voltage rectifier and doubler, tubes V121, V122, and V123. In the charging process capacitor *C204* is charged initially by the arriving burst that drives the first diode into conduction. Between bursts, capacitor *C204* discharges through the diode coupler and, over a period of time, will charge capacitor *C205* up to the peak amplitude of the pulse. Thus capacitor *C205* and the pulse itself are effective in causing the last diode (referred to as the high-voltage doubler) to conduct and charge capacitor *C203* to twice the amplitude of the transient pulse. The diode coupler functions as a time-constant circuit for the charging of capacitor *C205* but provides better stability than a series of resistors would.

A shunt-regulator tube V120 provides a constant 20,000-volt anode voltage for application to the color tube. It does so by placing a load on the output of the high-voltage supply that is a function of the bias applied to its grid via the arm of the potentiometer *R245*. When the output voltage tends to decline (as it does in a transition from a dark to a bright scene), the grid voltage on the regulator tube drops as well, and the regulator tube presents a lighter load to the high-voltage output and restores it to the proper level. A constant reference is supplied to the cathode of the regulator tube from the +400-volt low-voltage supply source. The d-c convergence voltage is also removed from the resistor network across the high-voltage supply.

Still another rectifier is connected across a portion of the direct-drive transformer, and this rectifies a pulse component of proper amplitude in order to develop the focusing potential. The focus voltage is taken from a potentiometer in the cathode circuit of the rectifier. This focus voltage, however, is modulated with a convergence signal before it is applied to the focusing electrodes of the color tube. Thus the focus voltage is applied through the secondaries of the convergence output transformers *T115* and *T116* before application to the picture tube.

The purpose of the convergence circuit is to modulate both the d-c convergence voltage and the focus voltage with a suitable correction waveform. The wave shape of the modulation is parabolic and in the form of a series of arcs at a radius of curvature that corresponds to the faceplate curvature of some monochrome picture tubes. This voltage, which is in the form of an arc, compensates for the fact that the distance between guns and phosphor dots is not the same at the end of the raster as it is in the center area. As a result, it is necessary to make two corrections. First, it is necessary to bring each beam into focus, despite this varying distance, and second, it is necessary to make the three beams converge properly (at varying distances from the gun) so that each beam strikes its associated color dot. The parabolic modulation (in the form of an arc of a circle) must therefore modulate not only the convergence potential which controls the convergence of the three beams but also the focus potential which controls the focus of each beam individually. Both horizontal and vertical convergence voltages need to be developed because of the horizontal and vertical motion of the scanning beam. Convergence voltages may be formed from signals derived from horizontal and vertical output tubes.

The horizontal convergence excitation is derived from a potentiometer in the cathode circuit of the horizontal sweep output tube, while vertical excitation is derived from a similar potentiometer in the cathode circuit of the vertical output tube. Because of the filtering action of the cathode capacitors the wave shape present in the cathode output circuit is parabolic and in the form of an arc. The extent or curvature of the arc for the horizontal convergence waveform is controlled by the inductor in the grid circuit of the convergence amplifier, tube V119-B. It is important to realize that the actual amplitude and shape of the correction waveform determine proper convergence at the center and edges of the screen. The amplitude of the wave controls the degree of correction and must be set for an optimum value, while the shape of the wave maintains correction uniformity from the center to edges of the raster. A similar shaping control, *R238* is employed and uses a feedback arrangement for the vertical convergence waveform.

The output transformers *T115* and *T116* are resonant at the line and field rates, thus emphasizing the convergence waveforms to produce the necessary 800- to 1500-volt peak convergence signal required for proper correction for a typical color tube. The two convergence waveforms combine in the transformer secondary and are applied through capacitor *C193* to the convergence electrode and through transformer taps, which reduce the amplitude of the correction voltage, to the focus circuit of the tricolor tube.

A number of convergence circuit controls are required. With the other convergence controls turned down, the d-c convergence control *R243* and the convergence magnets are first adjusted for proper convergence of the dot pattern at the center of the screen. Next, the so-called dynamic-convergence adjustments are made in order to produce correct dot overlap at center top, bottom, and sides of the raster. The horizontal dynamic-convergence control, for example, permits the adjustment of the dots from left to right on each side of center. If the dots can be made to converge on one side but not on the other side of center, the actual dynamic waveform must be varied by adjusting the phasing control L114. Similar controls, potentiometer *R250-B* and shaping control potentiometer *R238*, regulate the same functions in terms of vertical convergence. Inasmuch as the horizontal and vertical convergence adjustments are interacting, it is necessary to continue readjusting the various controls until the best convergence is attained.

The vertical deflection circuit is conventional, using a vertical blocking oscillator and vertical output tube to drive the vertical deflection coils through transformer *T120*. A vertical linearity control and the vertical convergence control are parts of the cathode circuit of the vertical output tube.

18. *Narrow-Band Chrominance Channel*

A narrow-band chrominance channel can be incorporated in a color receiver at a saving in cost but a sacrifice in color detail. The economy-type receiver takes advantage of the color-difference phasing of the demodulators. For example, if the inserted sine wave from the sub-carrier frequency generator at the receiver is made to coincide in phase with the color-difference angles, Fig. 27, the received chrominance signal can be demodulated into color-difference signals instead of the I and Q signal components. In this system, Fig. 45, the

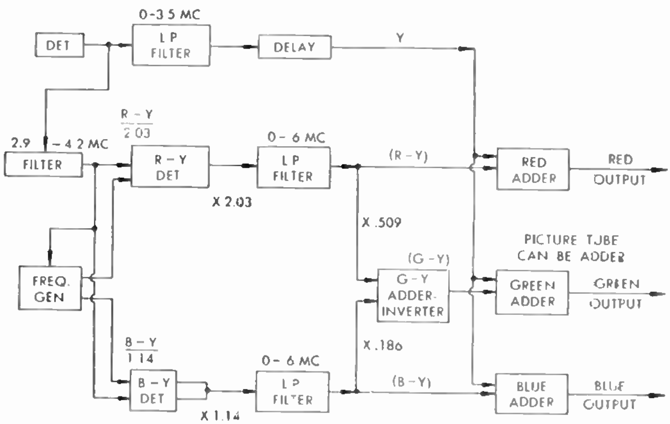


FIG. 45 Narrow Band Decoder

bandpass of the chrominance channel is limited to a range from zero to 0.6 megacycle in each channel to prevent crosstalk and interference. In effect, we are detecting the signal as two double-sideband components and intentionally sacrificing the wider-band I signal information in favor of circuit simplification.

In the narrow-band system the output of the $R - Y$ and $B - Y$ channels are combined and inverted to form the $G - Y$ signal. The luminance information is then added to the three color-difference signals, producing the three primary-color signal voltages. It is conceivable that the actual addition of the color-difference signals and the luminance signal can take place in the picture tube itself—by application of the color-difference signals to the individual grids of the guns in the picture tube, while the luminance signal is applied simultaneously to all cathodes of the color tube. Thus at least two adder stages are eliminated, and since the chrominance channels have equal bandpass, the problem of equalizing their relative delays does not exist. Likewise, the complexity of the bandpass filters required normally to eliminate interference between I and Q chrominance components is not present.

The matrixing of the color-difference signals can be accomplished in a simpler manner by regulating the gain of the various channels. For example, with the relative NSTC values for the color-difference signals, as indicated in Fig. 45 and on the vectors of Fig. 27, it is possible to derive the color-difference signals at proper relative levels in the demodulator system. If the relative-gain relationship between the $R - Y$ demodulator channel and the $B - Y$ demodulator channel is 2.03 over 1.14, the actual output of the $R - Y$ channel will be the $R - Y$ color-difference signal, and the output of the $B - Y$ channel will be the $B - Y$ color-difference signal. If we rearrange the standard luminance formula and present this information in the form of the $G - Y$ component, we find that by suitably mixing proper percentages of $R - Y$ and $B - Y$ color-difference signals with a 180-degree phase reversal (negative signs in front of $-(R - Y)$ and $-(B - Y)$ components in formula), it is possible to form the $G - Y$ color-difference signal, as follows.

$$\begin{aligned}
Y &= 0.59G + 0.3R + 0.11B \\
Y - Y &= (0.59G - 0.59Y) + (0.3R - 0.3Y) + (0.11B - 0.11Y) \\
0 &= 0.59(G - Y) + 0.3(R - Y) + 0.11(B - Y) \\
0.59(G - Y) &= -0.3(R - Y) - 0.11(B - Y) \\
G - Y &= -0.509(R - Y) - 0.186(B - Y)
\end{aligned}$$

Thus if the relative gain in the $G - Y$ adder is 0.509 to the $R - Y$ color-difference signal and 0.186 to the $B - Y$ color-difference signal, a $G - Y$ signal is present at the output of the adder and inverter. This example shows how the color-difference signals can be formed with a much simpler matrixing arrangement.

To prove mathematically that the color-difference signals $B - Y$ and $R - Y$ also carry the $G - Y$ information, the luminance equation is rearranged as follows:

$$\begin{aligned}
Y &= 0.59G + 0.3R + 0.11B \\
B - Y &= B - (0.59G + 0.3R + 0.11B) \\
B - Y &= -0.59G - 0.3R + 0.89B \text{ and} \\
R - Y &= R - (0.59G + 0.3R + 0.11B) \\
R - Y &= -0.59G + 0.7R - 0.11B
\end{aligned}$$

Now, by substitution of the two color-difference signals, the $G - Y$ resultant is obtained and matches the $G - Y$ signal derived by direct rearrangement of the luminance formula:

$$\begin{aligned}
G - Y &= -0.509(R - Y) - 0.186(B - Y) \\
G - Y &= -0.509(-0.59G + 0.7R - 0.11B) - 0.186(-0.59G - 0.3R \\
&\quad + 0.89B) \\
G - Y &= 0.41G - 0.3R - 0.11B
\end{aligned}$$

Direct:

$$\begin{aligned}
G - Y &= G - (0.59G + 0.3R + 0.11B) \\
G - Y &= 0.41G - 0.3R - 0.11B
\end{aligned}$$

19. *Decoding of NTSC Signal For Single-Gun Presentation*

When a single-gun color tube is used, the color information must be keyed in a definite color sequence to the control grid circuit. To do this it is necessary to form a keying or gating signal that corresponds to the rate at which the scanning beam changes from one color of phosphor to another. For example, in planning a decoding system for the chromatron tube, it is necessary to key the individual colors at the same rate that the scanning beam moves across the individual color strips, Figs. 42 and 46. Note that in color scanning of the phosphor strips of the chromatron tube, the colors are scanned in the order: red, green, blue, green, red, in one cycle—or green is scanned twice in one cycle as compared to one scan of red and one of blue. Consequently, in the keying or gating process it is necessary that the keying of the green signal occurs at twice the rate of the keying of the red and blue signals.

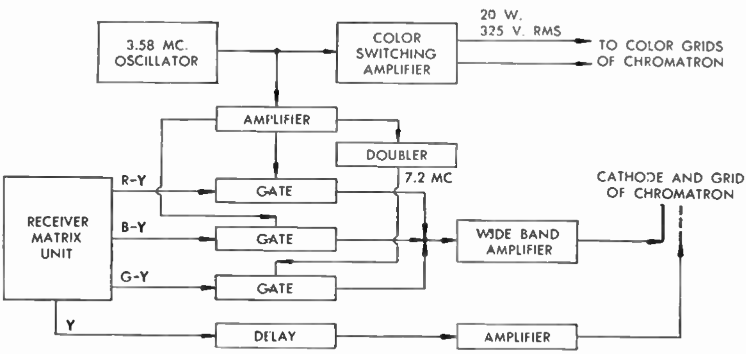


FIG. 46 Decoding NTSC Signal for Chromatron

A typical plan for keying the color-difference signal of the NTSC system is illustrated in Fig. 46. The color-difference signals are applied to three so-called gate tubes. To the same tubes we also apply a 3.58-megacycle keying wave which, through an amplifier and phase-splitter, is applied to the blue and red gates. After frequency-doubling the sine wave is also applied as a 7.2-megacycle keying signal to the green gate.

The keyed output of the gating stages is applied through a wide-band amplifier to the cathode of the chromatron tube. The signal is in the form of sharp pulses of color information that represent the amplitude of a given color at given instants of time. Thus, a dot pattern is traced across the individual color strips of the chromatron tube, and color information is presented in a dot-sequential fashion. Therefore, the gating process has, in effect, changed the NTSC color signal to a dot-sequential color signal for proper excitation of a single-gun color tube. The excitation of the individual color strips can be controlled by regulating the pulse width of the video signal contributed by each gate. Consequently, proper correction can be made for the relative efficiencies of the individual color strips.

The luminance signal which carries the picture detail is passed through the usual delay circuit and video amplifier to the control grid of the chromatron tube. Likewise, the 3.58-megacycle sine wave from the generator is increased in amplitude by a color-switching amplifier which builds up the sine wave to a level of approximately 300-volt rms, which is applied to the color grid of the chromatron and controls the color-scanning order of the electron beam. It is to be noted that the gating process and the color-scanning process are accomplished with the same signal and are therefore synchronized. The signal applied to the color grid makes certain that the scanning beam strikes a specific color strip at a given instant, while the gating process makes certain that the control grid, at that very same instant, is excited by a signal that represents this color.

Inasmuch as the NTSC signals are simultaneous ones, it is possible to sample or key the individual color signals even at as low a rate as the field frequency and thereby to convert the simultaneous NTSC signal into a sequential color display. In fact, a complementary approach can even be used at the transmitter. For example, it is possible to use a field sequential-color camera and convert the output signal of that camera into the color-difference simultaneous signal that is the standardized method of transmission.

20. *Color Receiver Adjustment and Alignment*

The following quoted information (pages 76 to 86), copyright 1953 by the RCA Service Company, Inc., is reproduced through the courtesy of the Radio Corporation of America.*

Receiver Set Up Procedure

Step One: Installation of Kinescope

a. Remove cabinet top (remove two screws and the entire top comes off).

b. The next step is the installation of the RCA Tri-Color Kinescope. The tube is somewhat bulkier than a conventional black and white kinescope of comparable screen size and care must be taken in handling.

1. Install the mu-metal shield on kinescope.
2. Install yoke assembly.
3. Install purifying coil on neck of kinescope with external leads to the rear.

c. The entire yoke, purifying coil, mu shield and kinescope assembly should be installed in cabinet.

d. Align the internal mask of the tube with the front of the cabinet, keeping the blue gun to the top.

Step Two: Location of Purifying Coil

a. Leave the yoke assembly in its normal position, but do not tighten down.

b. Set the purifying coil just forward of the kinescope gun structure. The back edge of the coil should be 4½ inches from the edge of the kinescope socket. If the purifying coil comes much closer to the kinescope gun structure, focus distortion of the beams may result.

c. Plug in the high voltage lead, yoke, purity coil and kinescope socket.

d. Turn set on.

* References to illustrations have been deleted.

STEP THREE: LINEARITY AND SIZE ADJUSTMENTS

a. Turn the chroma control R302 off [refer to front end papers] * (CCW †) to the minimum (no color) position.

b. Tune in a black and white transmission in the normal manner by adjusting fine tuning, contrast, brightness and AGC.

c. Make sure that the linearity of the picture is correct, as any adjustment of the linearity controls will affect dynamic convergence voltages.

d. Be very careful *not* to overscan the picture, as this will affect color purity. This has the effect of producing secondary electrons in sufficient quantity to contaminate color purity. Use the centering controls to determine effective raster size.

e. Check the horizontal oscillator control circuit to make sure it is correct. Any later adjustment will change horizontal dynamic voltages.

STEP FOUR: HIGH VOLTAGE SET UP

a. Using a high voltage probe with a vacuum tube volt-meter, measure the high voltage at the kinescope. The voltage should measure 20,000 volts.

b. If the voltage is too high or too low, adjust the voltage regulator control R245 until the proper 20 KV is obtained.

c. With the meter connected, adjust the brightness control from a low level brightness to a high level brightness. There should be no appreciable change in voltage. If the voltage does vary, a slight adjustment of the voltage regulator control may be necessary.

STEP FIVE: COLOR PURITY ADJUSTMENTS

a. Turn the contrast control R198 to a minimum to remove the picture signal.

b. Turn the red screen R227 up (CW ‡).

c. Turn the blue (R225B) and green (R225A) screens down (CCW). . . . [Result should be a red raster.]

d. Pull the convergence magnets out away from the neck of the kinescope.

e. Slide the yoke to the rear as far as possible.

f. The color purifying coil produces a uniform transverse magnetic field which serves to orient the three electron beams with respect to the central axes of the kinescope. Adjust the purifying coil rotation in connection with the purity current control R106, until the most uniform red appears in the *central area* of the tube. Disregard the purity condition along the sides and at the top and bottom. As little purifying coil current as possible should be used.

g. Adjust the screen purity for most uniform red by sliding the yoke forward until best purity is reached. Make sure that there are no neck shadows at this point due to improper yoke location. If some purity contamination occurs [not a pure uniform red], slight readjustment of the purifying coil and purity current may result in better purity.

h. Next check the blue purity by turning down the red screen R227 and turning up the blue screen. Green purity should be checked in the same manner. If some contamination results it may be necessary to make a compromise adjustment for best results on all three fields.

i. Color purity may be affected by an external magnetic field. If color purity makes an abrupt shift at any time during the adjustment procedure, look for an external field causing trouble (magnetic tools, etc.).

* [Brackets indicate author's addition.]
† [Counterclockwise.]
‡ [Clockwise.]

STEP SIX: SETTING THE KINESCOPE TO WHITE

a. With chroma and contrast controls to a minimum turn the brightness control to a maximum.

b. By the use of the red, blue and green screens adjust for a low brightness white.

STEP SEVEN: CONVERGENCE ADJUSTMENTS

A. General Considerations.

The object of the convergence adjustments is to cause the beams from the three parallel guns to be bent toward the central axis of the kinescope and to converge at the aperture mask. This is accomplished by the application of a high d-c potential modulated with signals of horizontal and vertical frequency. These signals have an approximation of a parabolic wave shape and are used to compensate for the longer distance the beam has to travel to reach the edges of the tube. The modulated DC, if used in conjunction with three permanent magnets could be considered a vernier convergence adjustment. Convergence adjustments are indicated as being best accomplished by a dot generator. This device provides a means of obtaining white dots on a black background. The [procedures] that deal with convergence are based on the dot system of convergence, although this does not necessarily mean that convergence can *only* be done by this method, or that this method as presented is the best.

B. Dot Generator Adjustment

The dot generator is adjusted by means of horizontal and vertical controls to get the dots approximately square and about one half an inch apart. Set the brightness control to a normal position to prevent "blooming" of the white dots.

C. Initial Receiver Adjustment

Turn the horizontal dynamic wave form amplitude control R250A to a minimum (CCW). Turn the vertical dynamic wave form amplitude, control R250B, and shaping control, R238, to a minimum (CCW).

Pull the convergence magnets as far away from the neck of the kinescope as possible.

Turn the DC convergence voltage R243 to a low value (CCW). This will give a pattern with the blue dot low, red dot to the right and green dot to the left.

[A] reverse triangular situation indicates convergence voltage is too high.

D. DC Convergence Adjustment.

Simultaneously adjust DC convergence voltage and position of the three convergence magnets to get dot overlap in the center of the raster. Proper dot overlap will cause single white dots in the center area of the kinescope.

NOTE: *Many of the convergence adjustments are interdependent; i.e., adjusting one may affect one or more other adjustments. As an example, adjusting the kinescope convergence magnets affects the* DC *convergence control setting; adjusting* DC *convergence affects the focus adjustment. It will be well to note that the convergence magnets should not be positioned too close to the kinescope neck in final adjustment, or beam focus distortion may result.*

E. Dynamic Convergence Adjustment.

The dynamic convergence adjustments are made next. Vertical and horizontal dynamic convergence adjustments provide correct dot overlap of the dot generator pattern at the top, bottom and sides of the raster. The two controls are interacting, and the adjustment of one affects the adjustment of the other. It has been found best to adjust vertical dynamic convergence first.

Vertical Dynamic Convergence

Adjust the vertical dynamic convergence control (R250B) until the extreme top and bottom dots show an equal displacement error.

Adjusting the DC convergence control will converge the dots on a vertical line down the center of the raster.

Horizontal Dynamic Convergence

Adjust the horizontal dynamic control R250A until the extreme left and extreme right center dots are equally displaced.

A change in DC convergence will converge the dots on the horizontal center line. If the horizontal dynamic voltage appears in phase error (as observed by one side of the raster not converged) it may be corrected by adjusting the horizontal dynamic phasing control L114.

F. Dynamic Convergence Check.

Because the horizontal and vertical convergence adjustments interact it will be necessary to check back and forth until the best overall convergence is reached over the major portion of the tube.

As a check on the accuracy of the dynamic wave forms, variation of the *DC* convergence voltage will indicate if further adjustment of dynamic voltages are needed. If, through the range of DC convergence voltage, the dots do not overlap, convergence cannot improve with dynamic voltage [adjustment]. If convergence at the edges will improve with [change] of DC convergence voltage, adjustment of the dynamic voltages will improve convergence. As an example, if in the area of mis-convergence the blue dot were low it would indicate that more dynamic voltage need be applied. If the blue dot were high it would indicate that the dynamic voltages were too high and [would] need [to] be reduced.

NOTE: *It has been found that in some tubes purity is improved if the tube is converged before purity adjustments are made. After adjusting purity in this manner, convergence must again be set.*

STEP EIGHT: WHITE ADJUSTMENT

(Remove the dot generator.)

a. Turn the chroma control and contrast control to minimum.

b. Turn brightness control to maximum.

c. Adjust the red screen control R227, blue screen R225B, and green screen control R225A to a low brightness white. The "color" of the white should be about the white produced by a low brightness setting on a standard black and white kinescope.

STEP NINE: HIGHLIGHT ADJUSTMENT

a. With the brightness still fully clockwise, turn the control to some mid-position.

b. Tune in a black and white picture.

c. Adjust the blue background control R216B and the green background control R216A until a white (same reference white as in step eight) appears on the high brightness high lights in the picture.

STEP TEN: LOW LIGHT ADJUSTMENT

a. With the contrast control still at mid-position, turn the brightness control down to a reduced value. Adjust the blue (R233A) and green (R233B) background controls to reach an equal white on the low lights in the picture. Steps Nine and Ten should be repeated until the "white" is made to track from high lights to low lights using the contrast control.

Servicing the Color Television Receiver

[A study of] the schematic diagram of the color TV receiver [indicates] the service problems . . . seem to be much greater than those in black and white receivers. Although some new and unfamiliar circuits are encountered, servicing need not be exceptionally difficult if a logical pattern of good trouble-shooting technique is followed. The expert technician follows the system of block diagram analysis to quickly isolate troubles. The system divides the receiver into sections or blocks according to functions. However, the purpose and theory of operation of each block of the receiver must be thoroughly understood by the technician in order [for him] to become proficient in the diagnosis and location of troubles. The schematic diagram of this RCA color receiver has been divided into functional blocks, not only for the purpose of understanding the operation of the receiver, but also as an aid to developing a servicing technique.

In isolating troubles in a color receiver it is well to remember that the receiver is a basic reproducer of black and white pictures, and that certain additional circuits are devoted entirely to color. It is logical to assume, then, that the first step to locate receiver malfunction would be to observe the reception of a black and white transmission.

CHECKING BLACK AND WHITE RECEPTION

Operating the color receiver as a black and white receiver will reveal:

1. Faults in circuits common to color receivers and black and white receivers. . . .
2. Faults in producing pure primary color fields (kinescope block).
3. Faults in convergence and focus (Kinescope, Vertical Deflection and Convergence blocks).
4. Faults that prevent the forming of uniform white or gray scale. (Kinescope and Color Matrix blocks.)

Pure Primary Color Field Troubles

Receiver faults in this category have the effect of producing a primary color shading over the entire field of an otherwise normal black and white picture. This type of malfunction is confined to the Kinescope block and should be located by following the procedure for obtaining color purity, as described in *Receiver Set Up Procedure.*

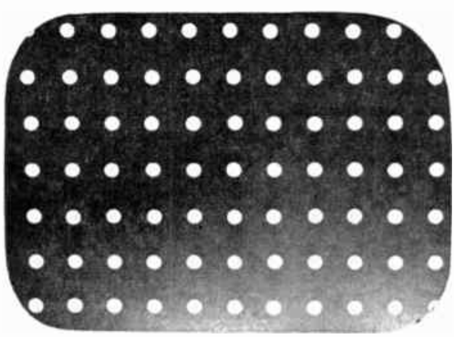


Fig. 46a Proper Dot Convergence

Convergence and Focus Troubles

If proper convergence over the entire face of the kinescope cannot be obtained, Fig. 46a, by following the setup procedure outlined in the previous section, the following steps should be taken:

1. If the tube cannot be made to converge in the center of the raster through a range of the DC convergence control, a check of the convergence voltage should be made. The voltage may be either too high or too low. Use of a meter with a high voltage probe will confirm this.
2. If the raster will converge in the center but both sides are out, check the horizontal dynamic voltages by rotating the control and noting the difference. If no difference is noted, check the dynamic convergence circuits with an oscilloscope. If convergence voltage is present but has insufficient amplitude, check the peaking of the convergence amplifier circuits. Remember also that any malfunction in the Horizontal Deflection or Horizontal Oscillator circuits will affect the convergence voltage.
3. If the raster comes into convergence on one side before the other, a recheck of the horizontal phasing adjustments should be made. Check the Convergence Amplifier tube and the Horizontal Output Amplifier tube in addition to the phase coil adjustment.
4. If the raster will converge on the sides but not at top and bottom, a check of the vertical dynamic amplitude and phase should be made. A fault in the Vertical Output stage can cause this voltage to be off.
5. If the convergence voltages are correct and the raster cannot be made to converge through a range of the DC convergence control, the kinescope should be replaced.
6. An effect of improper convergence may be noted when there is a fault in the Color Matrix block resulting in poor frequency response of one video stage only. This can be seen on the kinescope as a fringe of one color around the white dots of a convergence pattern. Check for peaking and proper response in the Color Adder section involved.
7. An effect very similar to the trouble outlined in 6 may be caused by low emission in one of the guns of the kinescope. The beam from the defective gun will defocus and "bloom" at the face plate causing a color fringe around the convergence pattern dots. The obvious remedy is to replace the kinescope, but only after making sure that proper voltages are present on the kinescope.
8. If an abrupt shift in convergence is noted make sure that the high voltage regulator circuit is working properly. A shift in high voltage will shift convergence voltages also.

Uniform White or Gray Scale

While viewing a black and white picture it should be possible to vary the brightness control from a high to a low setting without altering the "color" of white or gray. In other words, after the correct white is obtained on a raster there should be no color contamination of the highlights or low lights in a black and white picture. If such a condition does not exist, then "tracking" adjustments must be made as follows:

Highlight Adjustment

a. With the brightness control turned clockwise, turn the contrast control to some mid-position.

b. Tune in a black and white picture.

c. Adjust the blue gain control R216B and the green gain control R216A until a white of the same "color" as the raster appears on the high brightness high lights in the picture.

Low Light Adjustment

a. With the contrast control still at mid-position, turn the brightness control down to a reduced value. Adjust the blue (R233A) and green (R233B) background controls to reach a match on the low lights in the picture. It may be necessary to repeat the Highlight and Low Light adjustments until "tracking" is obtained.

CHECKING COLOR RECEPTION

If no troubles are apparent on a black and white picture, reception of color transmissions should next be checked. If any picture fault is seen, it is logical to assume that the fault is located in the Color Sync Block or Demodulation—Phase Inversion block. Troubles can occur in these two blocks that will not affect the reproduction of a black and white picture.

There are three types of troubles that can originate in one or both of these blocks:

1. No color reproduction
2. No color lock (synchronization)
3. Improper color rendition

Faults Resulting in No Color

Before looking for tube or component failures a check should be made to see that the set is properly adjusted to receive color and that the channel selected is transmitting a color program. The chroma control should be advanced and the fine tuning properly set. Also check the following:

1. Check operation of 3.58 MC Crystal Oscillator. A good check here is to measure the voltage at the suppressor grids of the Demodulator tubes. This will not only indicate whether the oscillator is functioning, but also whether or not the oscillator signal is reaching the Demodulator tubes. If the CW signals are driving the suppressor grids properly, the voltage should be approximately −5 volts. With no oscillation, fixed bias is the only voltage present—about −1.8 volts. If drive to the suppressor grids is lacking but grid voltage is present on the Crystal Oscillator, the oscillator may be greatly off frequency. Check frequency against a standard. Also check the 3.58 MC crystal, the transfer T124 and the Reactance Tube Circuit.

2. Check the Bandpass Amplifier tube and components up to the grids of the *I* and *Q* Demodulators. Although the Bandpass Amplifier is not in one of the blocks under discussion, troubles in this circuit can most definitely cause color faults not affecting black and white reception. No color reception in this circuit could be due to the Bandpass Amplifier tube, T126, or associated circuit components. It must be remembered that the Bandpass Amplifier tube can be driven to cut-off by a bias from the Color Killer tube; the killer tube will supply this cut-off bias if the grid bias to the Color Killer is lost due to the absence of burst.

No Color Lock (Sync)

Loss of color synchronization will show up as horizontal bands of color moving vertically.

If the color bands are very few in number, it indicates that the 3.58 MC oscillator is only slightly off frequency. Loss of color sync can be caused by:

1. Failure in the Color Phase Discriminator circuit
2. Failure in the Reactance Tube circuit
3. 3.58 MC oscillator slightly off frequency

IMPROPER COLOR RENDITION

Any fault that causes a change in the phase relationship between *I* and *Q* will produce improper colors:

1. Phase Control Misadjusted

Before trouble shooting for this fault it is advisable to check the phase control on the front panel of the receiver.

2. Quadrature Amplifier Phase Off

The Quadrature Amplifier is responsible for the 90 degree shift in phase of the *Q* CW signal to the *I* CW signal. Therefore, troubles here could be expected to cause improper color rendition.

3. *I* or *Q* Signal Missing

The *I* and *Q* channels pass a band of frequencies that are responsible for certain color components.

4. Quadrature Distortion

Narrow bands of incorrect color appearing at adjacent vertical edges of a sharp color transition is commonly called Quadrature Distortion. This occurs when the CW drives to the *I* and *Q* Demodulators are not 90 degrees displaced in phase.

Although malfunction of the Color Sync and Demodulation sections are most frequent causes of improper color rendition, this does not mean that color cannot suffer from faults in other parts of the receiver. Improper alignment of the RF and IF amplifiers can cause attenuation or loss of color information. This loss of high video frequencies, it might be added, could go unnoticed on a black and white transmission. Clipping in the picture IF or video sections could remove the burst reference, resulting in color sync trouble.

In a black and white transmission, it is well known that reflections can cause poor picture synchronization. In a color receiver, the *color sync* information may be cancelled by reflections and result in no *color* synchronization.

A fault that can occur in the Luminance Amplifier section that will affect color is the delay line. Approximately one microsecond of delay represents an appreciable misregistration of chrominance information with Y signal. This fault, of course, would not be apparent on a black and white transmission.

Faults in the power supply, such as hum in the raster, present no additional problems in a color receiver. However, filament-cathode leakage in tubes in the video amplifier channels will produce color hum bars. To aid in diagnosis it would be well for service technicians to know the predominant color *I* and *Q* channels. Hum appears as green and purple bars in the *Q* channel and as orange and cyan bars in the *I* channel.

Test Equipment for Color Servicing

. . . The following is the present thinking on test equipment requirements for the initial production RCA color television receivers:

1. OSCILLOSCOPE

Oscilloscopes that have good frequency and phase response from low frequency or from DC, up to 500 kilocycles, . . . are satisfactory for the majority of applications in servicing color receivers, including alignment, trouble-shooting in sync and deflection circuits, trouble-shooting in dynamic-convergence circuits, etc. For certain other applications, such as measurement of the 3.58 MC signals, a wide-band oscilloscope with flat response up to 4 MC, may be desirable.

The oscilloscope should have a compensated isolating probe to minimize loading effects, and it should have voltage calibration for the vertical amplifier in order to determine the amplitude of any waveform.

2. SWEEP GENERATOR

In addition the usual RF and IF ranges, the sweep generator should have a video-frequency range covering from approximately 50 kilocycles to 6 megacycles. The video range is required in checking the frequency response of the video amplifier, and in checking and adjusting the band-pass filter and the *I* and *Q* filters. Some brands and models of TV sweep generators do not include a video range, or do not cover video frequencies under 3 MC. . . . Other essential requirements in the sweep generator include flat voltage output and good linearity.

It is necessary to provide markers to identify specific frequencies on the sweep response curves of the video amplifier, the band-pass filter, and the *I* and *Q* filters. Absorption-type markers are satisfactory.

3. CALIBRATOR

The contour and frequency limits of the over-all RF-IF response curve are much more important in color receivers than in black and white receivers. For this reason it is essential to use an accurate crystal calibrated marker generator. . . .

4. VACUUM TUBE VOLTMETER

A vacuum tube voltmeter with high voltage probe will be necessary equipment for color servicing. The RCA VoltOhmyst with accessory probe is ideally suited for color purposes.

5. CONVERGENCE AND LINEARITY CHECKER

In order to permit proper observation of superimposition of three separate electron beams in the RCA Tri-Color Kinescope, a symmetrical pattern of light and dark patches, similar to a checkerboard, is desirable. Convergence can best be observed in one of two ways: by a pattern of equally spaced dots, or by a cross hatch pattern of horizontal and vertical lines. In either case, there must be a sharp transition between light and dark patches so that convergence may be observed horizontally, vertically and obliquely. Equal spacing between light and dark patches is desirable so that the convergence checker will also serve the dual function of a linearity checker. In the interests of servicing convenience, the convergence checker should be applied to a color receiver with a minimum of connections while maintaining reasonable initial cost.

6. COLOR BAR GENERATOR

[The study of] color signal development impresses the point that accurate phase relationships between the *I* signal, *Q* signal and burst must be maintained in order to produce proper color fidelity. Therefore, a signal comprised of static color bars representing this phase relationship will provide an accurate means of color phase adjustment. The color bar generator will also prove invaluable in adjusting the various gains and signal balance through the Color Matrix section, and in trouble-shooting the specialized color circuits. . . .

Alignment

VIDEO ALIGNMENT

1. Disconnect the [lead to] grid (pin 2) of the video amplifier (V114) and connect the sweep generator to pin 2.

2. Connect the scope to the cathode (pin 1) of V114.

3. Set the signal calibrator to 4.5 MC and adjust L110 for minimum response. Refer to Fig. 44A.

4. Remove the scope and connect to the cathode (pin 7) of the Second Video Amplifier (115A).

5. Adjust the contrast control so that a response is obtained on the scope.

6. Set the signal calibrator to 3.58 MC and adjust both cores of T114 for response as shown in Fig. 44B.

7. Connect the scope to each of the three kinescope grids successively and check for response as shown in Fig. 44C.

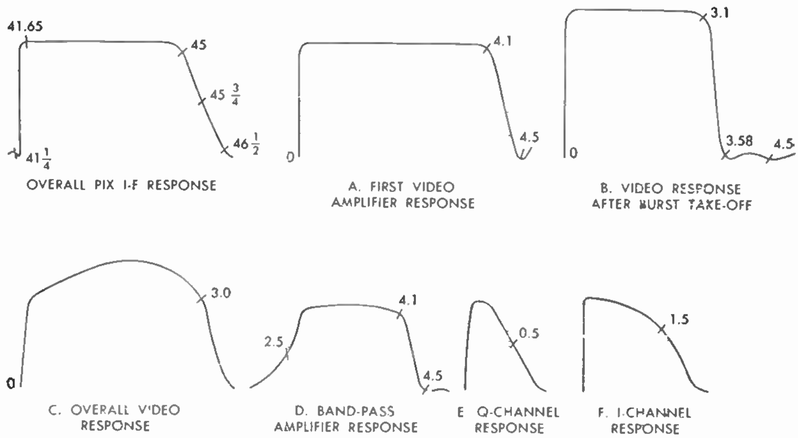


FIG. 44 Alignment Response Curves (Repeated)

8. Short out the resistor, *R296* in the Color Killer circuit.

9. Leave the sweep generator on the grid of the first Video Amplifier.

10. Connect the scope to high side of the chroma control.

11. Adjust T126 and L129 of bandpass filter until maximum gain and response [are] obtained as shown in Fig. 44D.

12. Remove the short on the grid of the Color Killer and connect the sweep generator to the arm of the chroma control.

13. Short the suppressor grids of the 6AS6 Demodulators to ground.

14. Connect the scope to the cathode (pin 8) of the "Q" Phase Splitter. A response curve should appear as shown in Fig. 44E.

15. Connect the scope to the junction of L133 and C280 in the *I* Amplifier, V134A. The response curve should appear as shown in Fig. 44F.

CHROMA SYNCHRONIZATION ALIGNMENT

1. Set the receiver fine tuning control to proper position for reception of a color bar pattern or other suitable color standard.

2. Short the grid (pin 2) of the Burst Amplifier (V129A) to ground.

3. Using a low capacity probe, connect the scope to "A" on T124.

4. Adjust the oscillator coil core for five volts peak to peak.

5. Connect the vacuum tube voltmeter on pin 8 of V129B.

6. Tune the primary of the Color Phasing Amplifier transformer (T123) for maximum indication. Tune the secondary for maximum indication while rocking the phase control (*R327*) about a mid-range position.

7. Remove the short on grid of the Burst amplifier.

8. Connect the scope to the cathode (pin 8) of the *Q* Phase Splitter.
9. Ground the junction of C239 and R291 in Reactance tube (V131A) grid circuit.
10. Adjust the core of L300 until the trace on the scope becomes stationary.
11. Remove the ground from the grid circuit of the Reactance Tube.
12. Reconnect the vacuum tube voltmeter to pin 8 on V129B.
13. Adjust the core of burst coil, T122 for maximum indication.
14. Adjust secondary of T114 for maximum indication.
15. Connect the vacuum tube voltmeter to the junction of C239 and R291.
16. Connect a 10 mmf capacitor across the 3.58 MC crystal to make the receiver fall out of color lock.
17. Adjust the AFC balance potentiometer for zero DC.
18. Remove the 10 mmf capacitor from the 3.58 MC crystal.
19. Connect the scope to junction of L133 and C280.
20. Observe the color bar trace on the scope and adjust the phase control to make the trace [normal].
21. Move the scope to the cathode (pin 8) of the *Q* Phase Splitter (V115B). Observe the trace and adjust the secondary of the Quadrature Amplifier transformer (T125) to obtain [proper] response.

I GAIN ADJUSTMENT

1. Adjust the fine tuning for proper reception of color bar signals or other color standard.
2. Connect the scope to the grid of the blue gun in kinescope.
3. Adjust the chroma and *I* gain controls to cancel all response excepting the blue and white bars.
4. Check red and green guns similarly for cancellation of undesirable colors.†

21. *Glossary of Terms Used in Color Television*

The following terms and definitions are included for quick reference in studying color television. Definitions of color terms suggested by the National Television Standards Committee (NTSC) are marked with asterisks.

Achromatic locus * (Achromatic region)—Chromaticities which may be acceptable reference standards under circumstances of common occurrence are represented in a chromaticity diagram by points in a region which may be called the "achromatic locus."

NOTE—The boundaries of the achromatic locus are indefinite, depending on the tolerances in any specific application. Acceptable reference standards of illumination (commonly referred to as "white light") are usually represented by points close to the locus of Planckian radiators having temperatures higher than about 2,000° K. While any point in the achromatic locus may be chosen as the reference point for the determination of dominant wavelength, complementary wavelength, and purity for specification of object colors, it is usually advisable to adopt the point representing the chromaticity of the luminator. . . . Having selected a suitable reference point, dominant wavelength may be determined

† See page 76 for source of material quoted, pages 76 to 86.

by noting the wavelength corresponding to the intersection of the spectrum locus with the straight line drawn from the reference point through the point representing the sample. When the reference point lies between the sample point and the intersection, the intersection indicates the complementary wavelength. Any point within the achromatic locus, chosen as a reference point, may be called an "achromatic point." Such points have also been called "white points."

Brightness *—The attribute of visual perception in accordance with which an area appears to emit more or less light.

NOTE—Luminance is recommended for the photometric quantity, which has been called "brightness." Luminance is a purely photometric quantity. Use of this name permits "brightness" to be used entirely with reference to the sensory response. The photometric quantity has been often confused with the sensation merely because of the use of one name for two distinct ideas. Brightness will continue to be used, properly, in nonquantitative statements, especially with reference to sensations and perceptions of light. Thus it is correct to refer to a brightness match, even in the field of a photometer, because the sensations are matched and only by inference are the photometric quantities (luminances) equal. Likewise, a photometer in which such matches are made will continue to be called an "equality-of-brightness" photometer.

A photo-electric instrument, calibrated in footlamberts, should not be called a "brightness meter." . . . A troublesome paradox is eliminated by the proposed distinction of nomenclature. The luminance of a surface may be doubled, yet it will be permissible to say that the brightness is not doubled, since the sensation which is called "brightness" is generally judged to be not doubled.

Bypass monochrome signal—A monochrome signal that is shunted around the color-sub-carrier modulator or demodulator.

Candle *—The unit of luminous intensity. One candle is defined as the luminous intensity of $\frac{1}{60}$th square centimeter of a blackbody radiator operating at the temperature of solidification of platinum. Values for standards having other spectral distributions are derived by the use of accepted luminosity factors.

Candlepower *—Luminous intensity expressed in candles.

Carrier color signal—The sidebands of the modulated color sub-carrier (plus the color sub-carrier, if not suppressed), which are added to the monochrome signal in order to convey color information.

Chroma * (Munsell chroma)—The dimension of the Munsell system of color which corresponds most closely to saturation.

NOTE—Chroma is frequently used, particularly in English works, as the equivalent of saturation.

Chromaticity *—The color quality of light definable by its chromaticity coordinates, or by its dominant (or complementary) wave length and its purity taken together.

Chromaticity co-ordinate *—The ratio of any one of the tristimulus values of a sample to the sum of the three tristimulus values.

Chromaticity diagram *—A plane diagram formed by plotting one of the three chromaticity co-ordinates against another.

NOTE—The most common chromaticity diagram at present is the CIE (International Commission of Illumination) (x, y) diagram plotted in rectangular co-ordinates.

Chrominance—The colorimetric difference between any color and a reference color of equal luminance, the reference color having a specified chromaticity.

Chrominance channel—In a color television system, any path which is intended to carry the carrier color signal ("carrier color signal" refers to signal carrier that carries color data).

Color *—The characteristics of light other than spatial and temporal inhomogeneities.

NOTE 1—The measure of color is three dimensional. One of the many ways of measuring color is in terms of luminance, dominant wavelength, and purity.

NOTE 2—Inhomogeneities, for example, particular distributions and variations of light, and characteristics of objects which are revealed by variations such as gloss, lustre, sheen, texture, sparkle, opalescence, and transparency, are not included among the color characteristics of objects.

Color burst—A few sine-wave cycles at color sub-carrier frequency (and the color-burst pedestal, if present), which is added to the "back porch" of the horizontal pedestal to permit the synchronizing of the color-carrier reference oscillator.

Color-burst pedestal—The rectangular, pulse-like component which may be part of the color burst. The amplitude of the color-burst pedestal is measured from the a-c axis of the sine-wave portion to the horizontal pedestal.

Color-carrier reference—A continuous signal having the same frequency as the color sub-carrier and having fixed phase with respect to the color burst. This signal is used for modulation at the transmitter and demodulation at the receiver.

Color co-ordinate transformation—Computation of the tri-stimulus values of colors in terms of one set of primaries from the tri-stimulus values of the same colors in another set of primaries.

NOTE—This computation may be performed electrically in a color television system.

Color-difference signal—An electrical signal which, when added to the monochrome signal, produces a signal representing one of the tri-stimulus values (with respect to a stated set of primaries) of the transmitted color.

Color edging—Spurious color at the boundaries of differently colored areas in the picture.

Color phase—(Of a given sub-carrier component). The phase, with respect to the color-carrier reference, of the component of the carrier color signal which transmits a particular color signal.

Color picture signal—The electrical signal which represents color picture information, consisting of a monochrome component plus a sub-carrier modulated with color information, excluding synchronizing signals.

Color sub-carrier—The carrier whose modulation sidebands are added to the monochrome signal in order to convey color information.

Color transmission—In television, the transmission of a signal which controls both the luminance values and the chromaticity values in a picture.

Compatibility—The nature of a color television system which permits substantially normal monochrome reception of the signal transmission by typical, unaltered monochrome receivers.

Complementary wavelength *—The wavelength of light of a single frequency, which matches the reference standard light when combined with a sample color in suitable proportions.

NOTE 1—The wide variety of purples which have no dominant wavelengths, including nonspectral violet, purple, magenta, and nonspectral red colors, are specified by use of their complementary wavelengths.

NOTE 2—Refer to Dominant Wavelength.

Composite color signal—The color picture, including blanking and all synchronizing signals.

Constant-luminance transmission—A method of color transmission in which the carrier color signal controls the chromaticity of the produced image without affecting the luminance, the luminance being controlled by the monochrome signal.

Delay distortion—That form of distortion which occurs when the envelope delay of a circuit or system is not constant over the frequency range required for transmission.

Dominant wavelength *—The wavelength of light of a single frequency, which matches a color when combined in suitable proportions with a reference standard light.

NOTE—Light of a single frequency is approximated in practice by the use of a range of wavelengths within which there is no noticeable difference of color. Although this practice is ambiguous in principle, the dominant wavelength is usually taken as the average wavelength of the band used in the mixture with the reference standard matching the sample. Many different qualities of light are used as reference standards under various circumstances. Usually the quality of the prevailing illumination is acceptable as the reference standard in the determination of the dominant wavelength of the colors of objects.

Envelope delay—The first derivative of the phase shift with reference to the frequency.

NOTE—If the phase is measured in radians and the frequency in radians per second, the envelope delay will be obtained in terms of seconds.

Equal-energy source *—A light source for which the time rate of emission of energy per unit of wavelength is constant throughout the visible spectrum.

Excitation purity (purity) *—The ratio of the distance from the reference point to the point representing the sample, to the distance along the same straight line from the reference point to the spectrum locus or to the purple boundary, both distances being measured (in the same direction from the reference point) on the CIE chromaticity diagram.

NOTE—The reference point is the point in the chromaticity diagram which represents the reference standard light mentioned in the definition of Dominant Wavelength.

Field—One of the two (or more) equal parts into which a frame is divided in interlaced scanning.

Footcandle *—A unit of illuminance when the foot is taken as the unit of length. It is the illuminance on a surface one square foot in area on which there is a uniformly distributed flux of one lumen, or the illuminance at a surface all points of which are at a distance of one foot from a uniform source of one candle.

Footlambert *—A unit of luminance equal to $1/\pi$ candle per square foot, or to the uniform luminance of a perfectly diffusing surface emitting or reflecting light at the rate of one lumen per square foot.

NOTE—A footcandle is a unit of incident light and a footlambert is a unit of emitted or reflected light. For a perfectly reflecting and perfectly diffusing surface, the number of footcandles is equal to the number of footlamberts.

Frequency overlap—In a color television system, that part of the frequency band which is common to monochrome and chrominance channels.

Gamma—In a color or monochrome channel, or part thereof, the coefficient expressing the selected evaluation of the slope of the used part of the log-versus-log plot relating input (abscissa) and output (ordinate) signal magnitudes, as measured from the point corresponding to some black level of reference.

Gamma correction—The modification of a transfer characteristic for the purpose of changing the value of gamma.

Hue *—The attribute of color perception that determines whether it is red, yellow, green, blue, purple, or the like.

NOTE 1—This is a subjective term corresponding to the psychophysical term Dominant (or Complementary) Wavelength.

NOTE 2—White, black, and gray are not considered as being hues.

Illuminance * (Illumination).—The density of the luminous flux on a surface; it is the quotient of the flux by the area of the surface when the latter is uniformly illuminated.

Lambert *—A unit of luminance equal to $1/\pi$ candle per square centimeter and, therefore, equal to the uniform luminance of a perfectly diffusing surface emitting or reflecting light at the rate of one lumen per square centimeter.

Light *—The aspect of radiant energy of which a human observer is aware through the visual sensations that arise from the stimulation of the retina of the eye. For the purposes of engineering, light is visually evaluated radiant energy.

NOTE 1—Light is psychophysical, neither purely physical nor purely psychological. Light is not synonymous with radiant energy, however restricted, nor is it merely sensation.

NOTE 2—The present basis for the engineering evaluation of light consists of the color-mixture data: $\bar{x}$, $\bar{y}$, $\bar{z}$, adopted in 1931 by the International Commission on Illumination.

Lumen *—The unit of luminous flux. It is equal to the flux through a unit solid angle (steradian) from a uniform point source of one candle, or to the flux on a unit surface all points of which are at unit distance from a uniform point source of one candle.

Luminance *—The luminous intensity of any surface in a given direction per unit of projected area of the surface as viewed from that direction.

NOTE—See Note under the term Brightness.

Luminance channel—In a color television system, any path which is intended to carry the luminance signal.

NOTE—The luminance channel may also carry other signals, for example, the carrier color signal (which may or may not be present).

Luminosity *—Ratio of luminous flux to the corresponding radiant flux at a particular wavelength. It is expressed in lumens per watt.

Luminosity coefficients *—The constant multipliers for the respective tristimulus values of any color, such that the sum of the three products is the luminance of the color.

Luminous efficiency *—The ratio of the luminous flux to the radiant flux.

NOTE—Luminous efficiency is usually expressed in lumens per watt of radiant flux. It should not be confused with the term "efficiency" as applied to a practical source of light, since the latter is based upon the power supplied to the source instead of the radiant flux from the source. For energy radiated at a single wavelength, luminous efficiency is synonymous with luminosity.

Luminous flux *—The time rate of flow of light.

Luminous intensity * (In any direction)—The ratio of the luminous flux emitted by a source or by an element of a source, in an infinitesimal solid angle, containing this direction, to the solid angle.

NOTE—Mathematically, a solid angle must have a point at its apex; the definition of Luminous Intensity, therefore, applies strictly only to a point source. In practice, however, light emanating from a source whose dimensions are negligible in comparison with the distance from which it is observed may be considered as coming from a point.

Matrix A. (Noun)—In color television, an array of coefficients symbolic of an operation to be performed, which operation results in a color-co-ordinate transformation. (This definition is consistent with mathematical usage.)

Matrix B. (Verb)—In color television, to perform a color-co-ordinate transformation by computation or by electrical, optical, or other means.

Matrixer (Matrix Unit, Matrix Circuit, etc.)—A device which performs a color-co-ordinate transformation by electrical, optical, or other means.

Moiré—In television the spurious pattern in the reproduced picture resulting from interference beats between two sets of periodic structures in the image.

Monochrome—Black and white. ("Monochrome" is the preferred term.)

Monochrome bandwidth (Of the signal)—The video bandwidth of the monochrome signal.

Monochrome bandwidth (Of the monochrome channel)—The video bandwidth of the monochrome channel.

Monochrome channel—In color television transmission, any path which is intended to carry the monochrome signal.

Monochrome signal—A. In monochrome television transmission, a signal wave for controlling the luminance values in the picture but not the chromaticity values.

B. In color television transmission, that part of the signal which has major control of the luminance of the color picture on a conventional monochrome receiver.

Monochrome transmission—In television, the transmission of a signal used for controlling the luminance values in the picture but not the chromaticity values.

Pickup spectral characteristic—The set of spectral responses of the device (camera), including the optical parts, which converts radiation into electrical signals, prior to any nonlinearizing and matrixing operations.

Primaries *—The colors of constant chromaticity and variable luminance, which, when mixed in proper proportions, are used to produce or specify other colors.

NOTE—Primaries need not be physically realizable.

Purity (Excitation Purity) *—*See* **Excitation purity.**

Purple boundary *—The straight line drawn between the ends of the spectrum locus.

Radiance*—The radiant flux per unit solid angle per unit of projected area of the source.

NOTE—The usual unit is the watt per steradian per square meter. This is the radiant analog of luminance.

Radiant flux *—The time rate of flow of radiant energy.

Radiant intensity *—The energy emitted per unit time, per unit solid angle about the direction considered; for example, watts per steradian.

Receiver primaries *—The colors of constant chromaticity and variable luminance produced by the receiver which, when mixed in proper proportions, are used to produce other colors.

NOTE—Usually three primaries are used: red, green, and blue.

Relative luminosity *—The ratio of the value of the luminosity at a particular wavelength to the value at the wavelength of maximum luminosity.

Saturation *—The attribute of any color perception possessing a hue that determines the degree of its difference from the achromatic color perception most resembling it.

NOTE 1—This is a subjective term corresponding to the psychophysical term Purity.

NOTE 2—The description of saturation is not commonly undertaken beyond the use of rather vague terms, such as vivid, strong, and weak. The terms brilliant, pastel, pale, and deep, which are sometimes used as descriptive of saturation, have connotations descriptive also of brightness.

Spectrum locus *—The locus of points representing the chromaticities of spectrally pure stimuli in a chromaticity diagram.

Tristimulus value—A tristimulus value is a numerical specification of a color in terms of three primaries.

White *

NOTE—In color television, the term White is used most commonly in the nontechnical sense. More specific usage is covered by the term Achromatic Locus, and this usage is explained in the Note under the term Achromatic Locus.

White object *—An object which reflects all wavelengths of light with substantially equal high efficiencies and with considerable diffusion.

Zero-sub-carrier chromaticity—The chromaticity which is intended to be displayed when the sub-carrier amplitude is zero.

LOGARITHMIC TABLES

TABLE I—BASE *e* TO THE MINUS POWER

x	e^{-x} Value	x	e^{-x} Value	x	e^{-x} Value	x	e^{-x} Value
0.00	1.0000						
0.01	.99005	0.51	.60050	1.01	.36422	1.51	.22091
0.02	.98020	0.52	.59452	1.02	.36060	1.52	.21871
0.03	.97045	0.53	.58860	1.03	.35701	1.53	.21654
0.04	.96079	0.54	.58275	1.04	.35345	1.54	.21438
0.05	.95123	0.55	.57695	1.05	.34994	1.55	.21225
0.06	.94176	0.56	.57121	1.06	.34646	1.56	.21014
0.07	.93239	0.57	.56553	1.07	.34301	1.57	.20805
0.08	.92312	0.58	.55990	1.08	.33960	1.58	.20598
0.09	.91393	0.59	.55433	1.09	.33622	1.59	.20393
0.10	.90484	**0.60**	.54881	**1.10**	.33287	**1.60**	.20190
0.11	.89583	0.61	.54335	1.11	.32956	1.61	.19989
0.12	.88692	0.62	.53794	1.12	.32628	1.62	.19790
0.13	.87810	0.63	.53259	1.13	.32303	1.63	.19593
0.14	.86936	0.64	.52729	1.14	.31982	1.64	.19398
0.15	.86071	0.65	.52205	1.15	.31664	1.65	.19205
0.16	.85214	0.66	.51685	1.16	.31349	1.66	.19014
0.17	.84366	0.67	.51171	1.17	.31037	1.67	.18825
0.18	.83527	0.68	.50662	1.18	.30728	1.68	.18637
0.19	.82696	0.69	.50158	1.19	.30422	1.69	.18452
0.20	.81873	**0.70**	.49659	**1.20**	.30119	**1.70**	.18268
0.21	.81058	0.71	.49164	1.21	.29820	1.71	.18087
0.22	.80252	0.72	.48675	1.22	.29523	1.72	.17907
0.23	.79453	0.73	.48191	1.23	.29229	1.73	.17728
0.24	.78663	0.74	.47711	1.24	.28938	1.74	.17552
0.25	.77880	0.75	.47237	1.25	.28650	1.75	.17377
0.26	.77105	0.76	.46767	1.26	.28365	1.76	.17204
0.27	.76338	0.77	.46301	1.27	.28083	1.77	.17033
0.28	.75578	0.78	.45841	1.28	.27804	1.78	.16864
0.29	.74826	0.79	.45384	1.29	.27527	1.79	.16696
0.30	.74082	**0.80**	.44933	**1.30**	.27253	**1.80**	.16530
0.31	.73345	0.81	.44486	1.31	.26982	1.81	.16365
0.32	.72615	0.82	.44043	1.32	.26714	1.82	.16203
0.33	.71892	0.83	.43605	1.33	.26448	1.83	.16041
0.34	.71177	0.84	.43171	1.34	.26185	1.84	.15882
0.35	.70469	0.85	.42741	1.35	.25924	1.85	.15724
0.36	.69768	0.86	.42316	1.36	.25666	1.86	.15567
0.37	.69073	0.87	.41895	1.37	.25411	1.87	.15412
0.38	.68386	0.88	.41478	1.38	.25158	1.88	.15259
0.39	.67706	0.89	.41066	1.39	.24908	1.89	.15107
0.40	.67032	**0.90**	.40657	**1.40**	.24660	**1.90**	.14957
0.41	.66365	0.91	.40252	1.41	.24414	1.91	.14808
0.42	.65705	0.92	.39852	1.42	.24171	1.92	.14661
0.43	.65051	0.93	.39455	1.43	.23931	1.93	.14515
0.44	.64404	0.94	.39063	1.44	.23693	1.94	.14370
0.45	.63763	0.95	.38674	1.45	.23457	1.95	.14227
0.46	.63128	0.96	.38289	1.46	.23224	1.96	.14086
0.47	.62500	0.97	.37908	1.47	.22993	1.97	.13946
0.48	.61878	0.98	.37531	1.48	.22764	1.98	.13807
0.49	.61263	0.99	.37158	1.49	.22537	1.99	.13670
0.50	.60653	**1.00**	.36788	**1.50**	.22313	**2.00**	.13534

x	e^{-x} Value	x	e^{-x} Value	x	e^{-x} Value
2.00	.13534				
2.01	.13399	2.51	.08127	3.05	.04736
2.02	.13266	2.52	.08046	3.10	.04505
2.03	.13134	2.53	.07966	3.15	.04285
2.04	.13003	2.54	.07887	3.20	.04076
2.05	.12873	2.55	.07808	3.25	.03877
2.06	.12745	2.56	.07730	3.30	.03688
2.07	.12619	2.57	.07654	3.35	.03508
2.08	.12493	2.58	.07577	3.40	.03337
2.09	.12369	2.59	.07502	3.45	.03175
2.10	.12246	**2.60**	.07427	**3.50**	.03020
2.11	.12124	2.61	.07353	3.55	.02872
2.12	.12003	2.62	.07280	3.60	.02732
2.13	.11884	2.63	.07208	3.65	.02599
2.14	.11765	2.64	.07136	3.70	.02472
2.15	.11648	2.65	.07065	3.75	.02352
2.16	.11533	2.66	.06995	3.80	.02237
2.17	.11418	2.67	.06925	3.85	.02128
2.18	.11304	2.68	.06856	3.90	.02024
2.19	.11192	2.69	.06788	3.95	.01925
2.20	.11080	**2.70**	.06721	**4.00**	.01832
2.21	.10970	2.71	.06654	4.10	.01657
2.22	.10861	2.72	.06587	4.20	.01500
2.23	.10753	2.73	.06522	4.30	.01357
2.24	.10646	2.74	.06457	4.40	.01228
2.25	.10540	2.75	.06393	4.50	.01111
2.26	.10435	2.76	.06329	4.60	.01005
2.27	.10331	2.77	.06266	4.70	.00910
2.28	.10228	2.78	.06204	4.80	.00823
2.29	.10127	2.79	.06142	4.90	.00745
2.30	.10026	**2.80**	.06081	**5.00**	.00674
2.31	.09926	2.81	.06020	5.10	.00610
2.32	.09827	2.82	.05961	5.20	.00552
2.33	.09730	2.83	.05901	5.30	.00499
2.34	.09633	2.84	.05843	5.40	.00452
2.35	.09537	2.85	.05784	5.50	.00409
2.36	.09442	2.86	.05727	5.60	.00370
2.37	.09348	2.87	.05670	5.70	.00335
2.38	.09255	2.88	.05613	5.80	.00303
2.39	.09163	2.89	.05558	5.90	.00274
2.40	.09072	**2.90**	.05502	**6.00**	.00248
2.41	.08982	2.91	.05448	6.25	.00193
2.42	.08892	2.92	.05393	6.50	.00150
2.43	.08804	2.93	.05340	6.75	.00117
2.44	.08716	2.94	.05287	7.00	.00091
2.45	.08629	2.95	.05234	7.50	.00055
2.46	.08543	2.96	.05182	8.00	.00034
2.47	.08458	2.97	.05130	8.50	.00020
2.48	.08374	2.98	.05079	9.00	.00012
2.49	.08291	2.99	.05029	9.50	.00007
2.50	.08208	**3.00**	.04979	**10.00**	.00005

TABLE IIA—FOUR-PLACE LOGARITHMS

N	0	1	2	3	4	5	6	7	8	9	1	2	3	4	5	6	7	8	9
10	0000	0043	0086	0128	0170	0212	0253	0294	0334	0374	4	8	12	17	21	25	29	33	37
11	0414	0453	0492	0531	0569	0607	0645	0682	0719	0755	4	8	11	15	19	23	26	30	34
12	0792	0828	0864	0899	0934	0969	1004	1038	1072	1106	3	7	10	14	17	21	24	28	31
13	1139	1173	1206	1239	1271	1303	1335	1367	1399	1430	3	6	10	13	16	19	23	26	29
14	1461	1492	1523	1553	1584	1614	1644	1673	1703	1732	3	6	9	12	15	18	21	24	27
15	1761	1790	1818	1847	1875	1903	1931	1959	1987	2014	3	6	8	11	14	17	20	22	25
16	2041	2068	2095	2122	2148	2175	2201	2227	2253	2279	3	5	8	11	13	16	18	21	24
17	2304	2330	2355	2380	2405	2430	2455	2480	2504	2529	2	5	7	10	12	15	17	20	22
18	2553	2577	2601	2625	2648	2672	2695	2718	2742	2765	2	5	7	9	12	14	16	19	21
19	2788	2810	2833	2856	2878	2900	2923	2945	2967	2989	2	4	7	9	11	13	16	18	20
20	3010	3032	3054	3075	3096	3118	3139	3160	3181	3201	2	4	6	8	11	13	15	17	19
21	3222	3243	3263	3284	3304	3324	3345	3365	3385	3404	2	4	6	8	10	12	14	16	18
22	3424	3444	3464	3483	3502	3522	3541	3560	3579	3598	2	4	6	8	10	12	14	16	17
23	3617	3636	3655	3674	3692	3711	3729	3747	3766	3784	2	4	6	7	9	11	13	15	17
24	3802	3820	3838	3856	3874	3892	3909	3927	3945	3962	2	4	5	7	9	11	12	14	16
25	3979	3997	4014	4031	4048	4065	4082	4099	4116	4133	2	4	5	7	9	10	12	14	16
26	4150	4166	4183	4200	4216	4232	4249	4265	4281	4298	2	3	5	7	8	10	11	13	15
27	4314	4330	4346	4362	4378	4393	4409	4425	4440	4456	2	3	5	6	8	9	11	12	14
28	4472	4487	4502	4518	4533	4548	4564	4579	4594	4609	2	3	5	6	8	9	11	12	14
29	4624	4639	4654	4669	4683	4698	4713	4728	4742	4757	1	3	4	6	7	9	10	12	13
30	4771	4786	4800	4814	4829	4843	4857	4871	4886	4900	1	3	4	6	7	9	10	11	13
31	4914	4928	4942	4955	4969	4983	4997	5011	5024	5038	1	3	4	5	7	8	10	11	12
32	5051	5065	5079	5092	5105	5119	5132	5145	5159	5172	1	3	4	5	7	8	9	11	12
33	5185	5198	5211	5224	5237	5250	5263	5276	5289	5302	1	3	4	5	7	8	9	11	12
34	5315	5328	5340	5353	5366	5378	5391	5403	5416	5428	1	2	4	5	6	8	9	10	11
35	5441	5453	5465	5478	5490	5502	5514	5527	5539	5551	1	2	4	5	6	7	9	10	11
36	5563	5575	5587	5599	5611	5623	5635	5647	5658	5670	1	2	4	5	6	7	8	10	11
37	5682	5694	5705	5717	5729	5740	5752	5763	5775	5786	1	2	4	5	6	7	8	9	11
38	5798	5809	5821	5832	5843	5855	5866	5877	5888	5899	1	2	3	5	6	7	8	9	10
39	5911	5922	5933	5944	5955	5966	5977	5988	5999	6010	1	2	3	4	5	7	8	9	10
40	6021	6031	6042	6053	6064	6075	6085	6096	6107	6117	1	2	3	4	5	6	8	9	10
41	6128	6138	6149	6160	6170	6180	6191	6201	6212	6222	1	2	3	4	5	6	7	8	9
42	6232	6243	6253	6263	6274	6284	6294	6304	6314	6325	1	2	3	4	5	6	7	8	9
43	6335	6345	6355	6365	6375	6385	6395	6405	6415	6425	1	2	3	4	5	6	7	8	9
44	6435	6444	6454	6464	6474	6484	6493	6503	6513	6522	1	2	3	4	5	6	7	8	9
45	6532	6542	6551	6561	6571	6580	6590	6599	6609	6618	1	2	3	4	5	6	7	8	9
46	6628	6637	6646	6656	6665	6675	6684	6693	6702	6712	1	2	3	4	5	6	7	7	8
47	6721	6730	6739	6749	6758	6767	6776	6785	6794	6803	1	2	3	4	5	6	7	7	8
48	6812	6821	6830	6839	6848	6857	6866	6875	6884	6893	1	2	3	4	5	6	7	7	8
49	6902	6911	6920	6928	6937	6946	6955	6964	6972	6981	1	2	3	4	4	5	6	7	8
50	6990	6998	7007	7016	7024	7033	7042	7050	7059	7067	1	2	3	3	4	5	6	7	8
51	7076	7084	7093	7101	7110	7118	7126	7135	7143	7152	1	2	3	3	4	5	6	7	8
52	7160	7168	7177	7185	7193	7202	7210	7218	7226	7235	1	2	3	3	4	5	6	7	7
53	7243	7251	7259	7267	7275	7284	7292	7300	7308	7316	1	2	2	3	4	5	6	6	7
54	7324	7332	7340	7348	7356	7364	7372	7380	7388	7396	1	2	2	3	4	5	6	6	7
N	0	1	2	3	4	5	6	7	8	9	1	2	3	4	5	6	7	8	9

The proportional parts are stated in full for every tenth at the right-hand side. The logarithm of any number of four significant figures can be read directly by add-

N	0	1	2	3	4	5	6	7	8	9	1 2 3	4 5 6	7 8 9
55	7404	7412	7419	7427	7435	7443	7451	7459	7466	7474	1 2 2	3 4 5	5 6 7
56	7482	7490	7497	7505	7513	7520	7528	7536	7543	7551	1 2 2	3 4 5	5 6 7
57	7559	7566	7574	7582	7589	7597	7604	7612	7619	7627	1 1 2	3 4 5	5 6 7
58	7634	7642	7649	7657	7664	7672	7679	7686	7694	7701	1 1 2	3 4 4	5 6 7
59	7709	7716	7723	7731	7738	7745	7752	7760	7767	7774	1 1 2	3 4 4	5 6 7
60	7782	7789	7796	7803	7810	7818	7825	7832	7839	7846	1 1 2	3 4 4	5 6 6
61	7853	7860	7868	7875	7882	7889	7896	7903	7910	7917	1 1 2	3 3 4	5 6 6
62	7924	7931	7938	7945	7952	7959	7966	7973	7980	7987	1 1 2	3 3 4	5 5 6
63	7993	8000	8007	8014	8021	8028	8035	8041	8048	8055	1 1 2	3 3 4	5 5 6
64	8062	8069	8075	8082	8089	8096	8102	8109	8116	8122	1 1 2	3 3 4	5 5 6
65	8129	8136	8142	8149	8156	8162	8169	8176	8182	8189	1 1 2	3 3 4	5 5 6
66	8195	8202	8209	8215	8222	8228	8235	8241	8248	8254	1 1 2	3 3 4	5 5 6
67	8261	8267	8274	8280	8287	8293	8299	8306	8312	8319	1 1 2	3 3 4	5 5 6
68	8325	8331	8338	8344	8351	8357	8363	8370	8376	8382	1 1 2	3 3 4	4 5 6
69	8388	8395	8401	8407	8414	8420	8426	8432	8439	8445	1 1 2	3 3 4	4 5 6
70	8451	8457	8463	8470	8476	8482	8488	8494	8500	8506	1 1 2	3 3 4	4 5 6
71	8513	8519	8525	8531	8537	8543	8549	8555	8561	8567	1 1 2	3 3 4	4 5 6
72	8573	8579	8585	8591	8597	8603	8609	8615	8621	8627	1 1 2	3 3 4	4 5 6
73	8633	8639	8645	8651	8657	8663	8669	8675	8681	8686	1 1 2	2 3 4	4 5 5
74	8692	8698	8704	8710	8716	8722	8727	8733	8739	8745	1 1 2	2 3 4	4 5 5
75	8751	8756	8762	8768	8774	8779	8785	8791	8797	8802	1 1 2	2 3 3	4 5 5
76	8808	8814	8820	8825	8831	8837	8842	8848	8854	8859	1 1 2	2 3 3	4 4 5
77	8865	8871	8876	8882	8887	8893	8899	8904	8910	8915	1 1 2	2 3 3	4 4 5
78	8921	8927	8932	8938	8943	8949	8954	8960	8965	8971	1 1 2	2 3 3	4 4 5
79	8976	8982	8987	8993	8998	9004	9009	9015	9020	9025	1 1 2	2 3 3	4 4 5
80	9031	9036	9042	9047	9053	9058	9063	9069	9074	9079	1 1 2	2 3 3	4 4 5
81	9085	9090	9096	9101	9106	9112	9117	9122	9128	9133	1 1 2	2 3 3	4 4 5
82	9138	9143	9149	9154	9159	9165	9170	9175	9180	9186	1 1 2	2 3 3	4 4 5
83	9191	9196	9201	9206	9212	9217	9222	9227	9232	9238	1 1 2	2 3 3	4 4 5
84	9243	9248	9253	9258	9263	9269	9274	9279	9284	9289	1 1 2	2 3 3	4 4 5
85	9294	9299	9304	9309	9315	9320	9325	9330	9335	9340	1 1 2	2 3 3	4 4 5
86	9345	9350	9355	9360	9365	9370	9375	9380	9385	9390	1 1 2	2 3 3	4 4 5
87	9395	9400	9405	9410	9415	9420	9425	9430	9435	9440	1 1 2	2 3 3	4 4 5
88	9445	9450	9455	9460	9465	9469	9474	9479	9484	9489	0 1 1	2 2 3	3 4 4
89	9494	9499	9504	9509	9513	9518	9523	9528	9533	9538	0 1 1	2 2 3	3 4 4
90	9542	9547	9552	9557	9562	9566	9571	9576	9581	9586	0 1 1	2 2 3	3 4 4
91	9590	9595	9600	9605	9609	9614	9619	9624	9628	9633	0 1 1	2 2 3	3 4 4
92	9638	9643	9647	9652	9657	9661	9666	9671	9675	9680	0 1 1	2 2 3	3 4 4
93	9685	9689	9694	9699	9703	9708	9713	9717	9722	9727	0 1 1	2 2 3	3 4 4
94	9731	9736	9741	9745	9750	9754	9759	9763	9768	9773	0 1 1	2 2 3	3 4 4
95	9777	9782	9786	9791	9795	9800	9805	9809	9814	9818	0 1 1	2 2 3	3 4 4
96	9823	9827	9832	9836	9841	9845	9850	9854	9859	9863	0 1 1	2 2 3	3 4 4
97	9868	9872	9877	9881	9886	9890	9894	9899	9903	9908	0 1 1	2 2 3	3 4 4
98	9912	9917	9921	9926	9930	9934	9939	9943	9948	9952	0 1 1	2 2 3	3 3 4
99	9956	9961	9965	9969	9974	9978	9983	9987	9991	9996	0 1 1	2 2 3	3 3 4
N	0	1	2	3	4	5	6	7	8	9	1 2 3	4 5 6	7 8 9

ing the proportional part corresponding to the fourth figure to the tabular number corresponding to the first three figures. There may be an error of 1 in the last place.

TABLE IIB—ANTILOGARITHMS TO FOUR PLACES

	0	1	2	3	4	5	6	7	8	9	1	2	3	4	5	6	7	8	9
.00	1000	1002	1005	1007	1009	1012	1014	1016	1019	1021	0	0	1	1	1	1	2	2	2
.01	1023	1026	1028	1030	1033	1035	1038	1040	1042	1045	0	0	1	1	1	1	2	2	2
.02	1047	1050	1052	1054	1057	1059	1062	1064	1067	1069	0	0	1	1	1	1	2	2	2
.03	1072	1074	1076	1079	1081	1084	1086	1089	1091	1094	0	0	1	1	1	1	2	2	2
.04	1096	1099	1102	1104	1107	1109	1112	1114	1117	1119	0	1	1	1	1	2	2	2	2
.05	1122	1125	1127	1130	1132	1135	1138	1140	1143	1146	0	1	1	1	1	2	2	2	2
.06	1148	1151	1153	1156	1159	1161	1164	1167	1169	1172	0	1	1	1	1	2	2	2	2
.07	1175	1178	1180	1183	1186	1189	1191	1194	1197	1199	0	1	1	1	1	2	2	2	2
.08	1202	1205	1208	1211	1213	1216	1219	1222	1225	1227	0	1	1	1	1	2	2	2	3
.09	1230	1233	1236	1239	1242	1245	1247	1250	1253	1256	0	1	1	1	1	2	2	2	3
.10	1259	1262	1265	1268	1271	1274	1276	1279	1282	1285	0	1	1	1	1	2	2	2	3
.11	1288	1291	1294	1297	1300	1303	1306	1309	1312	1315	0	1	1	1	2	2	2	2	3
.12	1318	1321	1324	1327	1330	1334	1337	1340	1343	1346	0	1	1	1	2	2	2	2	3
.13	1349	1352	1355	1358	1361	1365	1368	1371	1374	1377	0	1	1	1	2	2	2	3	3
.14	1380	1384	1387	1390	1393	1396	1400	1403	1406	1409	0	1	1	1	2	2	2	3	3
.15	1413	1416	1419	1422	1426	1429	1432	1435	1439	1442	0	1	1	1	2	2	2	3	3
.16	1445	1449	1452	1455	1459	1462	1466	1469	1472	1476	0	1	1	1	2	2	2	3	3
.17	1479	1483	1486	1489	1493	1496	1500	1503	1507	1510	0	1	1	1	2	2	2	3	3
.18	1514	1517	1521	1524	1528	1531	1535	1538	1542	1545	0	1	1	1	2	2	2	3	3
.19	1549	1552	1556	1560	1563	1567	1570	1574	1578	1581	0	1	1	1	2	2	2	3	3
.20	1585	1589	1592	1596	1600	1603	1607	1611	1614	1618	0	1	1	1	2	2	3	3	3
.21	1622	1626	1629	1633	1637	1641	1644	1648	1652	1656	0	1	1	1	2	2	3	3	3
.22	1660	1663	1667	1671	1675	1679	1683	1687	1690	1694	0	1	1	2	2	2	3	3	3
.23	1698	1702	1706	1710	1714	1718	1722	1726	1730	1734	0	1	1	2	2	2	3	3	3
.24	1738	1742	1746	1750	1754	1758	1762	1766	1770	1774	0	1	1	2	2	2	3	3	4
.25	1778	1782	1786	1791	1795	1799	1803	1807	1811	1816	0	1	1	2	2	3	3	3	4
.26	1820	1824	1828	1832	1837	1841	1845	1849	1854	1858	0	1	1	2	2	3	3	3	4
.27	1862	1866	1871	1875	1879	1884	1888	1892	1897	1901	0	1	1	2	2	3	3	3	4
.28	1905	1910	1914	1919	1923	1928	1932	1936	1941	1945	0	1	1	2	2	3	3	4	4
.29	1950	1954	1959	1963	1968	1972	1977	1982	1986	1991	0	1	1	2	2	3	3	4	4
.30	1995	2000	2004	2009	2014	2018	2023	2028	2032	2037	0	1	1	2	2	3	3	4	4
.31	2042	2046	2051	2056	2061	2065	2070	2075	2080	2084	0	1	1	2	2	3	3	4	4
.32	2089	2094	2099	2104	2109	2113	2118	2123	2128	2133	0	1	1	2	2	3	3	4	4
.33	2138	2143	2148	2153	2158	2163	2168	2173	2178	2183	0	1	1	2	2	3	3	4	4
.34	2188	2193	2198	2203	2208	2213	2218	2223	2228	2234	1	1	2	2	3	3	4	4	5
.35	2239	2244	2249	2254	2259	2265	2270	2275	2280	2286	1	1	2	2	3	3	4	4	5
.36	2291	2296	2301	2307	2312	2317	2323	2328	2333	2339	1	1	2	2	3	3	4	4	5
.37	2344	2350	2355	2360	2366	2371	2377	2382	2388	2393	1	1	2	2	3	3	4	4	5
.38	2399	2404	2410	2415	2421	2427	2432	2438	2443	2449	1	1	2	2	3	3	4	5	5
.39	2455	2460	2466	2472	2477	2483	2489	2495	2500	2506	1	1	2	2	3	3	4	5	5
.40	2512	2518	2523	2529	2535	2541	2547	2553	2559	2564	1	1	2	2	3	4	4	5	5
.41	2570	2576	2582	2588	2594	2600	2606	2612	2618	2624	1	1	2	2	3	4	4	5	6
.42	2630	2636	2642	2649	2655	2661	2667	2673	2679	2685	1	1	2	2	3	4	4	5	6
.43	2692	2698	2704	2710	2716	2723	2729	2735	2742	2748	1	1	2	2	3	4	4	5	6
.44	2754	2761	2767	2773	2780	2786	2793	2799	2805	2812	1	1	2	3	3	4	4	5	6
.45	2818	2825	2831	2838	2844	2851	2858	2864	2871	2877	1	1	2	3	3	4	5	5	6
.46	2884	2891	2897	2904	2911	2917	2924	2931	2938	2944	1	1	2	3	3	4	5	5	6
.47	2951	2958	2965	2972	2979	2985	2992	2999	3006	3013	1	1	2	3	3	4	5	6	6
.48	3020	3027	3034	3041	3048	3055	3062	3069	3076	3083	1	1	2	3	3	4	5	6	6
.49	3090	3097	3105	3112	3119	3126	3133	3141	3148	3155	1	1	2	3	4	4	5	6	6

TABLE IIB—ANTILOGARITHMS TO FOUR PLACES

	0	1	2	3	4	5	6	7	8	9	1	2	3	4	5	6	7	8	9
.50	3162	3170	3177	3184	3192	3199	3206	3214	3221	3228	1	1	2	3	4	4	5	6	7
.51	3236	3243	3251	3258	3266	3273	3281	3289	3296	3304	1	1	2	3	4	4	5	6	7
.52	3311	3319	3327	3334	3342	3350	3357	3365	3373	3381	1	1	2	3	4	5	5	6	7
.53	3388	3396	3404	3412	3420	3428	3436	3443	3451	3459	1	2	2	3	4	5	6	6	7
.54	3467	3475	3483	3491	3499	3508	3516	3524	3532	3540	1	2	2	3	4	5	6	6	7
.55	3548	3556	3565	3573	3581	3589	3597	3606	3614	3622	1	2	2	3	4	5	6	7	7
.56	3631	3639	3648	3656	3664	3673	3681	3690	3698	3707	1	2	2	3	4	5	6	7	8
.57	3715	3724	3733	3741	3750	3758	3767	3776	3784	3793	1	2	3	3	4	5	6	7	8
.58	3802	3811	3819	3828	3837	3846	3855	3864	3873	3882	1	2	3	3	4	5	6	7	8
.59	3890	3899	3908	3917	3926	3936	3945	3954	3963	3972	1	2	3	4	5	5	6	7	8
.60	3981	3990	3999	4009	4018	4027	4036	4046	4055	4064	1	2	3	4	5	6	7	8	8
.61	4074	4083	4093	4102	4111	4121	4130	4140	4150	4159	1	2	3	4	5	6	7	8	9
.62	4169	4178	4188	4198	4207	4217	4227	4236	4246	4256	1	2	3	4	5	6	7	8	9
.63	4266	4276	4285	4295	4305	4315	4325	4335	4345	4355	1	2	3	4	5	6	7	8	9
.64	4365	4375	4385	4395	4406	4416	4426	4436	4446	4457	1	2	3	4	5	6	7	8	9
.65	4467	4477	4487	4498	4508	4519	4529	4539	4550	4560	1	2	3	4	5	6	7	8	9
.66	4571	4581	4592	4603	4613	4624	4634	4645	4656	4667	1	2	3	4	5	6	7	9	10
.67	4677	4688	4699	4710	4721	4732	4742	4753	4764	4775	1	2	3	4	5	7	8	9	10
.68	4786	4797	4808	4819	4831	4842	4853	4864	4875	4887	1	2	3	5	6	7	8	9	10
.69	4898	4909	4920	4932	4943	4955	4966	4977	4989	5000	1	2	3	5	6	7	8	9	10
.70	5012	5023	5035	5047	5058	5070	5082	5093	5105	5117	1	2	3	5	6	7	8	9	10
.71	5129	5140	5152	5164	5176	5188	5200	5212	5224	5236	1	2	4	5	6	7	8	10	11
.72	5248	5260	5272	5284	5297	5309	5321	5333	5346	5358	1	2	4	5	6	7	9	10	11
.73	5370	5383	5395	5408	5420	5433	5445	5458	5470	5483	1	3	4	5	6	7	9	10	11
.74	5495	5508	5521	5534	5546	5559	5572	5585	5598	5610	1	3	4	5	6	8	9	10	12
.75	5623	5636	5649	5662	5675	5689	5702	5715	5728	5741	1	3	4	5	7	8	9	11	12
.76	5754	5768	5781	5794	5808	5821	5834	5848	5861	5875	1	3	4	5	7	8	9	11	12
.77	5888	5902	5916	5929	5943	5957	5970	5984	5998	6012	1	3	4	5	7	8	10	11	12
.78	6026	6039	6053	6067	6081	6095	6109	6124	6138	6152	1	3	4	6	7	8	10	11	13
.79	6166	6180	6194	6209	6223	6237	6252	6266	6281	6295	1	3	4	6	7	9	10	11	13
.80	6310	6324	6339	6353	6368	6383	6397	6412	6427	6442	1	3	4	6	7	9	10	12	13
.81	6457	6471	6486	6501	6516	6531	6546	6561	6577	6592	2	3	5	6	8	9	11	12	14
.82	6607	6622	6637	6653	6668	6683	6699	6714	6730	6745	2	3	5	6	8	9	11	12	14
.83	6761	6776	6792	6808	6823	6839	6855	6871	6887	6902	2	3	5	6	8	9	11	13	14
.84	6918	6934	6950	6966	6982	6998	7015	7031	7047	7063	2	3	5	7	8	10	11	13	15
.85	7079	7096	7112	7129	7145	7161	7178	7194	7211	7228	2	3	5	7	8	10	12	13	15
.86	7244	7261	7278	7295	7311	7328	7345	7362	7379	7396	2	3	5	7	8	10	12	14	15
.87	7413	7430	7447	7464	7482	7499	7516	7534	7551	7568	2	4	5	7	9	10	12	14	16
.88	7586	7603	7621	7638	7656	7674	7691	7709	7727	7745	2	4	5	7	9	11	12	14	16
.89	7762	7780	7798	7816	7834	7852	7870	7889	7907	7925	2	4	6	7	9	11	13	15	16
.90	7943	7962	7980	7998	8017	8035	8054	8072	8091	8110	2	4	6	7	9	11	13	15	17
.91	8128	8147	8166	8185	8204	8222	8241	8260	8279	8299	2	4	6	8	9	11	13	15	17
.92	8318	8337	8356	8375	8395	8414	8433	8453	8472	8492	2	4	6	8	10	12	14	15	17
.93	8511	8531	8551	8570	8590	8610	8630	8650	8670	8690	2	4	6	8	10	12	14	16	18
.94	8710	8730	8750	8770	8790	8810	8831	8851	8872	8892	2	4	6	8	10	12	14	16	18
.95	8913	8933	8954	8974	8995	9016	9036	9057	9078	9099	2	4	6	8	10	12	15	17	19
.96	9120	9141	9162	9183	9204	9226	9247	9268	9290	9311	2	4	6	9	11	13	15	17	19
.97	9333	9354	9376	9397	9419	9441	9462	9484	9506	9528	2	4	6	9	11	13	15	17	19
.98	9550	9572	9594	9616	9638	9661	9683	9705	9727	9750	2	4	7	9	11	13	16	18	20
.99	9772	9795	9817	9840	9863	9886	9908	9931	9954	9977	2	5	7	9	11	14	16	18	21

TABLE III—FOUR-PLACE TRIGONOMETRIC FUNCTIONS

[Characteristics of Logarithms omitted—determine by the usual rule from the value]

Radians	Degrees	Sine Value	Sine Log_{10}	Tangent Value	Tangent Log_{10}	Cotangent Value	Cotangent Log_{10}	Cosine Value	Cosine Log_{10}		
.0000	0° 00′	.0000	——	.0000	——	——	——	1.0000	.0000	90° 00′	1.5708
.0029	10	.0029	.4637	.0029	.4637	343.77	.5363	1.0000	.0000	50	1.5679
.0058	20	.0058	.7648	.0058	.7648	171.89	.2352	1.0000	.0000	40	1.5650
.0087	30	.0087	.9408	.0087	.9409	114.59	.0591	1.0000	.0000	30	1.5621
.0116	40	.0116	.0658	.0116	.0658	85.940	.9342	.9999	.0000	20	1.5592
.0145	50	.0145	.1627	.0145	.1627	68.750	.8373	.9999	.0000	10	1.5563
.0175	1° 00′	.0175	.2419	.0175	.2419	57.290	.7581	.9998	.9999	89° 00′	1.5533
.0204	10	.0204	.3088	.0204	.3089	49.104	.6911	.9998	.9999	50	1.5504
.0233	20	.0233	.3668	.0233	.3669	42.964	.6331	.9997	.9999	40	1.5475
.0262	30	.0262	.4179	.0262	.4181	38.188	.5819	.9997	.9999	30	1.5446
.0291	40	.0291	.4637	.0291	.4638	34.368	.5362	.9996	.9998	20	1.5417
.0320	50	.0320	.5050	.0320	.5053	31.242	.4947	.9995	.9998	10	1.5388
.0349	2° 00′	.0349	.5428	.0349	.5431	28.636	.4569	.9994	.9997	88° 00′	1.5359
.0378	10	.0378	.5776	.0378	.5779	26.432	.4221	.9993	.9997	50	1.5330
.0407	20	.0407	.6097	.0407	.6101	24.542	.3899	.9992	.9996	40	1.5301
.0436	30	.0436	.6397	.0437	.6401	22.904	.3599	.9990	.9996	30	1.5272
.0465	40	.0465	.6677	.0466	.6682	21.470	.3318	.9989	.9995	20	1.5243
.0495	50	.0494	.6940	.0495	.6945	20.206	.3055	.9988	.9995	10	1.5213
.0524	3° 00′	.0523	.7188	.0524	.7194	19.081	.2806	.9986	.9994	87° 00′	1.5184
.0553	10	.0552	.7423	.0553	.7429	18.075	.2571	.9985	.9993	50	1.5155
.0582	20	.0581	.7645	.0582	.7652	17.169	.2348	.9983	.9993	40	1.5126
.0611	30	.0610	.7857	.0612	.7865	16.350	.2135	.9981	.9992	30	1.5097
.0640	40	.0640	.8059	.0641	.8067	15.605	.1933	.9980	.9991	20	1.5068
.0669	50	.0669	.8251	.0670	.8261	14.924	.1739	.9978	.9990	10	1.5039
.0698	4° 00′	.0698	.8436	.0699	.8446	14.301	.1554	.9976	.9989	86° 00′	1.5010
.0727	10	.0727	.8613	.0729	.8624	13.727	.1376	.9974	.9989	50	1.4981
.0756	20	.0756	.8783	.0758	.8795	13.197	.1205	.9971	.9988	40	1.4952
.0785	30	.0785	.8946	.0787	.8960	12.706	.1040	.9969	.9987	30	1.4923
.0814	40	.0814	.9104	.0816	.9118	12.251	.0882	.9967	.9986	20	1.4893
.0844	50	.0843	.9256	.0846	.9272	11.826	.0728	.9964	.9985	10	1.4864
.0873	5° 00′	.0872	.9403	.0875	.9420	11.430	.0580	.9962	.9983	85° 00′	1.4835
.0902	10	.0901	.9545	.0904	.9563	11.059	.0437	.9959	.9982	50	1.4806
.0931	20	.0929	.9682	.0934	.9701	10.712	.0299	.9957	.9981	40	1.4777
.0960	30	.0958	.9816	.0963	.9836	10.385	.0164	.9954	.9980	30	1.4748
.0989	40	.0987	.9945	.0992	.9966	10.078	.0034	.9951	.9979	20	1.4719
.1018	50	.1016	.0070	.1022	.0093	9.7882	.9907	.9948	.9977	10	1.4690
.1047	6° 00′	.1045	.0192	.1051	.0216	9.5144	.9784	.9945	.9976	84° 00′	1.4661
.1076	10	.1074	.0311	.1080	.0336	9.2553	.9664	.9942	.9975	50	1.4632
.1105	20	.1103	.0426	.1110	.0453	9.0098	.9547	.9939	.9973	40	1.4603
.1134	30	.1132	.0539	.1139	.0567	8.7769	.9433	.9936	.9972	30	1.4573
.1164	40	.1161	.0648	.1169	.0678	8.5555	.9322	.9932	.9971	20	1.4544
.1193	50	.1190	.0755	.1198	.0786	8.3450	.9214	.9929	.9969	10	1.4515
.1222	7° 00′	.1219	.0859	.1228	.0891	8.1443	.9109	.9925	.9968	83° 00′	1.4486
.1251	10	.1248	.0961	.1257	.0995	7.9530	.9005	.9922	.9966	50	1.4457
.1280	20	.1276	.1060	.1287	.1096	7.7704	.8904	.9918	.9964	40	1.4428
.1309	30	.1305	.1157	.1317	.1194	7.5958	.8806	.9914	.9963	30	1.4399
.1338	40	.1334	.1252	.1346	.1291	7.4287	.8709	.9911	.9961	20	1.4370
.1367	50	.1363	.1345	.1376	.1385	7.2687	.8615	.9907	.9959	10	1.4341
.1396	8° 00′	.1392	.1436	.1405	.1478	7.1154	.8522	.9903	.9958	82° 00′	1.4312
.1425	10	.1421	.1525	.1435	.1569	6.9682	.8431	.9899	.9956	50	1.4283
.1454	20	.1449	.1612	.1465	.1658	6.8269	.8342	.9894	.9954	40	1.4254
.1484	30	.1478	.1697	.1495	.1745	6.6912	.8255	.9890	.9952	30	1.4224
.1513	40	.1507	.1781	.1524	.1831	6.5606	.8169	.9886	.9950	20	1.4195
.1542	50	.1536	.1863	.1554	.1915	6.4348	.8085	.9881	.9948	10	1.4166
.1571	9° 00′	.1564	.1943	.1584	.1997	6.3138	.8003	.9877	.9946	81° 00′	1.4137
		Cosine Value	Cosine Log_{10}	Cotangent Value	Cotangent Log_{10}	Tangent Value	Tangent Log_{10}	Sine Value	Sine Log_{10}	Degrees	Radians

TABLE III—FOUR-PLACE TRIGONOMETRIC FUNCTIONS

[Characteristics of Logarithms omitted—determine by the usual rule from the value]

Radians	Degrees	Sine Value	Sine Log10	Tangent Value	Tangent Log10	Cotangent Value	Cotangent Log10	Cosine Value	Cosine Log10		
.1571	9° 00′	.1564	.1943	.1584	.1997	6.3138	.8003	.9877	.9946	81° 00′	1.4137
.1600	10	.1593	.2022	.1614	.2078	6.1970	.7922	.9872	.9944	50	1.4108
.1629	20	.1622	.2100	.1644	.2158	6.0844	.7842	.9868	.9942	40	1.4079
.1658	30	.1650	.2176	.1673	.2236	5.9758	.7764	.9863	.9940	30	1.4050
.1687	40	.1679	.2251	.1703	.2313	5.8708	.7687	.9858	.9938	20	1.4021
.1716	50	.1708	.2324	.1733	.2389	5.7694	.7611	.9853	.9936	10	1.3992
.1745	10° 00′	.1736	.2397	.1763	.2463	5.6713	.7537	.9848	.9934	80° 00′	1.3963
.1774	10	.1765	.2468	.1793	.2536	5.5764	.7464	.9843	.9931	50	1.3934
.1804	20	.1794	.2538	.1823	.2609	5.4845	.7391	.9838	.9929	40	1.3904
.1833	30	.1822	.2606	.1853	.2680	5.3955	.7320	.9833	.9927	30	1.3875
.1862	40	.1851	.2674	.1883	.2750	5.3093	.7250	.9827	.9924	20	1.3846
.1891	50	.1880	.2740	.1914	.2819	5.2257	.7181	.9822	.9922	10	1.3817
.1920	11° 00′	.1908	.2806	.1944	.2887	5.1446	.7113	.9816	.9919	79° 00′	1.3788
.1949	10	.1937	.2870	.1974	.2953	5.0658	.7047	.9811	.9917	50	1.3759
.1978	20	.1965	.2934	.2004	.3020	4.9894	.6980	.9805	.9914	40	1.3730
.2007	30	.1994	.2997	.2035	.3085	4.9152	.6915	.9799	.9912	30	1.3701
.2036	40	.2022	.3058	.2065	.3149	4.8430	.6851	.9793	.9909	20	1.3672
.2065	50	.2051	.3119	.2095	.3212	4.7729	.6788	.9787	.9907	10	1.3643
.2094	12° 00′	.2079	.3179	.2126	.3275	4.7046	.6725	.9781	.9904	78° 00′	1.3614
.2123	10	.2108	.3238	.2156	.3336	4.6382	.6664	.9775	.9901	50	1.3584
.2153	20	.2136	.3296	.2186	.3397	4.5736	.6603	.9769	.9899	40	1.3555
.2182	30	.2164	.3353	.2217	.3458	4.5107	.6542	.9763	.9896	30	1.3526
.2211	40	.2193	.3410	.2247	.3517	4.4494	.6483	.9757	.9893	20	1.3497
.2240	50	.2221	.3466	.2278	.3576	4.3897	.6424	.9750	.9890	10	1.3468
.2269	13° 00′	.2250	.3521	.2309	.3634	4.3315	.6366	.9744	.9887	77° 00′	1.3439
.2298	10	.2278	.3575	.2339	.3691	4.2747	.6309	.9737	.9884	50	1.3410
.2327	20	.2306	.3629	.2370	.3748	4.2193	.6252	.9730	.9881	40	1.3381
.2356	30	.2334	.3682	.2401	.3804	4.1653	.6196	.9724	.9878	30	1.3352
.2385	40	.2363	.3734	.2432	.3859	4.1126	.6141	.9717	.9875	20	1.3323
.2414	50	.2391	.3786	.2462	.3914	4.0611	.6086	.9710	.9872	10	1.3294
.2443	14° 00′	.2419	.3837	.2493	.3968	4.0108	.6032	.9703	.9869	76° 00′	1.3265
.2473	10	.2447	.3887	.2524	.4021	3.9617	.5979	.9696	.9866	50	1.3235
.2502	20	.2476	.3937	.2555	.4074	3.9136	.5926	.9689	.9863	40	1.3206
.2531	30	.2504	.3986	.2586	.4127	3.8667	.5873	.9681	.9859	30	1.3177
.2560	40	.2532	.4035	.2617	.4178	3.8208	.5822	.9674	.9856	20	1.3148
.2589	50	.2560	.4083	.2648	.4230	3.7760	.5770	.9667	.9853	10	1.3119
.2618	15° 00′	.2588	.4130	.2679	.4281	3.7321	.5719	.9659	.9849	75° 00′	1.3090
.2647	10	.2616	.4177	.2711	.4331	3.6891	.5669	.9652	.9846	50	1.3061
.2676	20	.2644	.4223	.2742	.4381	3.6470	.5619	.9644	.9843	40	1.3032
.2705	30	.2672	.4269	.2773	.4430	3.6059	.5570	.9636	.9839	30	1.3003
.2734	40	.2700	.4314	.2805	.4479	3.5656	.5521	.9628	.9836	20	1.2974
.2763	50	.2728	.4359	.2836	.4527	3.5261	.5473	.9621	.9832	10	1.2945
.2793	16° 00′	.2756	.4403	.2867	.4575	3.4874	.5425	.9613	.9828	74° 00′	1.2915
.2822	10	.2784	.4447	.2899	.4622	3.4495	.5378	.9605	.9825	50	1.2886
.2851	20	.2812	.4491	.2931	.4669	3.4124	.5331	.9596	.9821	40	1.2857
.2880	30	.2840	.4533	.2962	.4716	3.3759	.5284	.9588	.9817	30	1.2828
.2909	40	.2868	.4576	.2994	.4762	3.3402	.5238	.9580	.9814	20	1.2799
.2938	50	.2896	.4618	.3026	.4808	3.3052	.5192	.9572	.9810	10	1.2770
.2967	17° 00′	.2924	.4659	.3057	.4853	3.2709	.5147	.9563	.9806	73° 00′	1.2741
.2996	10	.2952	.4700	.3089	.4898	3.2371	.5102	.9555	.9802	50	1.2712
.3025	20	.2979	.4741	.3121	.4943	3.2041	.5057	.9546	.9798	40	1.2683
.3054	30	.3007	.4781	.3153	.4987	3.1716	.5013	.9537	.9794	30	1.2654
.3083	40	.3035	.4821	.3185	.5031	3.1397	.4969	.9528	.9790	20	1.2625
.3113	50	.3062	.4861	.3217	.5075	3.1084	.4925	.9520	.9786	10	1.2595
.3142	18° 00′	.3090	.4900	.3249	.5118	3.0777	.4882	.9511	.9782	72° 00′	1.2566
		Value	Log10	Value	Log10	Value	Log10	Value	Log10		
		Cosine		Cotangent		Tangent		Sine		Degrees	Radians

TABLE III—FOUR-PLACE TRIGONOMETRIC FUNCTIONS

[Characteristics of Logarithms omitted—determine by the usual rule from the value]

Radians	Degrees	Sine Value	Sine Log_{10}	Tangent Value	Tangent Log_{10}	Cotangent Value	Cotangent Log_{10}	Cosine Value	Cosine Log_{10}		
.3142	**18° 00′**	.3090	.4900	.3249	.5118	3.0777	.4882	.9511	.9782	**72° 00′**	1.2566
.3171	10	.3118	.4939	.3281	.5161	3.0475	.4839	.9502	.9778	50	1.2537
.3200	20	.3145	.4977	.3314	.5203	3.0178	.4797	.9492	.9774	40	1.2508
.3229	30	.3173	.5015	.3346	.5245	2.9887	.4755	.9483	.9770	30	1.2479
.3258	40	.3201	.5052	.3378	.5287	2.9600	.4713	.9474	.9765	20	1.2450
.3287	50	.3228	.5090	.3411	.5329	2.9319	.4671	.9465	.9761	10	1.2421
.3316	**19° 00′**	.3256	.5126	.3443	.5370	2.9042	.4630	.9455	.9757	**71° 00′**	1.2392
.3345	10	.3283	.5163	.3476	.5411	2.8770	.4589	.9446	.9752	50	1.2363
.3374	20	.3311	.5199	.3508	.5451	2.8502	.4549	.9436	.9748	40	1.2334
.3403	30	.3338	.5235	.3541	.5491	2.8239	.4509	.9426	.9743	30	1.2305
.3432	40	.3365	.5270	.3574	.5531	2.7980	.4469	.9417	.9739	20	1.2275
.3462	50	.3393	.5306	.3607	.5571	2.7725	.4429	.9407	.9734	10	1.2246
.3491	**20° 00′**	.3420	.5341	.3640	.5611	2.7475	.4389	.9397	.9730	**70° 00′**	1.2217
.3520	10	.3448	.5375	.3673	.5650	2.7228	.4350	.9387	.9725	50	1.2188
.3549	20	.3475	.5409	.3706	.5689	2.6985	.4311	.9377	.9721	40	1.2159
.3578	30	.3502	.5443	.3739	.5727	2.6746	.4273	.9367	.9716	30	1.2130
.3607	40	.3529	.5477	.3772	.5766	2.6511	.4234	.9356	.9711	20	1.2101
.3636	50	.3557	.5510	.3805	.5804	2.6279	.4196	.9346	.9706	10	1.2072
.3665	**21° 00′**	.3584	.5543	.3839	.5842	2.6051	.4158	.9336	.9702	**69° 00′**	1.2043
.3694	10	.3611	.5576	.3872	.5879	2.5826	.4121	.9325	.9697	50	1.2014
.3723	20	.3638	.5609	.3906	.5917	2.5605	.4083	.9315	.9692	40	1.1985
.3752	30	.3665	.5641	.3939	.5954	2.5386	.4046	.9304	.9687	30	1.1956
.3782	40	.3692	.5673	.3973	.5991	2.5172	.4009	.9293	.9682	20	1.1926
.3811	50	.3719	.5704	.4006	.6028	2.4960	.3972	.9283	.9677	10	1.1897
.3840	**22° 00′**	.3746	.5736	.4040	.6064	2.4751	.3936	.9272	.9672	**68° 00′**	1.1868
.3869	10	.3773	.5767	.4074	.6100	2.4545	.3900	.9261	.9667	50	1.1839
.3898	20	.3800	.5798	.4108	.6136	2.4342	.3864	.9250	.9661	40	1.1810
.3927	30	.3827	.5828	.4142	.6172	2.4142	.3828	.9239	.9656	30	1.1781
.3956	40	.3854	.5859	.4176	.6208	2.3945	.3792	.9228	.9651	20	1.1752
.3985	50	.3881	.5889	.4210	.6243	2.3750	.3757	.9216	.9646	10	1.1723
.4014	**23° 00′**	.3907	.5919	.4245	.6279	2.3559	.3721	.9205	.9640	**67° 00′**	1.1694
.4043	10	.3934	.5948	.4279	.6314	2.3369	.3686	.9194	.9635	50	1.1665
.4072	20	.3961	.5978	.4314	.6348	2.3183	.3652	.9182	.9629	40	1.1636
.4102	30	.3987	.6007	.4348	.6383	2.2998	.3617	.9171	.9624	30	1.1606
.4131	40	.4014	.6036	.4383	.6417	2.2817	.3583	.9159	.9618	20	1.1577
.4160	50	.4041	.6065	.4417	.6452	2.2637	.3548	.9147	.9613	10	1.1548
.4189	**24° 00′**	.4067	.6093	.4452	.6486	2.2460	.3514	.9135	.9607	**66° 00′**	1.1519
.4218	10	.4094	.6121	.4487	.6520	2.2286	.3480	.9124	.9602	50	1.1490
.4247	20	.4120	.6149	.4522	.6553	2.2113	.3447	.9112	.9596	40	1.1461
.4276	30	.4147	.6177	.4557	.6587	2.1943	.3413	.9100	.9590	30	1.1432
.4305	40	.4173	.6205	.4592	.6620	2.1775	.3380	.9088	.9584	20	1.1403
.4334	50	.4200	.6232	.4628	.6654	2.1609	.3346	.9075	.9579	10	1.1374
.4363	**25° 00′**	.4226	.6259	.4663	.6687	2.1445	.3313	.9063	.9573	**65° 00′**	1.1345
.4392	10	.4253	.6286	.4699	.6720	2.1283	.3280	.9051	.9567	50	1.1316
.4422	20	.4279	.6313	.4734	.6752	2.1123	.3248	.9038	.9561	40	1.1286
.4451	30	.4305	.6340	.4770	.6785	2.0965	.3215	.9026	.9555	30	1.1257
.4480	40	.4331	.6366	.4806	.6817	2.0809	.3183	.9013	.9549	20	1.1228
.4509	50	.4358	.6392	.4841	.6850	2.0655	.3150	.9001	.9543	10	1.1199
.4538	**26° 00′**	.4384	.6418	.4877	.6882	2.0503	.3118	.8988	.9537	**64° 00′**	1.1170
.4567	10	.4410	.6444	.4913	.6914	2.0353	.3086	.8975	.9530	50	1.1141
.4596	20	.4436	.6470	.4950	.6946	2.0204	.3054	.8962	.9524	40	1.1112
.4625	30	.4462	.6495	.4986	.6977	2.0057	.3023	.8949	.9518	30	1.1083
.4654	40	.4488	.6521	.5022	.7009	1.9912	.2991	.8936	.9512	20	1.1054
.4683	50	.4514	.6546	.5059	.7040	1.9768	.2960	.8923	.9505	10	1.1025
.4712	**27° 00′**	.4540	.6570	.5095	.7072	1.9626	.2928	.8910	.9499	**63° 00′**	1.0996
		Value Cosine	Log_{10} Cosine	Value Cotangent	Log_{10} Cotangent	Value Tangent	Log_{10} Tangent	Value Sine	Log_{10} Sine	Degrees	Radians

TABLE III—FOUR-PLACE TRIGONOMETRIC FUNCTIONS

[Characteristics of Logarithms omitted—determine by the usual rule from the value]

Radians	Degrees	Sine Value	Sine Log₁₀	Tangent Value	Tangent Log₁₀	Cotangent Value	Cotangent Log₁₀	Cosine Value	Cosine Log₁₀		
.4712	27° 00′	.4540	.6570	.5095	.7072	1.9626	.2928	.8910	.9499	63° 00′	1.0996
.4741	10	.4566	.6595	.5132	.7103	1.9486	.2897	.8897	.9492	50	1.0966
.4771	20	.4592	.6620	.5169	.7134	1.9347	.2866	.8884	.9486	40	1.0937
.4800	30	.4617	.6644	.5206	.7165	1.9210	.2835	.8870	.9479	30	1.0908
.4829	40	.4643	.6668	.5243	.7196	1.9074	.2804	.8857	.9473	20	1.0879
.4858	50	.4669	.6692	.5280	.7226	1.8940	.2774	.8843	.9466	10	1.0850
.4887	28° 00′	.4695	.6716	.5317	.7257	1.8807	.2743	.8829	.9459	62° 00′	1.0821
.4916	10	.4720	.6740	.5354	.7287	1.8676	.2713	.8816	.9453	50	1.0792
.4945	20	.4746	.6763	.5392	.7317	1.8546	.2683	.8802	.9446	40	1.0763
.4974	30	.4772	.6787	.5430	.7348	1.8418	.2652	.8788	.9439	30	1.0734
.5003	40	.4797	.6810	.5467	.7378	1.8291	.2622	.8774	.9432	20	1.0705
.5032	50	.4823	.6833	.5505	.7408	1.8165	.2592	.8760	.9425	10	1.0676
.5061	29° 00′	.4848	.6856	.5543	.7438	1.8040	.2562	.8746	.9418	61° 00′	1.0647
.5091	10	.4874	.6878	.5581	.7467	1.7917	.2533	.8732	.9411	50	1.0617
.5120	20	.4899	.6901	.5619	.7497	1.7796	.2503	.8718	.9404	40	1.0588
.5149	30	.4924	.6923	.5658	.7526	1.7675	.2474	.8704	.9397	30	1.0559
.5178	40	.4950	.6946	.5696	.7556	1.7556	.2444	.8689	.9390	20	1.0530
.5207	50	.4975	.6968	.5735	.7585	1.7437	.2415	.8675	.9383	10	1.0501
.5236	30° 00′	.5000	.6990	.5774	.7614	1.7321	.2386	.8660	.9375	60° 00′	1.0472
.5265	10	.5025	.7012	.5812	.7644	1.7205	.2356	.8646	.9368	50	1.0443
.5294	20	.5050	.7033	.5851	.7673	1.7090	.2327	.8631	.9361	40	1.0414
.5323	30	.5075	.7055	.5890	.7701	1.6977	.2299	.8616	.9353	30	1.0385
.5352	40	.5100	.7076	.5930	.7730	1.6864	.2270	.8601	.9346	20	1.0356
.5381	50	.5125	.7097	.5969	.7759	1.6753	.2241	.8587	.9338	10	1.0327
.5411	31° 00′	.5150	.7118	.6009	.7788	1.6643	.2212	.8572	.9331	59° 00′	1.0297
.5440	10	.5175	.7139	.6048	.7816	1.6534	.2184	.8557	.9323	50	1.0268
.5469	20	.5200	.7160	.6088	.7845	1.6426	.2155	.8542	.9315	40	1.0239
.5498	30	.5225	.7181	.6128	.7873	1.6319	.2127	.8526	.9308	30	1.0210
.5527	40	.5250	.7201	.6168	.7902	1.6212	.2098	.8511	.9300	20	1.0181
.5556	50	.5275	.7222	.6208	.7930	1.6107	.2070	.8496	.9292	10	1.0152
.5585	32° 00′	.5299	.7242	.6249	.7958	1.6003	.2042	.8480	.9284	58° 00′	1.0123
.5614	10	.5324	.7262	.6289	.7986	1.5900	.2014	.8465	.9276	50	1.0094
.5643	20	.5348	.7282	.6330	.8014	1.5798	.1986	.8450	.9268	40	1.0065
.5672	30	.5373	.7302	.6371	.8042	1.5697	.1958	.8434	.9260	30	1.0036
.5701	40	.5398	.7322	.6412	.8070	1.5597	.1930	.8418	.9252	20	1.0007
.5730	50	.5422	.7342	.6453	.8097	1.5497	.1903	.8403	.9244	10	.9977
.5760	33° 00′	.5446	.7361	.6494	.8125	1.5399	.1875	.8387	.9236	57° 00′	.9948
.5789	10	.5471	.7380	.6536	.8153	1.5301	.1847	.8371	.9228	50	.9919
.5818	20	.5495	.7400	.6577	.8180	1.5204	.1820	.8355	.9219	40	.9890
.5847	30	.5519	.7419	.6619	.8208	1.5108	.1792	.8339	.9211	30	.9861
.5876	40	.5544	.7438	.6661	.8235	1.5013	.1765	.8323	.9203	20	.9832
.5905	50	.5568	.7457	.6703	.8263	1.4919	.1737	.8307	.9194	10	.9803
.5934	34° 00′	.5592	.7476	.6745	.8290	1.4826	.1710	.8290	.9186	56° 00′	.9774
.5963	10	.5616	.7494	.6787	.8317	1.4733	.1683	.8274	.9177	50	.9745
.5992	20	.5640	.7513	.6830	.8344	1.4641	.1656	.8258	.9169	40	.9716
.6021	30	.5664	.7531	.6873	.8371	1.4550	.1629	.8241	.9160	30	.9687
.6050	40	.5688	.7550	.6916	.8398	1.4460	.1602	.8225	.9151	20	.9657
.6080	50	.5712	.7568	.6959	.8425	1.4370	.1575	.8208	.9142	10	.9628
.6109	35° 00′	.5736	.7586	.7002	.8452	1.4281	.1548	.8192	.9134	55° 00′	.9599
.6138	10	.5760	.7604	.7046	.8479	1.4193	.1521	.8175	.9125	50	.9570
.6167	20	.5783	.7622	.7089	.8506	1.4106	.1494	.8158	.9116	40	.9541
.6196	30	.5807	.7640	.7133	.8533	1.4019	.1467	.8141	.9107	30	.9512
.6225	40	.5831	.7657	.7177	.8559	1.3934	.1441	.8124	.9098	20	.9483
.6254	50	.5854	.7675	.7221	.8586	1.3848	.1414	.8107	.9089	10	.9454
.6283	36° 00′	.5878	.7692	.7265	.8613	1.3764	.1387	.8090	.9080	54° 00′	.9425
		Value Cosine	Log₁₀	Value Cotangent	Log₁₀	Value Tangent	Log₁₀	Value Sine	Log₁₀	Degrees	Radians

TABLE III—FOUR-PLACE TRIGONOMETRIC FUNCTIONS

[Characteristics of Logarithms omitted—determine by the usual rule from the value]

Radians	Degrees	Sine Value	Sine Log_{10}	Tangent Value	Tangent Log_{10}	Cotangent Value	Cotangent Log_{10}	Cosine Value	Cosine Log_{10}		
.6283	36° 00′	.5878	.7692	.7265	.8613	1.3764	.1387	.8090	.9080	54° 00′	.9425
.6312	10	.5901	.7710	.7310	.8639	1.3680	.1361	.8073	.9070	50	.9396
.6341	20	.5925	.7727	.7355	.8666	1.3597	.1334	.8056	.9061	40	.9367
.6370	30	.5948	.7744	.7400	.8692	1.3514	.1308	.8039	.9052	30	.9338
.6400	40	.5972	.7761	.7445	.8718	1.3432	.1282	.8021	.9042	20	.9308
.6429	50	.5995	.7778	.7490	.8745	1.3351	.1255	.8004	.9033	10	.9279
.6458	37° 00′	.6018	.7795	.7536	.8771	1.3270	.1229	.7986	.9023	53° 00′	.9250
.6487	10	.6041	.7811	.7581	.8797	1.3190	.1203	.7969	.9014	50	.9221
.6516	20	.6065	.7828	.7627	.8824	1.3111	.1176	.7951	.9004	40	.9192
.6545	30	.6088	.7844	.7673	.8850	1.3032	.1150	.7934	.8995	30	.9163
.6574	40	.6111	.7861	.7720	.8876	1.2954	.1124	.7916	.8985	20	.9134
.6603	50	.6134	.7877	.7766	.8902	1.2876	.1098	.7898	.8975	10	.9105
.6632	38° 00′	.6157	.7893	.7813	.8928	1.2799	.1072	.7880	.8965	52° 00′	.9076
.6661	10	.6180	.7910	.7860	.8954	1.2723	.1046	.7862	.8955	50	.9047
.6690	20	.6202	.7926	.7907	.8980	1.2647	.1020	.7844	.8945	40	.9018
.6720	30	.6225	.7941	.7954	.9006	1.2572	.0994	.7826	.8935	30	.8988
.6749	40	.6248	.7957	.8002	.9032	1.2497	.0968	.7808	.8925	20	.8959
.6778	50	.6271	.7973	.8050	.9058	1.2423	.0942	.7790	.8915	10	.8930
.6807	39° 00′	.6293	.7989	.8098	.9084	1.2349	.0916	.7771	.8905	51° 00′	.8901
.6836	10	.6316	.8004	.8146	.9110	1.2276	.0890	.7753	.8895	50	.8872
.6865	20	.6338	.8020	.8195	.9135	1.2203	.0865	.7735	.8884	40	.8843
.6894	30	.6361	.8035	.8243	.9161	1.2131	.0839	.7716	.8874	30	.8814
.6923	40	.6383	.8050	.8292	.9187	1.2059	.0813	.7698	.8864	20	.8785
.6952	50	.6406	.8066	.8342	.9212	1.1988	.0788	.7679	.8853	10	.8756
.6981	40° 00′	.6428	.8081	.8391	.9238	1.1918	.0762	.7660	.8843	50° 00′	.8727
.7010	10	.6450	.8096	.8441	.9264	1.1847	.0736	.7642	.8832	50	.8698
.7039	20	.6472	.8111	.8491	.9289	1.1778	.0711	.7623	.8821	40	.8668
.7069	30	.6494	.8125	.8541	.9315	1.1708	.0685	.7604	.8810	30	.8639
.7098	40	.6517	.8140	.8591	.9341	1.1640	.0659	.7585	.8800	20	.8610
.7127	50	.6539	.8155	.8642	.9366	1.1571	.0634	.7566	.8789	10	.8581
.7156	41° 00′	.6561	.8169	.8693	.9392	1.1504	.0608	.7547	.8778	49° 00′	.8552
.7185	10	.6583	.8184	.8744	.9417	1.1436	.0583	.7528	.8767	50	.8523
.7214	20	.6604	.8198	.8796	.9443	1.1369	.0557	.7509	.8756	40	.8494
.7243	30	.6626	.8213	.8847	.9468	1.1303	.0532	.7490	.8745	30	.8465
.7272	40	.6648	.8227	.8899	.9494	1.1237	.0506	.7470	.8733	20	.8436
.7301	50	.6670	.8241	.8952	.9519	1.1171	.0481	.7451	.8722	10	.8407
.7330	42° 00′	.6691	.8255	.9004	.9544	1.1106	.0456	.7431	.8711	48° 00′	.8378
.7359	10	.6713	.8269	.9057	.9570	1.1041	.0430	.7412	.8699	50	.8348
.7389	20	.6734	.8283	.9110	.9595	1.0977	.0405	.7392	.8688	40	.8319
.7418	30	.6756	.8297	.9163	.9621	1.0913	.0379	.7373	.8676	30	.8290
.7447	40	.6777	.8311	.9217	.9646	1.0850	.0354	.7353	.8665	20	.8261
.7476	50	.6799	.8324	.9271	.9671	1.0786	.0329	.7333	.8653	10	.8232
.7505	43° 00′	.6820	.8338	.9325	.9697	1.0724	.0303	.7314	.8641	47° 00′	.8203
.7534	10	.6841	.8351	.9380	.9722	1.0661	.0278	.7294	.8629	50	.8174
.7563	20	.6862	.8365	.9435	.9747	1.0599	.0253	.7274	.8618	40	.8145
.7592	30	.6884	.8378	.9490	.9772	1.0538	.0228	.7254	.8606	30	.8116
.7621	40	.6905	.8391	.9545	.9798	1.0477	.0202	.7234	.8594	20	.8087
.7650	50	.6926	.8405	.9601	.9823	1.0416	.0177	.7214	.8582	10	.8058
.7679	44° 00′	.6947	.8418	.9657	.9848	1.0355	.0152	.7193	.8569	46° 00′	.8029
.7709	10	.6967	.8431	.9713	.9874	1.0295	.0126	.7173	.8557	50	.7999
.7738	20	.6988	.8444	.9770	.9899	1.0235	.0101	.7153	.8545	40	.7970
.7767	30	.7009	.8457	.9827	.9924	1.0176	.0076	.7133	.8532	30	.7941
.7796	40	.7030	.8469	.9884	.9949	1.0117	.0051	.7112	.8520	20	.7912
.7825	50	.7050	.8482	.9942	.9975	1.0058	.0025	.7092	.8507	10	.7883
.7854	45° 00′	.7071	.8495	1.0000	.0000	1.0000	.0000	.7071	.8495	45° 00′	.7854
		Value Cosine	Log_{10}	Value Cotangent	Log_{10}	Value Tangent	Log_{10}	Value Sine	Log_{10}	Degrees	Radians

TABLE IV—NAPIERIAN OR NATURAL LOGARITHMS

0-9

N	0	1	2	3	4	5	6	7	8	9
0.0		5.395	6.088	6.493	6.781	7.004	7.187	7.341	7.474	7.592
0.1	7.697 (Take tabular value − 10)	7.793	7.880	7.960	8.034	8.103	8.167	8.228	8.285	8.339
0.2	8.391	8.439	8.486	8.530	8.573	8.614	8.653	8.691	8.727	8.762
0.3	8.796	8.829	8.861	8.891	8.921	8.950	8.978	9.006	9.032	9.058
0.4	9.084	9.108	9.132	9.156	9.179	9.201	9.223	9.245	9.266	9.287
0.5	9.307	9.327	9.346	9.365	9.384	9.402	9.420	9.438	9.455	9.472
0.6	9.489	9.506	9.522	9.538	9.554	9.569	9.584	9.600	9.614	9.629
0.7	9.643	9.658	9.671	9.685	9.699	9.712	9.726	9.739	9.752	9.764
0.8	9.777	9.789	9.802	9.814	9.826	9.837	9.849	9.861	9.872	9.883
0.9	9.895	9.906	9.917	9.927	9.938	9.949	9.959	9.970	9.980	9.990
1.0	0.00000	0995	1980	2956	3922	4879	5827	6766	7696	8618
1.1	9531	*0436	*1333	*2222	*3103	*3976	*4842	*5700	*6551	*7395
1.2	0.1 8232	9062	9885	*0701	*1511	*2314	*3111	*3902	*4686	*5464
1.3	0.2 6236	7003	7763	8518	9267	*0010	*0748	*1481	*2208	*2930
1.4	0.3 3647	4359	5066	5767	6464	7156	7844	8526	9204	9878
1.5	0.4 0547	1211	1871	2527	3178	3825	4469	5108	5742	6373
1.6	7000	7623	8243	8858	9470	*0078	*0682	*1282	*1879	*2473
1.7	0.5 3063	3649	4232	4812	5389	5962	6531	7098	7661	8222
1.8	8779	9333	9884	*0432	*0977	*1519	*2058	*2594	*3127	*3658
1.9	0.6 4185	4710	5233	5752	6269	6783	7294	7803	8310	8813
2.0	9315	9813	*0310	*0804	*1295	*1784	*2271	*2755	*3237	*3716
2.1	0.7 4194	4669	5142	5612	6081	6547	7011	7473	7932	8390
2.2	8846	9299	9751	*0200	*0648	*1093	*1536	*1978	*2418	*2855
2.3	0.8 3291	3725	4157	4587	5015	5442	5866	6289	6710	7129
2.4	7547	7963	8377	8789	9200	9609	*0016	*0422	*0826	*1228
2.5	0.9 1629	2028	2426	2822	3216	3609	4001	4391	4779	5166
2.6	5551	5935	6317	6698	7078	7456	7833	8208	8582	8954
2.7	9325	9695	*0063	*0430	*0796	*1160	*1523	*1885	*2245	*2604
2.8	1.0 2962	3318	3674	4028	4380	4732	5082	5431	5779	6126
2.9	6471	6815	7158	7500	7841	8181	8519	8856	9192	9527
3.0	9861	*0194	*0526	*0856	*1186	*1514	*1841	*2168	*2493	*2817
3.1	1.1 3140	3462	3783	4103	4422	4740	5057	5373	5688	6002
3.2	6315	6627	6938	7248	7557	7865	8173	8479	8784	9089
3.3	9392	9695	9996	*0297	*0597	*0896	*1194	*1491	*1788	*2083
3.4	1.2 2378	2671	2964	3256	3547	3837	4127	4415	4703	4990
3.5	5276	5562	5846	6130	6413	6695	6976	7257	7536	7815
3.6	8093	8371	8647	8923	9198	9473	9746	*0019	*0291	*0563
3.7	1.3 0833	1103	1372	1641	1909	2176	2442	2708	2972	3237
3.8	3500	3763	4025	4286	4547	4807	5067	5325	5584	5841
3.9	6098	6354	6609	6864	7118	7372	7624	7877	8128	8379
4.0	8629	8879	9128	9377	9624	9872	*0118	*0364	*0610	*0854
4.1	1.4 1099	1342	1585	1828	2070	2311	2552	2792	3031	3270
4.2	3508	3746	3984	4220	4456	4692	4927	5161	5395	5629
4.3	5862	6094	6326	6557	6787	7018	7247	7476	7705	7933
4.4	8160	8387	8614	8840	9065	9290	9515	9739	9962	*0185
4.5	1.5 0408	0630	0851	1072	1293	1513	1732	1951	2170	2388
4.6	2606	2823	3039	3256	3471	3687	3902	4116	4330	4543
4.7	4756	4969	5181	5393	5604	5814	6025	6235	6444	6653
4.8	6862	7070	7277	7485	7691	7898	8104	8309	8515	8719
4.9	8924	9127	9331	9534	9737	9939	*0141	*0342	*0543	*0744
5.0	1.6 0944	1144	1343	1542	1741	1939	2137	2334	2531	2728
N	0	1	2	3	4	5	6	7	8	9

TABLE IV—NAPIERIAN OR NATURAL LOGARITHMS

10-99

10	2.30259	**25**	3.21888	**40**	3.68888	**55**	4.00733	**70**	4.24850	**85**	4.44265
11	2.39790	26	3.25810	41	3.71357	56	4.02535	71	4.26268	86	4.45435
12	2.48491	27	3.29584	42	3.73767	57	4.04305	72	4.27667	87	4.46591
13	2.56495	28	3.33220	43	3.76120	58	4.06044	73	4.29046	88	4.47734
14	2.63906	29	3.36730	44	3.78419	59	4.07754	74	4.30407	89	4.48864
15	2.70805	**30**	3.40120	**45**	3.80666	**60**	4.09434	**75**	4.31749	**90**	4.49981
16	2.77259	31	3.43399	46	3.82864	61	4.11087	76	4.33073	91	4.51086
17	2.83321	32	3.46574	47	3.85015	62	4.12713	77	4.34381	92	4.52179
18	2.89037	33	3.49651	48	3.87120	63	4.14313	78	4.35671	93	4.53260
19	2.94444	34	3.52636	49	3.89182	64	4.15888	79	4.36945	94	4.54329
20	2.99573	**35**	3.55535	**50**	3.91202	**65**	4.17439	**80**	4.38203	**95**	4.55388
21	3.04452	36	3.58352	51	3.93183	66	4.18965	81	4.39445	96	4.56435
22	3.09104	37	3.61092	52	3.95124	67	4.20469	82	4.40672	97	4.57471
23	3.13549	38	3.63759	53	3.97029	68	4.21951	83	4.41884	98	4.58497
24	3.17805	39	3.66356	54	3.98898	69	4.23411	84	4.43082	99	4.59512

100-409

N	0	1	2	3	4	5	6	7	8	9
10	4.6 0517	1512	2497	3473	4439	5396	6344	7283	8213	9135
11	4.7 0048	0953	1850	2739	3620	4493	5359	6217	7068	7912
12	8749	9579	*0402	*1218	*2028	*2831	*3628	*4419	*5203	*5981
13	4.8 6753	7520	8280	9035	9784	*0527	*1265	*1998	*2725	*3447
14	4.9 4164	4876	5583	6284	6981	7673	8361	9043	9721	*0395
15	5.0 1064	1728	2388	3044	3695	4343	4986	5625	6260	6890
16	7517	8140	8760	9375	9987	*0595	*1199	*1799	*2396	*2990
17	5.1 3580	4166	4749	5329	5906	6479	7048	7615	8178	8739
18	9296	9850	*0401	*0949	*1494	*2036	*2575	*3111	*3644	*4175
19	5.2 4702	5227	5750	6269	6786	7300	7811	8320	8827	9330
20	9832	*0330	*0827	*1321	*1812	*2301	*2788	*3272	*3754	*4233
21	5.3 4711	5186	5659	6129	6598	7064	7528	7990	8450	8907
22	9363	9816	*0268	*0717	*1165	*1610	*2053	*2495	*2935	*3372
23	5.4 3808	4242	4674	5104	5532	5959	6383	6806	7227	7646
24	8064	8480	8894	9306	9717	*0126	*0533	*0939	*1343	*1745
25	5.5 2146	2545	2943	3339	3733	4126	4518	4908	5296	5683
26	6068	6452	6834	7215	7595	7973	8350	8725	9099	9471
27	9842	*0212	*0580	*0947	*1313	*1677	*2040	*2402	*2762	*3121
28	5.6 3479	3835	4191	4545	4897	5249	5599	5948	6296	6643
29	6988	7332	7675	8017	8358	8698	9036	9373	9709	*0044
30	5.7 0378	0711	1043	1373	1703	2031	2359	2685	3010	3334
31	3657	3979	4300	4620	4939	5257	5574	5890	6205	6519
32	6832	7144	7455	7765	8074	8383	8690	8996	9301	9606
33	9909	*0212	*0513	*0814	*1114	*1413	*1711	*2008	*2305	*2600
34	5.8 2895	3188	3481	3773	4064	4354	4644	4932	5220	5507
35	5793	6079	6363	6647	6930	7212	7493	7774	8053	8332
36	8610	8888	9164	9440	9715	9990	*0263	*0536	*0808	*1080
37	5.9 1350	1620	1889	2158	2426	2693	2959	3225	3489	3754
38	4017	4280	4542	4803	5064	5324	5584	5842	6101	6358
39	6615	6871	7126	7381	7635	7889	8141	8394	8645	8896
40	9146	9396	9645	9894	*0141	*0389	*0635	*0881	*1127	*1372
N	0	1	2	3	4	5	6	7	8	9

Above 409, use the formula $\log_e 10n = \log_e n + \log_e 10 = \log_e n + 2.30258509$, or the formula $\log_e n = \log_e 10 \cdot \log_{10} n = 2.30258509 \log_{10} n$.

INDEX

PART I*

Boldface numbers refer to entries in Volume II of *Television for Radiomen.* (Volume II begins on page 338. Part II, which is devoted to color television, follows page 662 in Volume II.)

* Index to pages 1–662 of Part I. Index to pages 1–93 of Part II follows Index to Part I.

INDEX
PART II*

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* Index to pages 1–662 of Part I precedes index to pages 1–93 of Part II.