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OUR COVER

. . . features the Home Instrument Television Engineering activity in Indianapolis, Indiana. In the background J. McDonald (left) and K. Siwko discuss the integrated-circuit application in the 12-inch, transistorized, black-and-white, portable TV receiver. Standing in center: L. Cochran, P. McCabe, and D. Willis examine the highly complex deflection yoke for the rectangular, 90°, color picture tube. Chassis in foreground features the more compact design of today's color receivers that use shorter, rectangular picture tubes. Inset at top is a "Lundberg," 21-inch, color TV console being enjoyed in the viewer's home.

The Challenging Field of Home Instruments

In the last five years RCA's home instrument business increased several fold. Although color television is a major factor in this increase, black and white television receivers were sold in a volume in 1965 exceeded only in one previous year in RCA's history. The radio and Victrola sector too is in a strong growth era.

Not only are the basic product lines being produced in larger volumes, but rapid increases in the number of models and variations are being required for the market place. The engineering work load is, in consequence, proportionately increasing and time cycles are under strong compressive pressures. It is simultaneously true that transistors, integrated circuits, and other technological advances along with radically revised production and merchandising systems must be evaluated and innovated in a world-wide competitive atmosphere.

The engineering effort in RCA Victor Home Instruments is increasingly designed to complement and facilitate the production and merchandising sectors and can no longer exist, if it ever could, without complete, skillful integration into the total business effort.

This issue describes some of television engineering's recent efforts in design. In 1965 more than 2,000,000 RCA television receivers were manufactured from this engineering group's designs and sold by the RCA Sales Corporation.

Magnitudes of this size dramatically illustrate that no detail is trivial and that tremendous technical, manufacturing, and merchandising sophistication is necessary. The evaluation of products is not long delayed, and only the best is tolerated by the consumer when labeled RCA Victor.

This is the challenge for now and the future.

E. I. Anderson,
*Manager, Operations,**
RCA Victor
Home Instruments Division
Indianapolis, Indiana

*Since this message was written, Mr. Anderson has been appointed Vice President, Value Assurance, RCA Sales Corporation.

RCA ENGINEER

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PAPERS

NOTES

DEPARTMENTS

A TECHNICAL JOURNAL PUBLISHED BY **RADIO CORPORATION OF AMERICA,** PRODUCT ENGINEERING 2-8, CAMDEN, N. J.

• To disseminate to RCA engineers technical information of professional value. • To publish in an appropriate manner important technical developments at RCA, and the role of the engineer. • To serve as a medium of interchange of technical information between various groups at RCA. • To create a community of engineering interest within the company by stressing the interrelated nature of all technical contributions. • To help publicize engineering achievements in a manner that will promote the interests and reputation of RCA in the engineering field. • To provide a convenient means by which the RCA engineer may review his professional work before associates and engineering management. • To announce outstanding and unusual achievements of RCA engineers in a manner most likely to enhance their prestige and professional status.

An index to RCA ENGINEER articles appears annually in the April-May issue.

To the home-instrument engineer, the consumers' demands for minimum cost, reliability, high performance, modern styling, beauty, and easy maintenance are paramount factors of final design. The importance of these factors is discussed and related to the engineer's role in providing a product that satisfies these demands and that can be mass produced economically.

The Engineer and the Corporation

ENGINEERING FOR PRODUCT PERFORMANCE AND ECONOMY —KEY TO THE CUSTOMER

C. HOYT, Staff Engineer
RCA Victor Home Instruments Div., Indianapolis, Ind.

Over the years RCA has gained an enviable reputation for its reliable engineering and high-quality products. To accomplish these goals in the design and production of RCA Victor Home Instruments requires special emphasis on product performance, high reliability, and customer appeal at an economical cost. This delicate balance of sometimes opposing factors constitutes a special challenge to the home instruments engineer, especially in light of today's high volume of business requiring mass-production methods.

ENGINEERING FOR LARGE-VOLUME PRODUCTION

The engineering and the eventual mass production of a modern RCA Victor Home Instruments' product can be likened to the same processes involving the automobile; both are the combined result of highly sophisticated research, design, engineering, material procurement, and production techniques. The final result is a very complex product that is useful and, in many ways, necessary to the vast majority of the population of the United States. The key to the success of both industries is that they have the ability to mass produce a product demanded by so many people and yet keep it within the consumer's ability to purchase and maintain.

In the field of home-instrument electronics, market complications and technical complexities at least equivalent to those of the automobile industry are present. The considerations of initial cost and maintenance of color TV receivers, for example, must be compatible with the economic realities of millions of U. S. citizens. Color TV and, to a lesser degree, other home-instrument products represent highly complicated electronic equipments which have achieved high-volume use through advances first in engineering and subsequently in production and maintenance techniques.

The economies provided by these products are possible only because materials and systems are kept abreast of advancing technology. For example, product engineering is done for vast quantities, the products are developed for low piece-cost tools, and the factories are geared to run with the utmost efficiency. To achieve high levels of performance from the many home-instrument products coming off the high-speed production lines, requires a sophistication and a thoroughness of engineering that is neither necessary nor feasible for the production of smaller numbers of equipments. Before such production lines are ever set up, the engineering cycle must include provisions in circuitry and mechanics for specified probability distributions for each of the hundreds of parts. Mass-quantity designs cannot be marginal for any reasonable accumulation of tolerances. For 10 radar sets it may be desirable to use circuits or components that can be selected or adjusted during production to yield the desired performance; but, for a million TV receivers, it is obviously impossible to select or adjust as many as 100 completed instruments in an hour. Therefore, if the required high-performance level is to be met, the burden of the higher order of refinement of design, production, and test rests on the engineer.

Final manuscript received February 10, 1966

ENGINEERING FOR ECONOMY

The home-instrument engineer is competing against worldwide professional competence in the production of consumer-pleasing performance; moreover he must do this on renewable short-term leases. He is also achieving economy, in very real time, through a blending of technical design with costs of material, labor, machine, and capital utilization. A typical result of this blending is shown in Fig. 1 on a year-to-year basis.

The true cost of a product can be quite difficult to assess, and in many cases the absolute true cost is never really known. The simplest and most direct portions of the true cost that can be assessed accurately are the piece prices, direct-labor costs, and tooling costs; even informative engineering costs and warranty charges can be determined.

The relative costs of factory production systems, however, are difficult to ascertain. For instance, which are more economical, the etched-circuit boards or the conventional hand-wired circuits shown in Fig. 2? To answer this question, the factors of engineering, tooling, machines, factories, capital, flexibility, performance, reliability, and service must be considered. The outcome may well depend upon total quantities, number of model years of basic similarity, and even such intangibles as product image and advertising philosophy. The cost of tooling *hard* tools for large-volume production can be excessive and is an important factor in the cost determinations.

Although it is difficult to determine accurately whether etched-circuit costs are different than hand-wiring, available records indicate that reliability of the former is significantly better. Also, the controlled lead dress of etched circuits minimizes the circuit variations the designer must face, permitting better average performance.

The tubes used in color and monochrome TV receivers have improved remarkably in recent years. Such improvements as dark heaters, uniform cathodes, frame grids, advanced glass technology, improved materials, and processing illustrate the great engineering advances in this area. Consequently, today's tubes are up to 10 times more reliable than previous versions and yield better performance for less cost. Circuits, tubes, and components are all designed for the best possible blend of cost, performance, reliability, etc.

Transistors and transistor circuits also are improving, but the economic crossover of transistor and vacuum-tube circuits in TV has been delayed. It is anticipated that transistor and integrated circuits will make greater penetrations into TV over the next 3 to 5 years.

The challenge of economy and low costs to the home-instruments engineer are as important and as demanding as performance considerations. A device that performs well but costs excessively does *not* reflect good engineering. The low-cost design that does not provide maximum or adequate performance does *not* represent good engineering either, since high reliability is considered mandatory. Furthermore, the engineering results are impartially and inexorably checked each year in the fiercely competitive market place.

ENGINEERING FOR PERFORMANCE AND RELIABILITY

Since millions of the same components, millions of the same circuits, and thousands of the same instruments will be produced, it is necessary to design individual components and circuits for an accurately specified role. This emphasis on component engineering results in better components for the application at less cost, giving the consumer a better performing product that is competitively priced.

The finished product will ultimately be called upon to perform in a predictable, realistic environment. It is very important that the conditions of performance and use be accurately known by the design engineer. This, of course, is a key factor influencing the eventual consumer price of the product. Many nonconsumer electronic products can be called upon to perform in extremely harsh or often unpredictable environments. For these reasons the equipment must be built to withstand extreme shock, vibration, moisture, hostile chemicals, extremely high and low temperature, and fungus. Individual parts must be built for contingencies that are not accurately specified and must be extensively tested and checked, which rapidly increases the cost of the part.

Unlike nonconsumer products, home-instrument products must be built to perform reliably under the conditions imposed by the mass market. Television receivers (for example) must be capable of operating at an altitude of 20,000 feet to insure minimum corona at both altitude and humidity extremes; and line-operated equipment will seldom

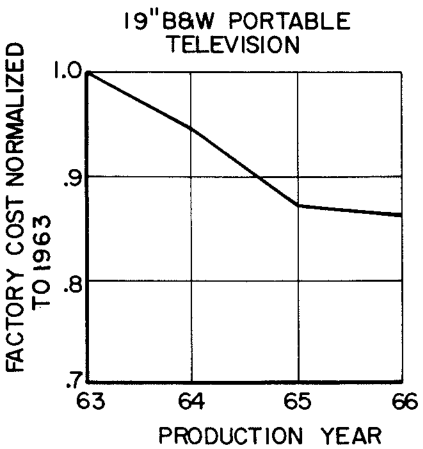
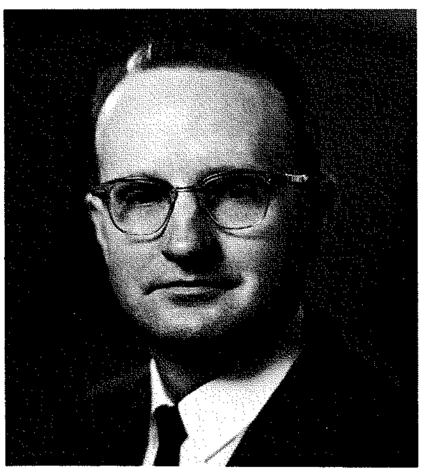


Fig. 1—Reduction in costs obtained in production of 19-inch, monochrome, portable TV receiver.

CLYDE W. HOYT is a graduate of Morningside College and Iowa State University in Arts and Electrical Engineering. He has been actively engaged in many phases of Home Instruments engineering; his work in television dates back to the introduction of the 630TS in 1946. In recent years he has been Staff Engineer for Home Instruments Division. Mr. Hoyt holds approximately 24 patents, primarily in the TV area.



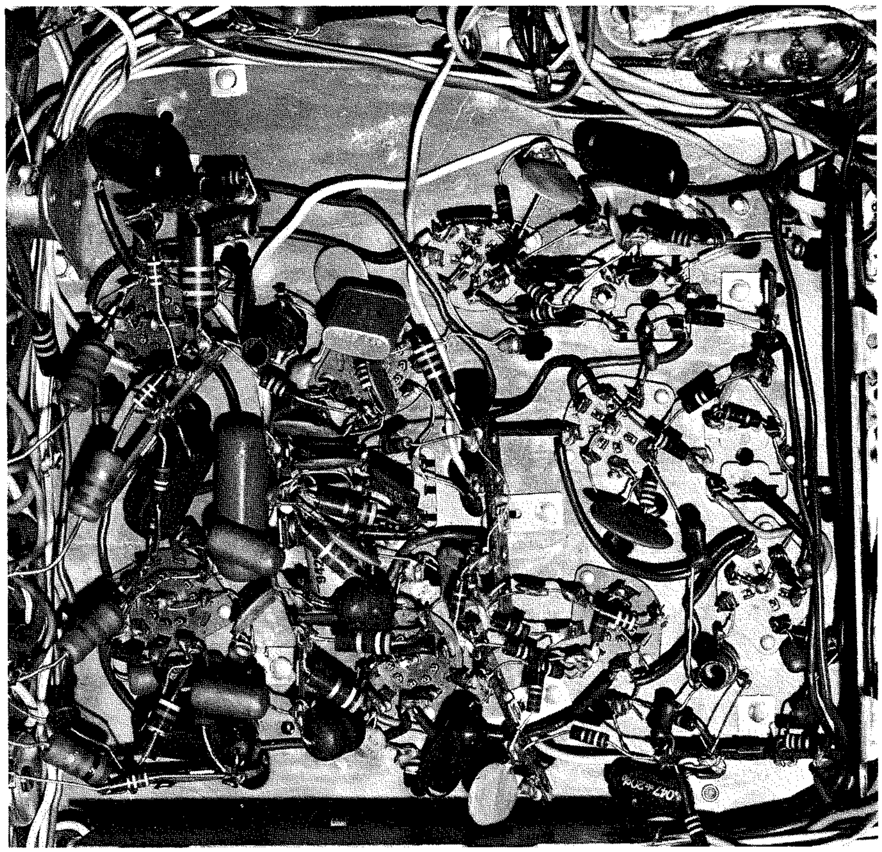
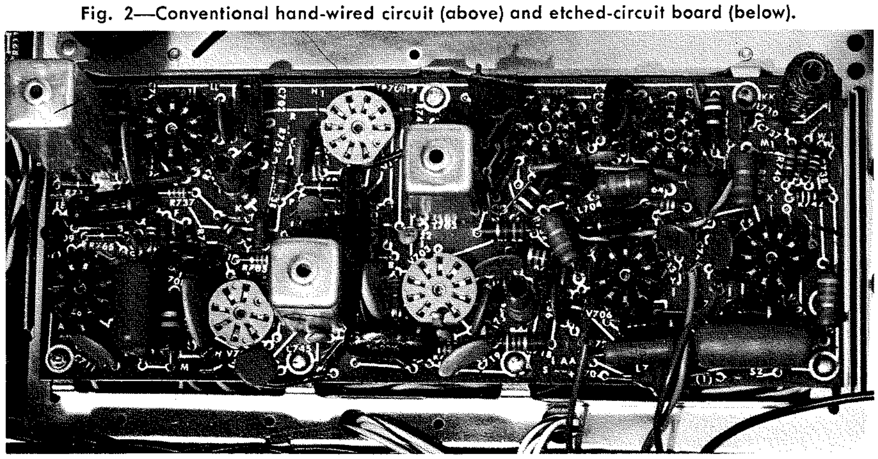


Fig. 2—Conventional hand-wired circuit (above) and etched-circuit board (below).



be called upon to operate in temperatures beyond the range of 60° to 120° F. Home-instrument products must tolerate shipping and drop tests, but they will generally be stationary during operation. Obviously, then, electronic and mechanical parts can be produced in volume at substantially lower cost when conditions are exactly known.

It is of interest to note that although home instrument products are not designed to take the treatment imposed by some space rocket environmental tests, a recent successful check of black-and-white and color television on a space test centrifuge and shaker, Fig. 3, showed that consumer shipping and drop tests demand a rugged, reliable product.

Every effort is made in determining the mechanical configuration of a part to keep assembly, tuning, and trimming to a minimum. Etched-circuit boards with many machine-inserted parts are used wherever possible (Fig. 4). Fixed-tuned coils molded in plastic of high dimensional stability to tight electrical tolerances are used to reduce factory aligning. Packaged electronic circuits which combine many resistors and capacitors on a porcelain substrate are used to reduce size and the number of connections, and to increase reliability. Power wire-wrap connections are used extensively in making connections to circuit boards to reduce labor costs and to provide high reliability.

STYLES AND MODELS

Molded plastic and vacuum-formed cabinets, having tastefully designed metal extrusions for trim and strength, are used where the extremely high volume justifies the high tooling costs.

Modern technology even shows up in wood cabinets. Decorative shapes, molded of wood fibers and finished with the rest of the cabinet, do a remarkable job of simulating intricately carved pieces, and many TV cabinets are now shipped to the assembly plant finished and knocked down, to be assembled around the front die-cast decorative metal mask and metal bottom which become structural parts of the finished cabinet.

Like the automobile industry, the home-instrument manufacturing operation must produce dozens of different models in each product class, and still maintain a cost picture that reflects a factory geared to run huge quantities. Again like the auto industry, this is accomplished by designing many models of a product category with identical components but with differences in cabinet style and easily added accessories.

HIGH-SPEED TESTS

As might be expected, the high order of teamwork between engineering and manufacturing necessary for the economical assembly of such instruments in high volume, must also carry on into test. The finished instruments must perform well. The configuration of individual components must permit ease of test as well as ease of assembly. Many components, complete circuit boards, and entire chassis are tested automatically. Many parameters can be checked at one time; the results are read out as *go* or *no-go*, and the units are sorted, automatically and very rapidly. In some cases, the area of trouble in a reject circuit can be pinpointed automatically, thus simplifying the troubleshooter's job considerably.

AN ENGINEERING CHALLENGE

From the foregoing it is clear that many of the decisions leading to the continuing acceptance of RCA's home-instrument products are the responsibility of the engineer. Making a device that will perform a function is not enough. To successfully woo the demanding consumer, the engineer must use his acquired knowledge and inborn ingenuity to produce a product with maximum performance and reliability for minimum cost. In addition, if he is to work effectively he must communicate well with a variety of people, such as factory process and test technicians, technical writers, servicemen, merchandisers, *and his management.*

Fig. 3—Accelerated life tests under demanding conditions are often used to prove-in the ruggedness of receiver designs.

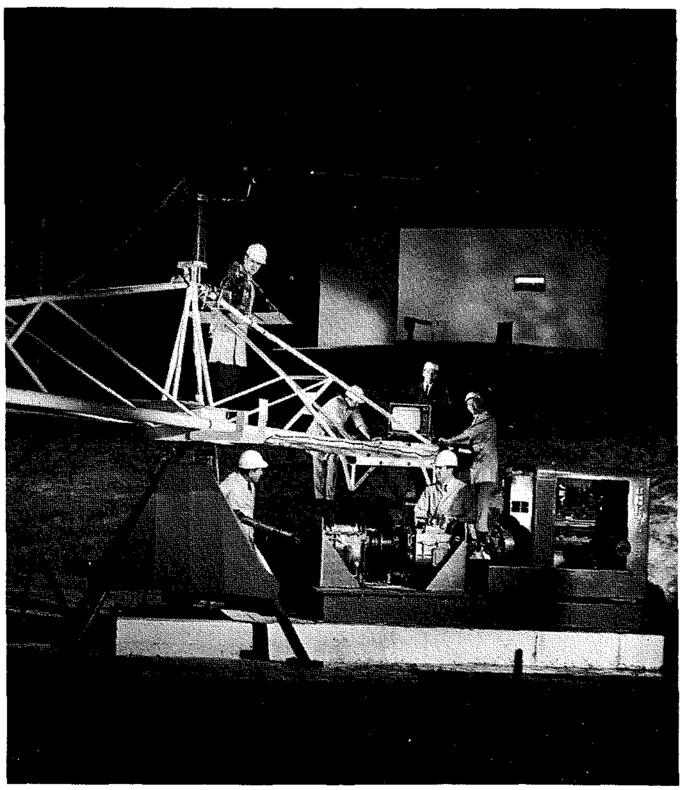
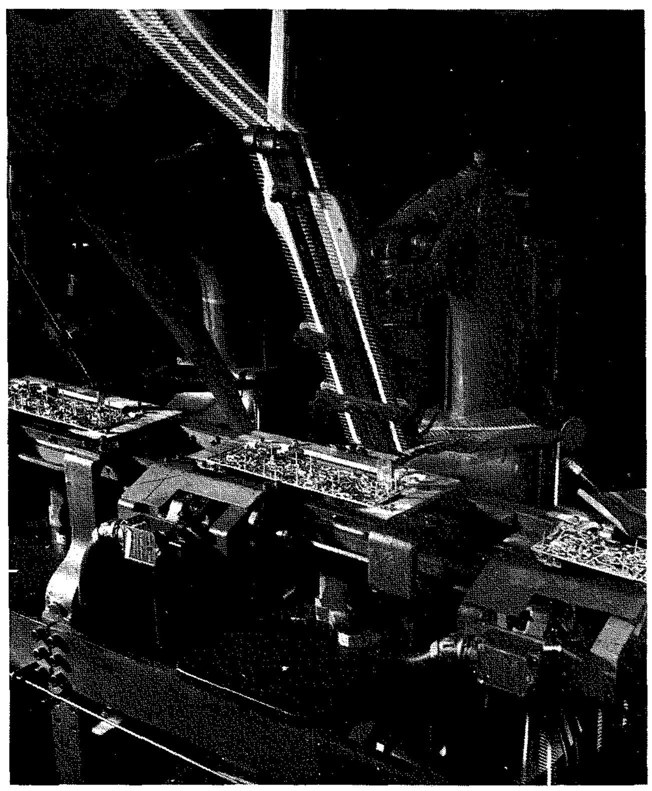


Fig. 4—Machine insertion of parts in etched-circuit boards helps keep assembly operations to a minimum.



RCA VICTOR HOME INSTRUMENT ENGINEERING PROGRESS

Engineering progress in radio and TV home instruments over the past 30 years has followed the pattern of gradual but steady growth described in this paper. The great emphasis placed on customer needs is discussed and several new technical features of modern instruments are described.

L. R. KIRKWOOD, Chief Engineer

RCA Victor Home Instruments Div., Indianapolis, Ind.

When one reviews the engineering progress in radio and television home instruments over the past 30 years, the gradual rate of change does not seem dramatic. But, a direct comparison of the first entertainment radios with the modern high-fidelity instruments of today provides a striking contrast. Even more dramatic, perhaps, would be a comparison of modern color television receivers with early radios; yet, modern color TV receivers sell at prices as low as those of many early deluxe console radios.

Such substantial progress has been achieved through a gradual uphill engineering and design battle with performance, reliability, cost, manufacturing, and service maintenance problems.

This paper describes some interesting technical changes and innovations and reviews the growth of radio and TV as major RCA products.

BIRTH OF THE TV INDUSTRY

In 1946 RCA introduced the 630-TS, a black-and-white TV receiver using a 10-inch picture tube. The plans, manufacturing processes, and test specifications were made available to many other manufacturers, providing major impetus to launching the television industry in the world. The design of this set was the result of RCA engineering knowledge and experience derived from an extensive development program dating back to the 1930's. A few pre-World War II receivers, such as the TRK12, were produced and much valuable information was gained. During World War II substantial advancements in the state of the TV art were made by RCA engineers in developing military TV applications such as Block III.

Since 1946 the development of TV receivers has followed several trends. Performance and operation have become much more critically tailored to the customers' requirements. New products have been introduced each year, and usually a new generation emerges every third year. Technically, the circuits and components show considerable refinement. New materials and devices have been applied, and, in some cases these have been created for particular applications. Fewer components and simpler circuits do a better job, last longer, and improve serviceability (Fig. 1). The resulting value enhancement is particularly noteworthy. This is illustrated further by comparing the selling price of $375 for the 630TS to the selling price of $150 for today's 19-inch receiver.

Final manuscript received February 7, 1966.

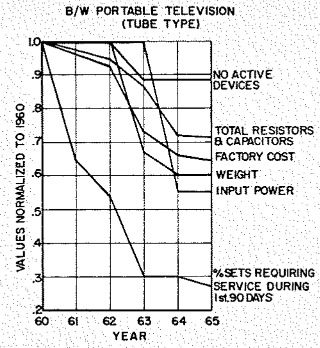


Fig. 1—Improvements in tube-type, monochrome, portable TV receivers, 1960-1965.

TV RECEIVER PROGRESS

The TV tuner, with its signal-sorting function, is perhaps the most critical part of a TV set, insofar as acceptable pictures are concerned.

Improvements in Tuners

Some notable improvements in TV tuners have been made in the last few years. Electrical and mechanical factors influencing oscillator stability have been refined to the point where "one-set" fine tuning is commonplace, with oscillator resetability typically accurate to within ± 50 kHz (kc/s) at 221 MHz (Mc/s) on Channel 7.

The application of the nuvistor and the frame-grid RF tubes now yields a channel-13 noise factor of 6.0 dB in comparison with the 13 dB figure of a few years ago.

All of our present UHF tuners now use a transistor oscillator, providing indefinitely long life where a major reliability problem previously existed. The smaller size and the cool running of the transistor allow internal mounting, which simplifies the radiation problem.

Fewer Stages and Improved Pictures

All other portions of the TV receivers, both black-and-white and color, have undergone similar growth, offering the customer more product for less money. Where four and five IF stages were deemed necessary some years ago, today's receivers offer good performance using two frame-grid IF stages aided by higher RF, mixer, and video gains. Noise immunity has been improved, and the application of voltage-dependent resistors has resulted in much-improved picture stabilization. Horizontal deflection circuits have benefited from higher efficiency yokes, new ferrites, and tubes with substantially increased power sensitivity. Solid-state power rectifiers are used throughout the line, and resetable circuit breakers have replaced fuses, resulting in fewer service calls.

Even the smaller wood cabinets have undergone a radical change in construction. Many are built and finished, then shipped to the point of instrument assembly in sections; these are then assembled around a die-cut mask which becomes a structural cabinet member.

TV Portables

By far the most noteworthy contribution in black-and-white TV in recent years has been the introduction of the 12-inch transistorized portable receiver. This set is a deluxe-performance, ultra-reliable unit that is currently enjoying an outstanding field service record. The basic circuitry of this receiver, under development for more than 5 years, will be used as a basis for many future solid-state designs, altering only those portions dictated by advances in device art, circuits, and picture-tube sizes. The second generation of this set incorporates an integrated circuit chip in the sound IF and detector circuits, replacing 26 components for improved reliability and economy.

MAINTENANCE IMPROVEMENTS

The color TV receiver has moved at a progressive rate toward a better, more reliable, and less expensive product that can be offered in numerous picture-tube sizes (Fig. 2). However, the prime target for color TV has been improved field acceptance, including greater ease of setup, operation, and maintenance. Color receivers now require only 20% more service than monochrome sets.

Simplified Setup and Servicing

To simplify installation of the color receivers, a setup switch has been added for ease of setting purity, color temperature, and tracking. The number of convergence controls has been substantially reduced, and control interaction has been minimized. A video peaking control varies picture sharpness.

Some new sets have automatic chroma control, automatic frequency control of the tuner oscillator, and even include detent tuning of UHF.

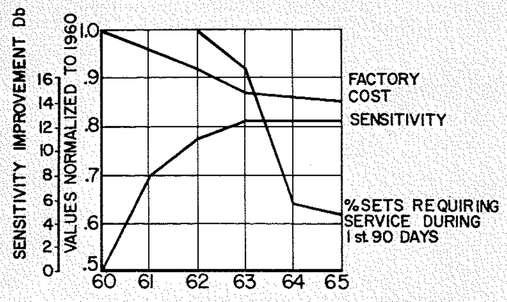


Fig. 2—Improvements in 21-inch, color TV receivers, 1960-1965.

LOREN R. KIRKWOOD received his BSEE from Kansas State University. He joined RCA in 1930. His work through 1941 included important contributions to radio and early high-fidelity equipment. Between 1941 and 1946 he engaged in the engineering of military communications equipment. From 1946 to 1950 he developed one of the first AC-DC, AM/FM home radios, and contributed to the 45-rpm record player. He directed receiver activities and development of all RCA color television demonstrations and field tests from 1950 to 1959. In 1951 he received the RCA Award of Merit for his work in color TV. Between 1959 and 1963 he was Manager, TV Product Engineering. He was named Chief Engineer, RCA Victor Home Instruments in 1963. Mr. Kirkwood is a Fellow of the IEEE.

To simplify the servicing of color receivers, current products use a minimum number of tube types. In the 19-inch color set, for example, 30% of all tubes are one type—the 6GH8 triode-pentode. To protect kinescopes and circuitry, built-in spark gaps are used extensively throughout the sets; resetable circuit breakers give added set protection.

Automatic Degaussing

A particularly noteworthy contribution that greatly improved customer acceptance of color TV was the introduction of automatic degaussing to the industry in 1964. This feature makes it unnecessary to call a serviceman to degauss the set when its orientation is changed in the home. The degaussing coils are energized every time the set is turned on. The necessary gradual decay of the coil-driving current is provided by the power-supply charging current; a thermistor and a voltage-dependent resistor provide the switching action which effectively takes the coils out of the circuit before complete set warmup. This action takes place before the picture appears, and there is no distraction to the viewer.

SOLID-STATE CIRCUITRY

Solid-state devices, with their advantages of high efficency and cool operation, will be used increasingly in future TV receivers. With solid-state circuitry, packaging requirements are considerably less demanding, ventilation is less of a problem, and, in some instances, greater latitude is afforded in selecting enclosure size and shape. Consequently, styling innovations may result, since the stylist has fewer restrictions he has greater freedom to create models and variations in accordance with consumers' desires.

More sophisticated use of integrated circuitry will be seen in the future, resulting in reduced labor and cost and increased reliability. Until a display-device breakthrough is made, deflection circuitry remains the problem area.

RADIO-VICTROLA PRODUCTS

Paced by advancing device technology, the application of transistors to high-volume home instrument products has grown to the point where virtually all RCA radios, high-fidelity, and tape products are fully transistorized.

Integrated Circuits and Transistors

The design trends that are developing as more thought is given to integrated solid-state circuit applications is a mild revolution in itself. One immediate outcome of such thinking has been the elimination of many large, expensive capacitors from radio and audio circuitry.

More extensive use of silicon transistors, with their inherent lower leakage currents and improved temperature characteristics, accounts for somewhat less-complicated and less-costly solid-state circuitry. It should be noted that even during the commercial birth of transistors, the all-tube radio-phonograph products were duplicated with transistors for the same or less money, with the exception of the highly polished designs of tube, line-operated, minimum dollar AM radio.

An engineering accomplishment that led to an extremely successful product was the design of the "Big D," a battery-operated, personal, portable AM radio. Although it operates from a 3-volt supply using two low-cost D-size cells, this set still provides RCA's *big-set sound*.

Advanced Styling

Many factors have contributed to the remarkable growth of home-instrument phonograph products, which show a 60% increase in RCA unit sales since 1960, compared to a 19% growth in industry unit sales. Advanced styling has contributed substantially to the growth of this product, and engineering advances have provided the consumer with a far better product, as well as many operational innovations, for less money.

Pickups with Better Performance

Phonograph pickups have gone through several generations of redesign for smaller size, compatibility with transistor amplifiers, and greatly improved high-frequency response. The new flip-over stylus assembly doubles the pickup compliance for improved tracking, and lower needle force results in less needle and record wear. It is virtually indestructible and provides an ease of service that is truly remarkable for such a critical part of the reproducing system. The latest pickup design has enabled engineering to provide high-quality ceramic elements throughout the stereo line, replacing crystal pickups in the minimum-dollar stereo portables. Pickup units with diamond stylii are now used where cost previously dictated the use of sapphires. New factory measuring techniques and equipment have been developed for the accurate and rapid measurement of pickup performance.

Improved Record Changer

Today's record changer, mass produced by RCA, represents many generations of design which have contributed toward making it a highly-styled, ultra-quiet and reliable unit.

Manual Portables Transistorized

The design of the old tube versions of the monophonic manual portable phonograph has undergone the careful scrutiny of many topnotch engineers—each doing his best to increase the performance/dollar of these high-volume units. Transistorizing such units posed a real design challenge. By using the phono motor for a line transformer and complementary-symmetry audio output units, significantly better performance was obtained at far less cost.

Stereo Consoles

The FM performance of our stereo consoles follows the basic standard of unsurpassed performance — even when compared with top deluxe component equipment. All FM front ends are built with precision molded coils, eliminating factory aligning of many circuits.

THE FUTURE

The radio-phono product, like the television product, will show an extensive use of computers for more comprehensive design studies. Tuned circuits requiring no manual adjustments will be used wherever feasible. Broadening the line to provide competitive products in all categories related to home entertainment should strengthen our share of the home instrument market even more, and provide an ever-increasing challenge to the design engineer.

A LOW-COST, HIGH-PERFORMANCE, 19-INCH COLOR TV RECEIVER

The design of the CTC-19 19-inch color TV receiver is the result of a concerted development program to produce a more compact color chassis with substantially the same performance as previous chassis, but at lower cost. The design objectives, chassis comparisons, board and chassis layouts, unique signal processing of video and sync-AGC and chroma circuitry are discussed in this paper. Performance is compared with the CTC-17, 25-inch color receiver.

L. A. COCHRAN and D. WILLIS
TV Product Engineering
RCA Victor Home Instruments Div., Indianapolis, Ind.

FOR several years RCA relied entirely on the performance of one basic color chassis to maintain its position as the leader in the industry. With the advent of the 25-inch, 90° picture tube, necessary deflection modifications in addition to new performance features were added to this basic chassis.

The increased competition in the color TV industry made it apparent that if RCA was to maintain its position of leadership, it would have to produce a new lower cost color chassis that would meet the performance standards set by previous chassis. The CTC-19 color chassis was designed with this goal in mind. Because of reduced size and weight, this chassis was readily adapted for use with the new 19-inch, 90° picture tube developed by the Electronic Components and Devices Division. The culmination of three years of design and development, RCA's new 19-inch color receiver now supplements its existing line of 21-inch and 25-inch color receivers.

CIRCUIT DESIGN

The areas of design that received the greatest attention were the IF, video, sync-AGC, and chroma circuitry. To achieve cost and performance objectives, the low-cost 6GH8A pentode-triode was used wherever possible; in all, seven 6GH8A's were used: one in the video circuitry, one in the sync-AGC circuitry, and five in the chroma processing circuitry.

No major design changes were made in the high-voltage and deflection circuitry; however, certain modifications were necessary due to the lower power requirements of the 19-inch kinescope and the use of new tube types. Board design and chassis layout in this area received considerable attention.

IF Circuits

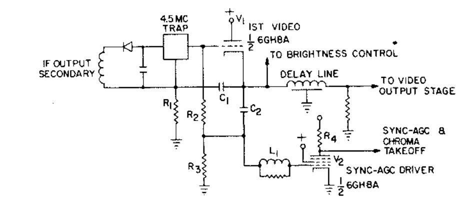
To lower costs, the IF circuitry for the 19-inch chassis was designed with two stages instead of three; to keep the gain of the IF as high as possible, several measures were taken. High-gain frame-grid tubes were used in the mixer and IF stages, and all interstage networks were double-tuned. The double-tuned circuits required an entirely different co-channel-sound-trapping arrangement than had been employed in previous chassis. A deep-absorption trap acts upon the secondary of the IF output transformer to keep co-channel sound from reaching the picture detector. The double-tuned output transformer with an absorption trap provides about 6 dB more gain than that of the single-tuned transformer with a null-type trap used in the three-stage IF amplifier. A shallow co-channel-sound-absorption trap acts upon the link transformer at the input of the IF. This trap serves two purposes: 1) it increases the sound rejection as seen at the second detector, and 2) it prevents cross modulation which would otherwise occur in the first IF stage. To provide alignment flexibility, a trimmer capacitor in the link circuit acts as an IF bandwidth control. No IF shielding is needed other than the shields for the interstage transformers. The nominal RF-IF sensitivity of this chassis is within 3 dB of previous chassis.

Final manuscript received January 12, 1966.

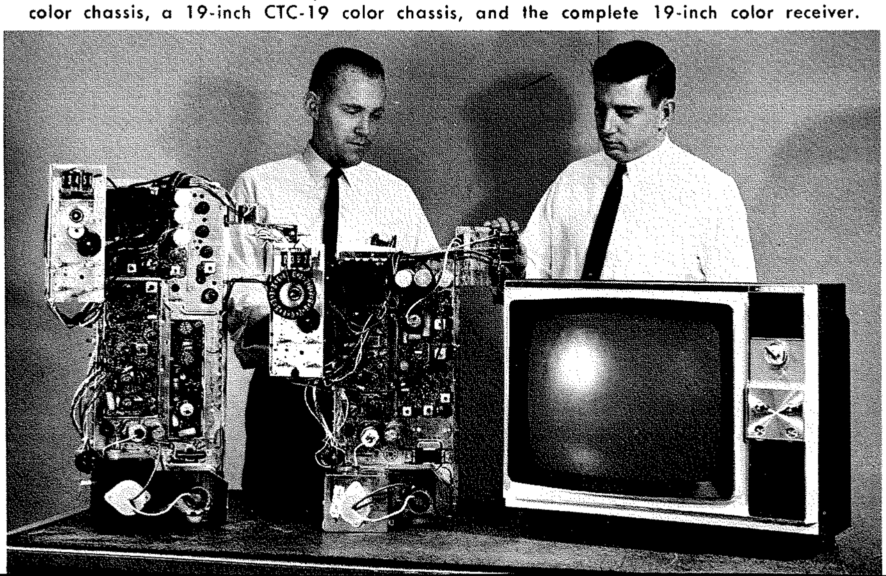
Video Circuits

The unique features of the video circuits in this chassis are in the first video and the sync-AGC driver stages. Both stages are in the same envelope, and they perform the same functions as the first and second stages in the three-stage video circuit. For low-frequency video information, V_1 acts as a cathode follower, thus providing a suitable source for driving the delay line (Fig. 1). At higher video frequencies, the detector is connected through C_1 directly across the grid and cathode of V_1, causing these frequencies to be accentuated at the cathode of V_1 and, therefore, at the delay line input. Sync and AGC drive is supplied to V_2 from the detector through divider R_1—R_3; chroma drive for V_2 is derived from the cathode of V_1 through C_2.

Fig. 1—Simplified schematic of CTC-19 1st video circuit.



The authors, L. A. Cochran (left) and D. H. Willis are shown with (l to r): a 25-inch CTC-17 color chassis, a 19-inch CTC-19 color chassis, and the complete 19-inch color receiver.



D. H. WILLIS was graduated with high honors from the University of Maryland in 1962 with a BSEE. He is presently doing graduate work at Purdue University. Following graduation Mr. Willis joined RCA as an engineering trainee and later was assigned to the Home Instruments Division. He is now engaged in the design of signal-processing circuitry of color television receivers. Mr. Willis is a member of Tau Beta Pi, Eta Kappa Nu, and Omicron Delta Kappa.

LARRY A. COCHRAN joined RCA as a summer student in June 1962. After receiving his BSEE from Purdue University in January 1963, he was assigned to the color television design group of the Home Instruments Division, Indianapolis, Indiana. He is presently working toward his MSEE at Purdue. Since joining RCA, Mr. Cochran has been associated with the design and development of many aspects of the color-signal-processing circuitry. He has been responsible for the design of new color synchronizing circuits and demodulation systems. Mr. Cochran is a member of Eta Kappa Nu and Tau Beta Pi.

Chroma frequencies are boosted further at the grid of V_2 by the resonance of L_1 with the input capacitance of V_2. As a result of these connections, the chroma output at the plate of V_2 is twice that obtained at the corresponding point in the previous chassis. A cost saving can be realized in the chroma processing circuits due to the higher chroma level at this point.

The circuit configuration of the video-output stage is basically unchanged. A pentode with higher g_m achieved sufficient kinescope drive with less video input; some of the component values were changed to accommodate the different output tube.

AGC-Sync Circuits

The automatic gain control (AGC) circuit was modified significantly. Most changes involved the networks feeding video information to the AGC tube and a new cathode-biasing circuit for this tube. Information is sent only to the AGC grid through a simple resistor divider network. The AGC control is a potentiometer that controls the amount of resistance division. When the bottom leg of the resistor divider is disconnected from ground, the AGC tube is turned *on*, cutting off the RF and IF stages. This action results in a blank raster which can be added as an extra position on the service switch for checking the purity of the receiver. The same cutoff action is desired when the service switch is in the collapsed sweep position used to adjust color temperature of the receiver.

The noise inversion for the sync separator is essentially unchanged; noise protection for the AGC function, however, is entirely different. The 6GH8A pentode does not have a separate tube pin for the suppressor grid as did the AGC tube in previous chassis. Since this electrode is not available for AGC noise inversion, sufficient noise protection was obtained by providing an equivalent cathode-degenerating resistance for the AGC tube.

The positive bias applied to the AGC cathode is derived so that it is affected by the B+ current drawn by the IF output stage; this was done to prevent *lockout*. When a lockout condition starts to occur, the increased B+ current drawn by the IF output stage lowers the positive bias on the AGC cathode, turning the AGC *on* and keeping it out of lockout. Although this method has been used with limited success in the past, it has now been applied to such a degree that AGC lockout is prevented under all conditions.

The AGC output filtering circuit has two additional components which provide a more nearly ideal frequency response characteristic. This low-pass response enables the AGC to handle airplane flutter while preventing the AGC from degrading horizontal pull-in performance.

The sync separator circuit is basically unchanged, except that component values were altered for compatibility with the different triode used.

Horizontal-Deflection and HV Circuits

Although the deflection circuitry is basically the same as in previous chassis, there have been some modifications that are worth noting. In the horizontal AFC circuitry, a feedback connection between the horizontal hold control and the cathode of the control tube has been added to improve circuit performance (Fig. 2). When the oscillator frequency is lowered by rotating the horizontal hold control toward minimum resistance, a positive voltage is developed on the cathode of the control tube causing the tube to cut off more quickly. Since the hold-in range depends on the control tube being biased in its active region, this arrangement limits the system hold-in range. Improved burst keying is realized by reducing the phase error which can exist between horizontal flyback and video synchronizing information. The addition of capacitor C_1 increases pull-in range by eliminating R_1 as a DC load on the phase detector. By increasing the pull-in hold-in ratio, the likelihood of the receiver remaining in horizontal synchronization when changing from channel to channel is increased.

The lower power requirements of the 19-inch kinescope permit the use of lower cost damper and horizontal-output tubes on this chassis. A new horizontal output tube (6KM6) developed by the Electronic Components and Devices Division utilizes cavity-trap-plate construction for controlling snivets (disturbances on the screen of the kinescope caused by oscillations in the horizontal output tube). Better circuit efficiency can be realized from this tube due to its high plate-to-screen ratio and its ability to deliver a high peak current at a low plate voltage. A new damper tube (6BS3) with lower power requirements has also been used on this chassis.

Chroma Circuitry

Major cost reductions have been realized in the chroma circuitry of the 19-inch chassis through the adoption of new circuit designs using low-cost pentode triodes. Performance goals match those of previous RCA color chassis.

The increased burst level available from the video circuits allows a lower g_m burst keyer to be used while maintaining the burst output necessary to synchronize the color oscillator. An open-loop subcarrier regeneration system eliminates the need for the killer detector, phase detector, and reactance tube normally found in automatic frequency and phase control (AFPC) oscillator systems. The 3.58-MHz (Mc/s) oscillator is pulled into phase synchronization by injection of the burst signal directly into the crystal portion of the oscillator circuit. By making the value of R_1 small (Fig. 3), the burst-drive circuit appears as a low impedance in series with the oscillator input, and the possibility of oscillator frequency drift is minimized. Trimmer capacitor C_1, at the grid of the oscillator, is used for oscillator tuning.

When a burst signal appears at the grid of the oscillator, a grid bias change of −4 volts develops and killer information can be derived directly from this point through the high impedance of R_2. The high selectivity of the crystal provides good killer noise immunity. The killer circuit consists of a grid-controlled rectifier biased by the information obtained from the oscillator grid.

Oscillator driver transformer T_2 supplies the low-impedance drive required for the color demodulators in this chassis. The network of R_3, C_3, and L_3 provides the leading phase shift needed for demodulation on the X and Z axis.

The color demodulator circuits used in the 19-inch chassis are of completely new design and are unique in the manner in which they are driven. Demodula-

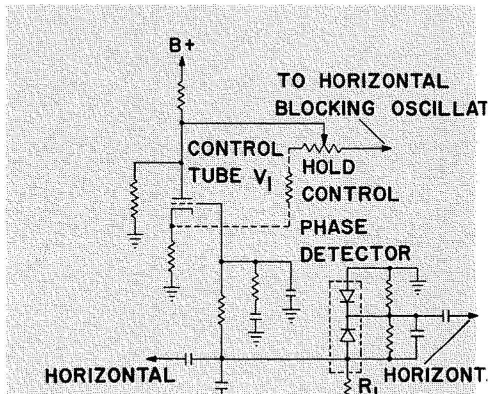


Fig. 2—Horizontal AFC circuit.

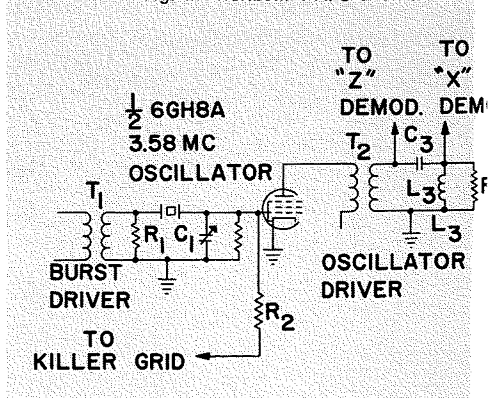


Fig. 3—Simplified 3.58-MHz oscillator circuit.

tion is achieved by driving the screen grids of a 6GH8A with a chrominance signal and the control grids with a reference signal from the 3.58-MHz oscillator. The desired demodulation characteristic, using the screen grid, can be obtained only if a DC potential in the order of 3 volts is applied to the screens. This operating point (Fig. 4) provides a linear output with a sufficient swing to drive the color difference amplifiers. Due to the low screen voltage, conduction occurs only when the peaks of the reference signal drive the control grid into its positive grid region. Spurious amplitude modulation on the reference signal is limited by the appropriate choice of grid-leak time constants. Due to a screen-grid g_m of 250 μS (umhos), a gain of only 2 is obtained from the demodulators; however, an overall gain of 30 to the kinescope grids is derived from the demodulator-difference amplifier combination.

The color difference amplifiers are the triode sections of three 6GH8A pentode triode units; the pentode sections are used for the X and Z demodulators and the color bandpass. The correct color difference signals are obtained, as in previous chassis, by matrixing in a cathode resistor common to all three difference amplifiers.

Since the difference amplifiers are DC coupled to the kinescope grids, it was necessary to insure adequate DC stability. First, heavy DC feedback from plate to grid was obtained by means of R_F (Fig. 5). Then the DC gain of the difference amplifiers was divided down to 50% of the AC gain by resistors R_1 and R_2. The kinescope biasing arrangement consists of appropriately chosen voltages derived from a B+ divider string and applied to the kinescope grid through R_2. The impedance of the kinescope bias arrangement is made small compared to R_2 so that: 1) variations in bias have a negligible effect on DC gain, and 2) crosstalk is minimized.

CTC-17 AND CTC-19 COMPARISON

The more important characteristics of the CTC-19 and CTC-17 chassis are compared in Table I. The chassis layout of the CTC-19 is shown in Fig. 6. Since fewer tube envelopes are used in the CTC-19, power consumption has been reduced by 12% under that of the CTC-17. To achieve cooler operation, the high-voltage enclosure in the CTC-19 has been rotated 90° and a shield has been added between the shunt regulator and high-voltage rectifier. Also, the horizontal output tube has been moved farther away from the high-voltage enclosure to further reduce the ambient temperature of the flyback transformer. Video and chroma drive available at the kinescope grids and cathodes are identical in both chassis. Since the kinescope screen voltages on both chassis are derived from the *B-boosted boost* supply, identical contrast levels are achieved. Through the use of frame-grid mixer and IF tubes, the overall sensitivity from the antenna to the kinescope in the CTC-19 is very close to that of the CTC-17. Although the design limitations on the color synchronizing circuits in the CTC-19 were more stringent than in the CTC-17, commercially acceptable color performance has been realized.

TABLE 1 — Comparison of CTC-17 and CTC-19 Chassis

Characteristic	CTC-17	CTC-19
Power consumption	330 W	290 W
Weight	31.5 lbs.	26.0 lbs.
Size	10⅝"x26"	9¾"x20½"
No. of tubes	26	21
Sensitivity (input for 1-volt DC output at 2nd detector)	8.5 uV	9.5 uV
Video drive	150 V	150 V
Kinescope anode voltage	25 kV	24 kV
Sound sensitivity	4.3 uV	4.7 uV
B+ supply voltage	405 V	405 V
Anode power	36 W	28 W

ACKNOWLEDGEMENTS

The authors gratefully acknowledge the contributions of J. C. Marsh (IF design) and J. N. Pratt and N. W. Hursh (horizontal deflection and high voltage). Acknowledgement is also due J. A. Konkel, Leader, and G. E. Kelly, Manager, for their suggestions and support throughout.

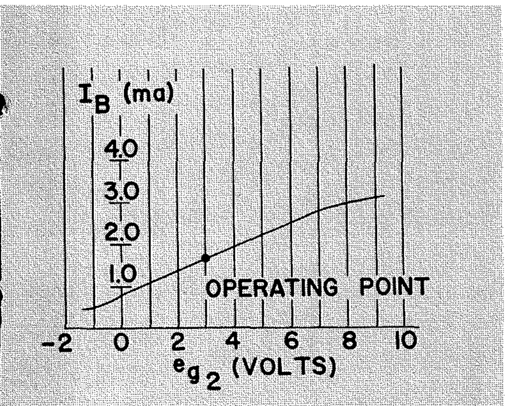


Fig. 4—Typical demodulator characteristic.

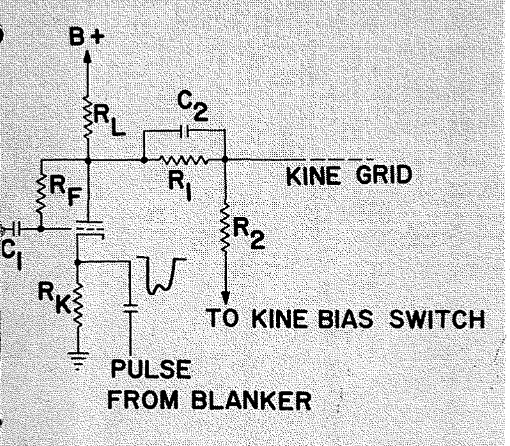


Fig. 5—Simplified difference amplifier.

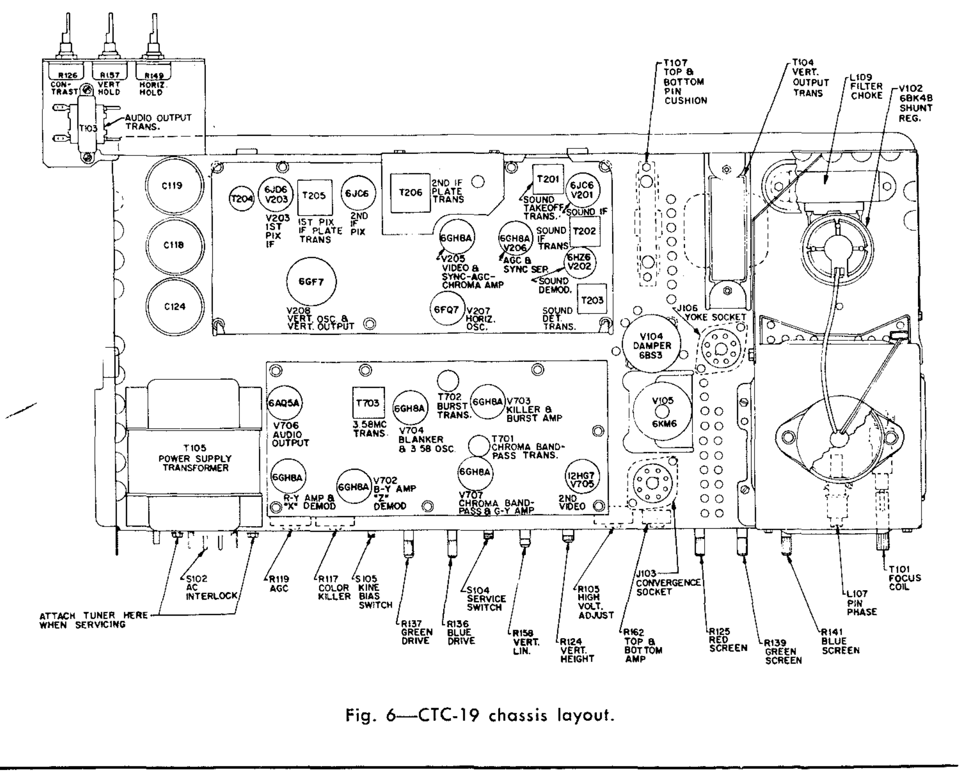


Fig. 6—CTC-19 chassis layout.

THE 25-INCH RECTANGULAR COLOR TV RECEIVER

This paper describes the design of the CTC-17 color TV receiver. Although this receiver has a 25-inch rectangular picture tube, the cabinet is smaller than that of its 21-inch predecessor. Important circuit considerations, such as high-voltage design, horizontal deflection, convergence, color purity and correction, degaussing, and component design, are discussed.

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PRIOR to 1964, color television receivers employed a 21-inch, 70°-deflection, round, color picture tube. These receivers performed well, but there was a growing demand for a more natural, rectangular viewing area and a smaller, less bulky cabinet. To meet this demand, a 25-inch, 90°-deflection color tube having a rectangular picture-viewing area was developed. The rectangular viewing area provided a more conventional, more pleasing presentation and a larger picture area. The new cabinet designs were approximately 4½ inches shallower than the cabinets housing the 21-inch 70° tube.

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GENERAL DESIGN GOALS: 25-INCH COLOR TV RECEIVER

The new RCA 25-inch rectangular color TV receiver (CTC-17) employs many of the basic signal circuits used in previous RCA color TV receivers. The main areas of redesign involved the deflection circuits and related functions. Redesign was dictated by the greater 90°-deflection angle, the larger picture area, a smaller neck diameter, and a new gun design.

The basic circuit functions and interconnections are shown in Fig. 1; arrows indicate the direction of flow of information through the receiver. Except for the audio signal, all of the information flow ends up at the picture tube. Discounting heaters, 20 different electrical or electromagnetic signals and 5 permanent-magnet controls are applied to the picture tube (Table I).

Of the 25 primary signal and control functions listed in Table I, the last 15 required varying degrees of redesign and development for application to the 25-inch rectangular color tube. Circuit design and operation of these functions are discussed in subsequent sections.

Picture tube considerations that required new circuit designs and some new control functions were: 1) a wider deflection angle, 2) new gun design and smaller diameter neck, and 3) a larger picture area. This new tube configuration greatly affected the deflection yoke design and the circuits driving the yoke, and the yoke-picture tube combination affected the dynamic convergence requirements.

The new, smaller diameter neck and new, smaller gun design affected the

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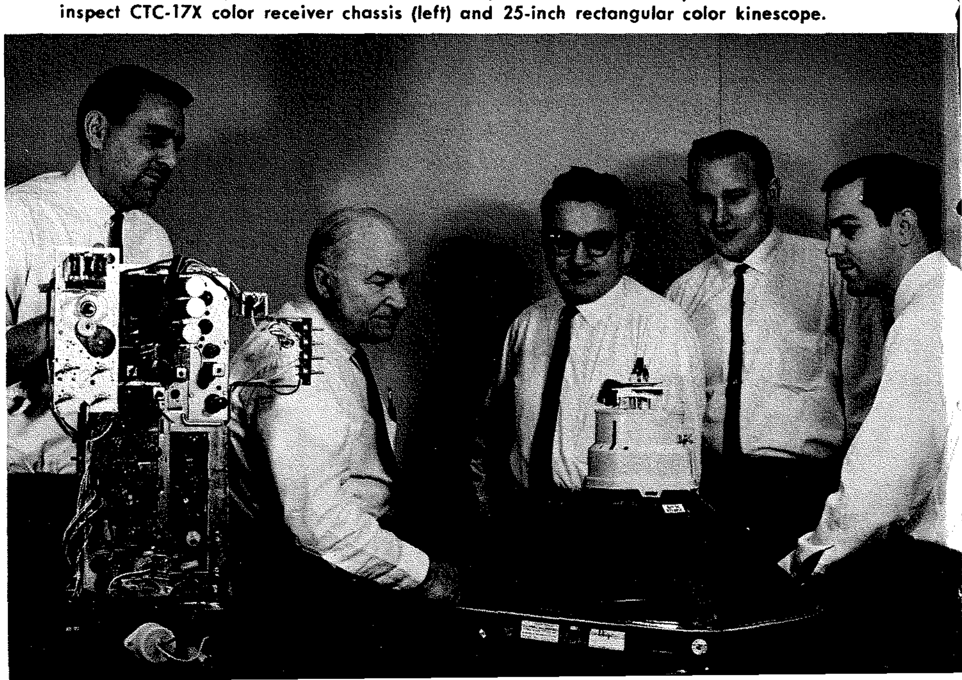
RONALD R. NORLEY received his BSEE degree from the University of Maryland in 1962. Following graduation he spent six months in the United States Army. On completion of this tour of duty he joined RCA as a specialized trainee. Later he joined The Home Instruments Division, where he has been involved in the circuit design and development of color receivers. He is currently specializing in applied magnetics. Mr. Norley is a member of Eta Kappa Nu and IEEE and is a recipient of the "Certificate of Distinguished Scholarship" from the University of Maryland.

PHILIP G. McCABE was graduated with honors from West Virginia University in 1960 with a BSEE degree. He joined RCA the same year as a specialized trainee. After completing the training program he was assigned to the Black and White TV Design and Development group of the RCA Victor Home Instruments Division, Indianapolis, where he worked on television deflection circuits. Since 1962, he has been involved with deflection and correction circuitry in color TV design. Mr. McCabe is a member of Eta Kappa Nu and Tau Beta Pi.

JOHN STARK, JR. received his BSEE degree from the University of Wisconsin in 1950. Following graduation he joined the RCA Training Program and was assigned to the Color Product Development Section of Home Instruments in Camden, N.J. Since 1961 he has been a Group Leader in the Color Television Section of Home Instruments at Indianapolis, Indiana. He has done circuit design in practically all areas of the color TV receiver. Mr. Stark holds a number of patents and joint patents in various areas. He is a member of IEEE, Tau Beta Phi, Eta Kappa Nu, and Phi Eta Sigma.

EUGENE LEMKE graduated from Tri-State College in 1955. He joined the RCA Home Instruments Div. the same year as a specialized trainee and was later assigned to the Indianapolis resident engineering group. There he was involved in RF and IF circuit and transducer engineering. In 1960 he joined the product design and development group, where he was responsible for the design of semiconductor remote control systems for color and monochrome TV. Later he was assigned to the color TV group, where he has been engaged in the development and design of specialized drive and correction devices and circuits associated with high-deflection-angle color kinescopes.

The authors (l to r) P. G. McCabe, B. E. Denton, J. Stark, R. R. Norley, and E. Lemke inspect CTC-17X color receiver chassis (left) and 25-inch rectangular color kinescope.



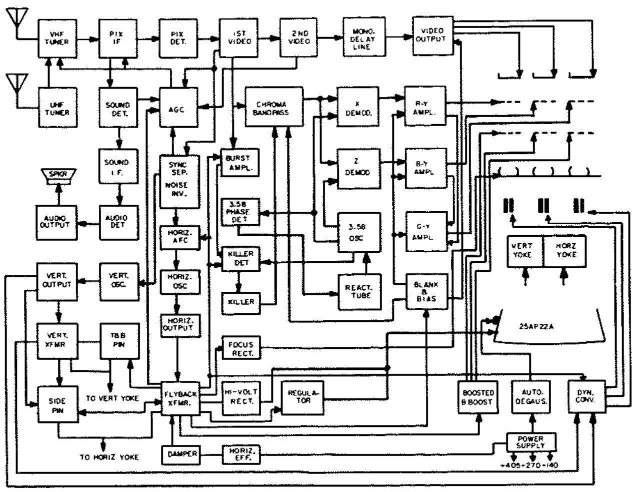


Fig. 1—Block diagram of main circuit stages of the CTC-17, 25-inch color TV receiver.

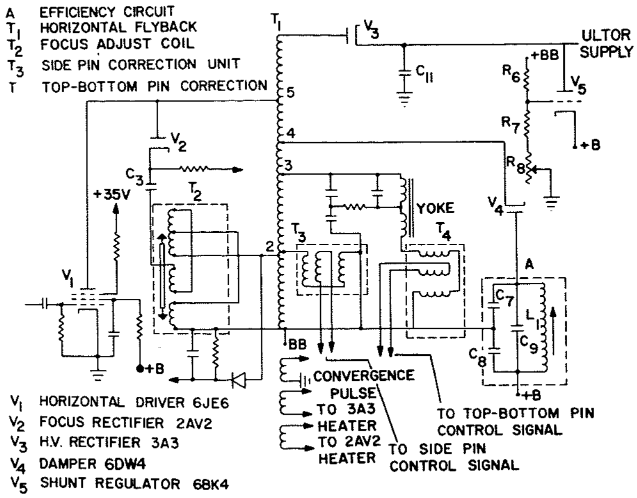


Fig. 2—CTC-17 horizontal and high-voltage circuit.

mechanical designs of yoke and pole-piece exciters as well as the electrical circuit design. Static convergence, blue-lateral, and neck-purity devices required new approaches and new mechanical designs. Other design considerations included: 1) a greater current demand on the high-voltage system because of the larger picture area, 2) magnetic shielding, and 3) degaussing problems arising from the wider angle of deflection.

Yoke and picture-tube design for optimum convergence and purity resulted in a raster with *pincushion distortion.* This distortion was corrected by additional circuitry to modulate horizontal and vertical deflection.

HORIZONTAL DEFLECTION AND HIGH-VOLTAGE DESIGN

In designing the horizontal deflection and high-voltage circuits (Fig. 2) for the CTC-17 chassis, the following factors were taken into consideration: 1) picture tube (25AP22A) basic requirements; 2) flyback transformer requirements; 3) performance; 4) auxiliary circuits; 5) features; and 6) precautions. These factors are discussed briefly in subsequent paragraphs.

High-Voltage Circuits

The maximum high voltage at a high-line-voltage condition (130 volts) for the 90° color picture tube (25AP22A) is 27,500 volts. To operate within these limits, it was decided to design for a value of 25 kV at a normal line voltage of 120 volts. It was further decided that the high voltage should be regulated to approximately 25 kV for all values of beam current between zero and 1400 μA to help provide the desired brightness, spot size, and highlight for overall bright scenes. The anode capacity of the picture tube provides enough stored energy to supply the beam current required for small-area highlights. The design specifications of the 25AP22A call for a focus range of 16.8 to 20% of the ultor supply voltage, or between 4200 and 5000 volts DC. This focus supply must provide for a leakage current from the focus anode of minus 45 μA to plus 15 μA. A screen supply in excess of 1000 volts DC, with respect to ground, is required.

Deflection Circuit

A flyback auto-transformer used with the proper deflection yoke produces full deflection at low line voltage (108 volts AC) (Fig. 2). The combination has a natural resonant frequency of approximately 40 kHz (kc/s), which provides a retrace time of approximately 12.5 μsec. A high-voltage winding with leakage inductance tuned to approximately the third harmonic of the retrace pulse (to minimize ringing) must be capable of providing a retrace pulse of sufficient amplitude and impedance for the desired high voltage at specified beam current. Under maximum load and normal operating conditions, the temperature of the flyback transformer must not exceed 100°C. Several taps must be provided on the transformer for such auxiliary functions as: focus adjust, pincushion correction, boosted B-boost voltage, burst keying, blanking, and AGC. A special winding on the core of the transformer opposite the high-voltage winding provides a convergence pulse referenced to ground. Provisions are made for supplying heater power for the 3A3 high-voltage rectifier and the 2AV2 focus rectifier.

Performance

With bogie components, the regulation of the high-voltage supply is approxi-

TABLE 1—Basic Control Elements and Signal Functions

Control Element	*Signal Function*	*Control Element*	*Signal Function*
Red cathode	Red monochrome drive (fixed) and vertical retrace blanking	Red internal pole pieces	Static radial convergence correction of red beam
Green cathode	Green monochrome drive (variable, to adjust for phosphor efficiencies) and vertical retrace blanking	Green internal pole pieces	Horizontal dynamic convergence correction of green beam
		Green internal pole pieces	Vertical dynamic convergence correction of green beam
Blue cathode	Blue monochrome drive (variable, to adjust for phosphor efficiencies) and vertical retrace blanking	Green internal pole pieces	Static radial convergence correction of green beam
Red grid	-(R-Y) chroma drive and horizontal retrace blanking	Blue internal pole pieces	Horizontal dynamic convergence correction of blue beam
Green grid	-(G-Y) chroma drive and horizontal retrace blanking	Blue internal pole pieces	Vertical dynamic convergence correction of blue beam
Blue grid	-(B-Y) chroma drive and horizontal retrace blanking	Blue internal pole pieces	Static radial convergence correction of blue beam
Red screen grid	Red gun cutoff adjustment	Blue lateral device	Static lateral convergence correction of blue beam
Green screen grid	Green gun cutoff adjustment		
Blue screen grid	Blue gun cutoff adjustment		
Focus electrode	Common focus potential for all three guns (variable)	Neck purity device	Permanent magnet predeflection spot landing correction (all three beams)
Ultor	Final high voltage (adjustable, regulated)	Degaussing coils	Earth and stray magnetic field neutralization
Red internal pole pieces	Horizontal dynamic convergence correction of red beam	Vertical deflecting coils	Vertical scanning of beams from a specific deflection center
Red internal pole pieces	Vertical dynamic convergence correction of red beam	Horizontal deflecting coil	Horizontal scanning of beams from a specific deflection center

mately flat from zero to 1400 μA, or more; at least 35 watts of picture power is produced at normal AC line voltage, and full deflection is provided at 102 volts AC line. The linearity deviation on either side of center is controlled to less than 5% with the adjustment of the efficiency circuit. When the efficiency circuit ("A" in Fig. 2) is properly adjusted, 39 to 40% of the horizontal-driver-tube input plate power is converted into ultor power (35 watts or more under bogie conditions).

SPECIAL AUXILIARY CIRCUITS

Focusing

The focus voltage is set to the correct value by adjusting the *focus-adjust* coil. The focus voltage output is proportional to the algebraic difference between the plate pulse of the horizontal driver stage and the output pulse from the focus adjust coil. When the coil is turned fully clockwise, the output pulse from the focus-adjust coil will be positive and the rectified voltage appearing across C_3 will be a minimum of 4100 volts. If the coil is turned fully counterclockwise, the output pulse from the focus-adjust coil will be negative and the rectified focus voltage appearing across C_3 will be a maximum of 5300 volts. This focus range is greater than that specified above, but it was necessary for certain combinations of tolerance used in the components.

Efficiency Circuit

The efficiency circuit ("A" in Fig. 2), consisting of C_7, C_8, C_9, and L_1, is a very flexible circuit which provides a waveform in series with +B which tends to compensate, in part, for the voltage drop across the damper and other components in the circuit. When this waveform is of the proper shape (and since horizontal input power is constant), the characteristic stretch on the left side of the picture tube is practically eliminated. Capacitors C_7 and C_8 together constitute the B-boost capacitors; the values of C_9 and L_1 may be altered to adjust the operating impedance of the efficiency circuit. When the capacitance ratio of C_7 to C_8 is increased, the input power to the horizontal driver stage will be decreased. If the capacitance ratio is decreased, the input power to the horizontal driver circuit will be increased. If the ratio of C_7 to C_8 is held constant and both capacitors are decreased in value, the B-boost voltage will increase a small amount and the center of the picture will be stretched. Conversely, if the ratio is held constant and the value of each capacitor is increased, the B-boost voltage will decrease a small amount and the center portion of the picture will be compressed. The capacitors are critical enough to require the use of selected pairs rather than low-tolerance components.

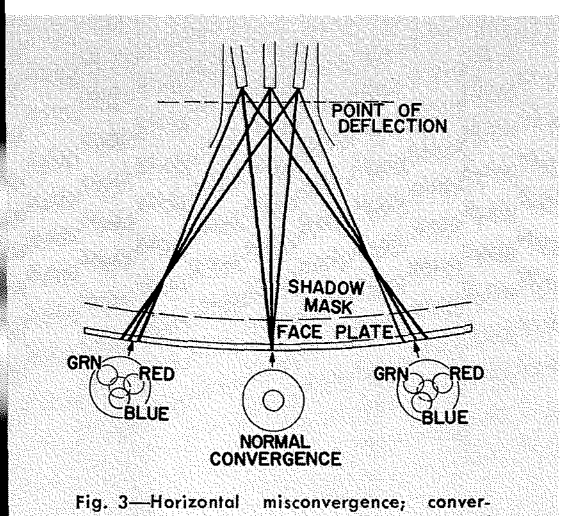


Fig. 3—Horizontal misconvergence; convergence errors due to horizontal deflection of electron beams.

Shunt Regulator Circuit

A very-high-mu triode (V_5), capable of dissipating approximately 40 watts of power and operating at very high plate voltages, functions as a shunt load across the high-voltage output (Fig. 2). Resistors R_6, R_7, and R_8 form a bleeder string between B-boost and ground; R_6 and R_7 are a matched pair. Their values are such that, when R_8 is adjusted to approximately midpoint and the cathode of the shunt regulator tube (V_5) is connected to +B, the control grid of V_5 is biased, causing the tube to load and hold the ultor supply voltage at 25 kV at zero beam current. When the picture tube is adjusted to draw beam current, the increased load on the ultor supply causes the value of B-boost voltage to decrease; at the same time, the bias on V_5 increases, thus increasing the plate impedance of V_5, which decreases its loading on the ultor supply. Thus the picture tube load and the shunt regulator load combine to form a constant load on the ultor supply resulting in constant high voltage over the range of power capabilities.

Precautions have been taken to confine x-rays or radiation associated with the regulator tube. As further precaution, the flyback enclosure has been made tight so that it will not sustain combustion should arcing occur in the compartment.

CONVERGENCE

In a black-and-white picture tube, one electron gun is mounted concentric with the axis of the tube; electrons emitted from this gun strike a phosphor screen, producing viewable pictures. In a shadow-mask color picture tube, three separate electron guns are mounted in a triad around the axis of the tube. The forward ends of these electron guns are tilted slightly toward the axis of the tube so that the electrons from the guns coincide at the center of the faceplate.

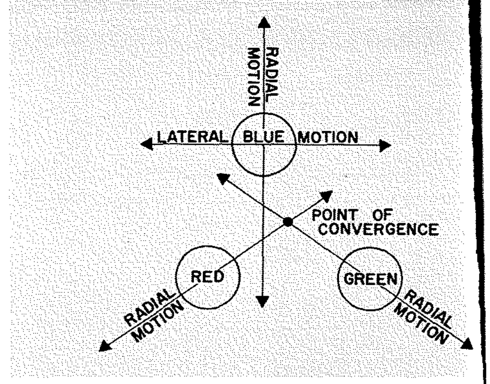


Fig. 4—Center convergence; radial motion accomplished by DC magnets in pole piece exciters, lateral motion accomplished by blue lateral device.

The phosphor screen is not composed of one continuous coating of phosphor, but is made up of thousands of symmetrical dot trios of three separate phosphors. Each phosphor is capable of emitting only one color, either red, green, or blue, when activated by electron bombardment. Directly behind the phosphor screen, at a fixed distance, is the shadow mask. This shadow mask is made up of a multitude of tiny holes through which the electron beams must pass to strike the phosphors on the faceplate. When the electron beams from the three guns strike the same point on the phosphor screen the electron beams are said to be *converged*.

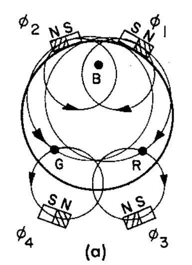
When the electron beams are deflected away from the center of the screen, it can be seen (Fig. 3) that the distances from the electron guns to the screen are not the same. Therefore, the point where the beams coincide does not occur at the phosphor screen and misconvergence occurs. Under these conditions, only one point on the screen is converged, that being the undeflected point or the point on the screen where the electron beams ideally coincide. Referring to Fig. 3, it is seen that the further away from the center of the screen the beams are deflected, the larger becomes the separation or misconvergence of the beams. Thus, any increase in deflection angle of a shadow-mask color picture tube is accompanied by an increase in convergence errors as the beams travel away from the center of the phosphor screen. Partial compensation for this error is obtained in the 25-inch, 90°, shadow-mask color picture tube by the use of a smaller neck diameter, which permits the electron guns to be mounted closer to the tube axis; hence, the angle of tilt of each electron gun with respect to the tube axis is necessarily decreased.

To correct for the convergence errors, as the beams are deflected on the screen, convergence pole-piece exciters are mounted external to the tube and are driven by dynamic waveforms at both horizontal and vertical scan rates. The three beams are deflected radially and independently of each other prior to entering the deflection field of the yoke. The convergence circuitry is designed for correction along both the horizontal and vertical axes. Any point on the screen that is not on the vertical or horizontal axis will require the addition of horizontal and vertical waveforms to achieve convergence. The addition of both correction waveforms is accomplished in the pole-piece exciters. The above discussion assumes that the slight tilt of the electron guns is sufficient for the beams to converge at the center of the screen. In practice, however, this is not the case, for the manufacturing variations of picture tube and deflecting yokes must be considered.

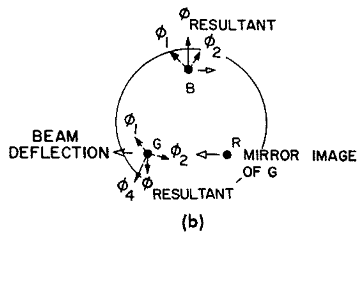
To enable the electron beams to converge at the center of the screen, four degrees of movement must be supplied (Fig. 4). The radial movement of the three electron beams is accomplished by adjustable permanent magnets mounted in the convergence pole-piece exciters to control the flux through the internal pole pieces of the picture tube. This system has the advantage of directing the magnetic flux from the external pole pieces to the internal pole pieces with minimum distortion of the electron beams; also, a more sturdy arrangement for the permanent magnets is provided, with less chance of misalignment.

The fourth degree of movement is supplied primarily to the blue beam, and this movement is perpendicular to the radial movement of the blue beam. In previous color picture tubes of larger neck diameters, the blue-lateral motion was obtained with little difficulty by

Fig. 5a—Abbreviated schematic flux plot due to magnets, and b) resultant flux action on each beam and corresponding direction of beam deflection.



(a)



(b)

using an internal magnetic strap. This strap directed a magnetic flux field perpendicular to the blue electron beam and perpendicular to the horizontal center line of the tube. The magnetic field was supplied by a permanent magnet mounted on the neck of the tube adjacent to the strap; the blue beam could be moved horizontally by varying the effective strength of the magnet.

In the smaller neck (90° color picture tube), the internal magnetic strap is omitted, because the nearness of internal components make it undesirable from an interaction standpoint. To provide the lateral motion required for the blue beam, and at the same time prevent adverse action on the red and green beams, a device incorporating four small barium ferrite magnets was devised. These four magnets are positioned around the neck of the tube in such a manner that the resultant magnetic flux acting on each electron beam is perpendicular to the horizontal center line of the picture tube, thus deflecting the electron beams horizontally. The resultant flux (Fig. 5) at the blue beam is opposite the resultant flux at both the red and green beams; thus, blue-beam deflection is opposite that of the red and green beams. This permits convergence of the blue beam to the red and green beams more readily than if only the blue beam were deflected. The strength of the resultant flux is varied by simultaneously moving the four magnets vertically away from the neck of the tube. The resultant flux fields can be completely reversed by reversing the positions of magnets 1 and 2 with 3 and 4. Since the bottom two magnets (Fig. 5) are further from the neck of the tube than the top two magnets, the resultant flux field at the red and green beams is substantially less than the flux field at the blue beam; consequently the most significant movement is the lateral movement of the blue beam. Although movement of the red and green beams in the opposite direction to blue-beam movement is not detrimental, the ease of setup is enhanced by permitting only one beam a large degree of movement. The relative beam movement of blue to red and green under these conditions is 4 or 5:1.

PURITY

When the beams from the three electron guns strike their respective phosphors (e.g. the electron beam from the red gun strikes only the red light-emitting phosphor), this condition is known as *purity*.

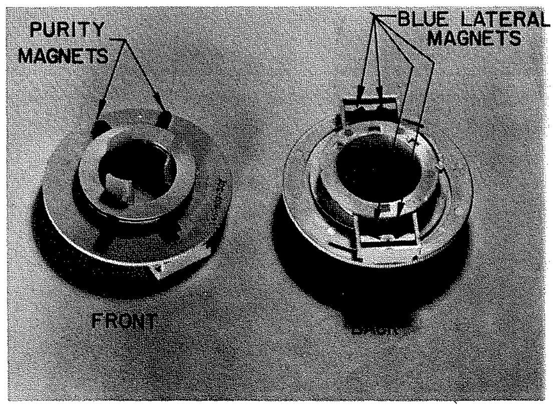
Purity is attained by accurately designing the deflection yoke and picture tube so that the beams passing through the shadow mask land on their respective phosphors. However, to compensate for manufacturing variations, an additional corrective device must be employed. This device is a purifying magnet around the neck of the tube in the vicinity of the cathodes with its field perpendicular to the axis of the tube. This field will deflect the electron beams and permit alignment of these beams on their respective phosphors. The strength of the purifying device must be variable from a minimum condition, where little or no correction is required, to a condition of maximum correction or 5 mils of beam register movement. Also, the control of the electron beams must be variable in direction as well as in magnitude. These conditions are met by providing two ring magnets similar to the 70° purity magnets but having an elliptical inner configuration and a circular outer configuration. When these ring magnets are magnetized across the minor axis, the resultant flux field perpendicular to the axis of the picture tube is essentially straight and uniform. The resultant strength can be controlled by rotating one magnet with respect to the other, and the direction of the resultant field can be controlled by rotating both magnets together around the neck of the tube. The configuration of these purity magnets permits the use of a very economical method of magnetizing these rings.

The purity magnets and the blue-lateral device are incorporated into one assembly to assure correct positioning of the two devices, with respect to each other, at all times (Fig. 6).

PURITY ERRORS AND CORRECTION

Correct purity can be obtained with the devices just mentioned, but any other magnetic field that crosses the path of the electron beams will further deflect the beams. This added deflection can cause the electrons to strike the wrong phosphors and thus produce color impurities.

Fig. 6—Purity and blue lateral neck device.



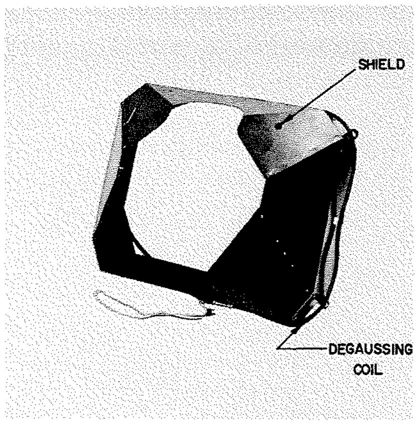


Fig. 7—Magnetic shield and degaussing coil for the 25-inch rectangular tube.

Magnetic fields that may cause purity errors include the following:

1) Magnetic fields radiated by components in the television receiver.
2) Unusual magnetic field conditions, external to the television receiver, that may induce a magnetic field into the metal parts of the picture tube.
3) The earth's magnetic field.

The 25-inch, 90° color picture tube did not present any new considerations in the first two causes of purity errors. However, in protecting the 25-inch, 90° color picture tube from purity errors caused by the earth's magnetic field, some problems were encountered that were not experienced with the 21-inch 70° color picture tube.

Effect of Earth's Magnetic Field

In considering the effect of the earth's magnetic field on purity, it is convenient to represent the magnetic field as consisting of a vertical and a horizontal component. The vertical component of the earth's magnetic field will cause an added deflection of the electron beams in a horizontal direction. The amount of added deflection will depend on the geographical area in which the TV receiver is used. The added deflection caused by the horizontal component of the earth's magnetic field will depend not only on the geographical area, but also on the orientation of the TV receiver in a given geographical area.[1] Improved shielding from the earth's magnetic field is required for the 25-inch, 90° color tube, as compared to the 21-inch, 70° color tube, due to the increased picture area of the 25-inch picture tube.

Increased Picture Area

When the color receiver is oriented north or south, the horizontal component of the earth's field affects the electron beam due to the radial distance the beam travels away from the axis of the tube. Compared to a zero magnetic field condition, the electron beams are moved tangentially with respect to the center of the tube. The largest tangential movement occurs at the outer edges of the picture tube with no beam movement at the center of the tube where the electron beams are parallel with the horizontal component of the earth's magnetic field. Due to built-in tolerances in the 21-inch, 70° color tube the tangential movements caused by a north or south orientation were not great enough to cause serious visible color errors. However, the increased viewing area of the 25-inch, 90° color tube increases the radial distance the electron beams travel. Therefore, greater tangential movements of the electron beams are encountered. These tangential movements are great enough to require definite shielding correction.

The need for maximum shielding of the 25-inch, 90° color picture tube from the horizontal component when the color receiver is facing north or south dictates a magnetic shield design of the basic form shown in Fig. 7. This shield also minimizes the beam deflection caused by the vertical component of the earth's magnetic field, and less compensation is needed.

DEGAUSSING

Degaussing improves the shielding properties of the magnetic shield, which requires degaussing after any change in external magnetic field conditions. Degaussing also eliminates the purity error caused by unusual temporary magnetic field conditions (gaussing of the picture tube).

Degaussing is achieved by driving a coil with an AC current which is large initially and decays to zero; current passing through the coil creates a magnetic flux field that accomplishes the degaussing. The coil is mounted on the shield so that the shield and shadow-mask-frame assembly properly direct the degaussing flux. The same degaussing methods are used in both the 21-inch, 70° color receiver and the 25-inch, 90° color receiver.

SHIELD MATERIAL

A low-grade silicon steel is used for the magnetic shield of the 25-inch color receiver. Previously, low-carbon steel was often used as a magnetic shielding material for color picture tubes. However, thin-gauge silicon steel provides better shielding than low-carbon steel of almost twice the thickness. The reduced thickness of the silicon-steel shielding results in a weight saving and keeps total material cost about equal. Silicon steel is also easier to degauss, which permits the use of a less expensive degaussing coil.

DEFLECTION YOKE

The design of a deflection yoke for use with a shadow-mask picture tube involves numerous performance considerations; but, on a very broad plane, these considerations can be abbreviated under the headings of convergence, register, and pincushion. Broadly stated, any two of these considerations can be optimized at the sacrifice of the third. Because of the increased deflection angle of the 90° rectangular picture tube and because of the addition of corners, which were not present with the round 70° shadow-mask picture tube, a yoke design was chosen for the CTC-17 that would optimize convergence and register. Resulting pincushion errors could be corrected by circuitry external to the deflection yoke.

The resultant yoke design provided very satisfactory convergence and register without requiring external correction circuitry for corner convergence. Also, with the present design there is no need for a dynamic blue-lateral circuit to make the blue beam converge along the horizontal axis. Convergence of the blue beam is accomplished with only the application of radial convergence.

It should be emphasized that there must be close cooperation between the yoke and tube design engineers before either device is finalized and a finished product is realized.

PINCUSHION

For reasons indicated above, the 90° color picture tube places more stringent requirements on the deflection yoke. At the present state of the art, it is necessary to compromise pincushion raster distortion, for optimization of purity and register, to such a degree that external correction circuits are required to produce an acceptable square raster. The established method of correction with the fixed permanent magnets used in black-and-white TV receivers is not acceptable for a color picture tube because of the detrimental distortion of beam trio register. Both vertical and horizontal deflection must be corrected for pincushion distortion.

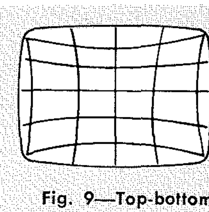
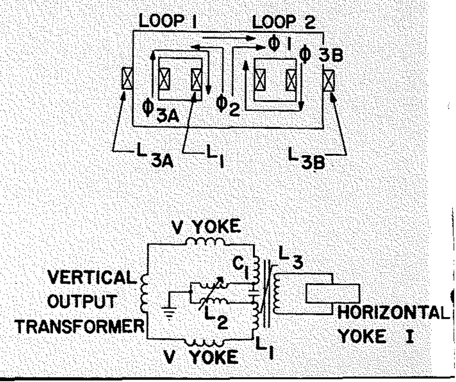


Fig. 8—Straight line distortion due to pincushion errors.

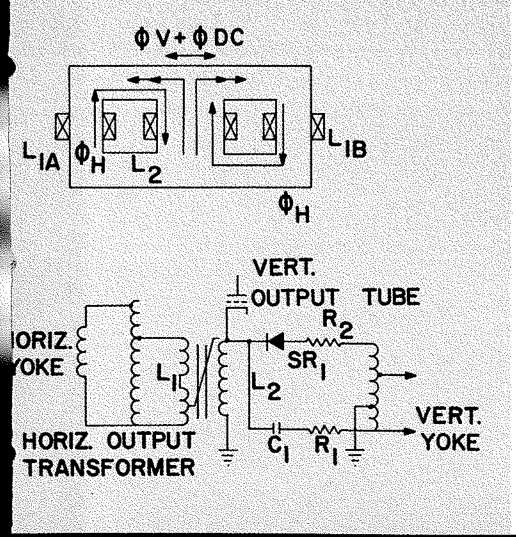
Fig. 9—Top-bottom pincushion-correction circuit.



Vertical-deflection pincushion distortion results in maximum curvature of the horizontal lines at the top and bottom of the raster, but of opposite polarity; the amount of distortion decreases towards the horizontal center line of the raster (Fig. 8). Analysis of the above condition indicates that correction information can be introduced into the vertical deflection yoke. Such correction must be maximum at the horizontal center of the raster to correct the distortion of the horizontal lines and must vary at a horizontal rate to provide negative correction at the horizontal extremities of the raster. The transition must be parabolic and continuous and the degree of correction must diminish toward the vertical center of the raster and increase again at the lower part of the raster, but in opposite phase.

This distortion could be corrected by using a relatively conventional push-pull vacuum tube or transistor modulators that properly combine vertical and horizontal waveforms. However, a more sophisticated and economical means was developed using a push-pull saturable reactor; operation can be explained by analysis of flux paths (Fig. 9). Flux path ϕ_1 is due to a fixed-biasing magnet similar in action to the biasing of a tube. Flux path ϕ_2 is proportional to the vertical yoke current and ϕ_3 is proportional to the horizontal yoke current. When the vertical current is zero, as in the case at the center of the raster, the unit is balanced and flux path $\phi_{3A} = \phi_{3B}$. These two fluxes are opposite in phase, thus resulting in a zero induced voltage in vertical coil L_1. The other extreme is when vertical current is at its positive maximum, as is the case at the top of the raster; then in loop 1 saturating $\phi = \phi_1 - \phi_2$ and in loop 2 saturating $\phi = \phi_1 + \phi_2$. Because of the resultant unbalance, different levels of core saturation exist in loops 1 and 2. Flux path ϕ_3 is no longer balanced in the center leg and is equal to $\phi_{3A} - \phi_{3B}$, producing an induced voltage at the horizontal rate in L_1. Conversely, when the vertical current is at its negative maximum the resultant induced voltage in L_1 is proportional to $\phi_{3B} - \phi_{3A}$. At all points in between these extremes the ϕ_{3A} and ϕ_{3B} ratio is directly dependent upon the magnitude of ϕ_2, or the vertical yoke current. From the above it is demonstrated that both a decreasing correction towards the center and a reversal of phase between the vertical halves of the raster have been achieved.

Fig. 10—Side pincushion-correction circuit.



To obtain proper phasing, L_1 and L_2 are tuned with C_1. This results in a sinusoidal horizontal rate correction output as compared to the ideal parabolic output introduced in the vertical deflection yoke. However, because of the long retrace time of the horizontal scan, the utilized portion of the sine wave is an acceptable approximation.

Normal production tolerance in L_1 and C_1 require external compensation to obtain proper L/C relationship for extreme probability limits. Because it would be undesirable and impractical to tune the modulator, an additional coil (L_2) was added. Since the coil is external to the modulator field, the effective coil inductance can be varied over considerable range with minute effects on the generator output (L_1).

Horizontal-deflection pincushion distortion can be treated simply as undesirable variations in scan width (Fig. 8). A dynamic width control was required which would parabolically and at a field rate, reduce the width at top and bottom with little effect in the center. This was achieved by a balanced variable reactor (Fig. 10) with L_2 suitably biased from the cathode of the vertical output tube. L_{1A} *and* L_{1B} represent a shunting inductance across a portion of the horizontal yoke current and remain in balance with respect to L_2. To properly shape the vertical drive current, C_1, SR_1, R_1, and R_2 are used so that when integrated by L_2, a parabolic vertical waveform results.

When this current is added to the bias current in L_1, the resultant flux variation controls the amount of saturation in the core; thus, the resultant inductance change of L_1 varies the amount of shunted yoke current. Since this variation is accomplished in a parabolic vertical form, proper dynamic-width control is achieved to correct for horizontal or side pincushion-raster deflection. By proper control of the device's variables, no further adjustments or compensations are required in the receiver.

The basic difference between the two described modulators is that the top and bottom (or vertical) device is basically a controlled generator, whereas the side (or horizontal) device is a variable reactor. Reactors of the form described are quite suitable for this application. The small size, stability, low impedance, and virtual indestructability from overloading and arcing make possible the addition of fairly sophisticated circuits without detriment to overall receiver reliability.

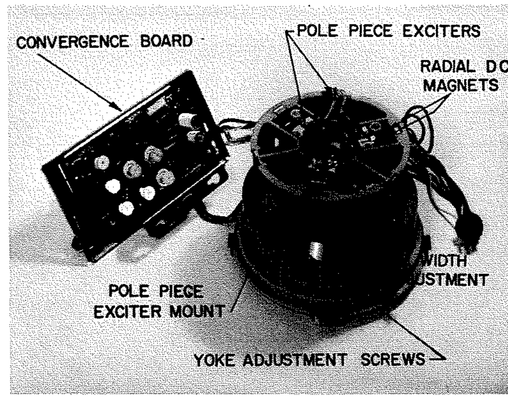


Fig. 11—Yoke mount and accompanying devices.

YOKE MOUNT

The design of the yoke mount involved numerous points that were unique to the 90° rectangular tube. The new design did not permit the yoke to align itself properly on the neck of the picture tube, as was possible in the past. In addition to holding the yoke in position on the neck of the tube, a provision had to be made to permit axial movement of the yoke to provide for purity setup. In conjunction with the yoke mount, the convergence pole-piece exciters were mounted on a device that would snap onto the rear end of the yoke mount and thus guarantee the positioning of the pole-piece exciters. A feature of the pole-piece device permits minor corrections of the blue raster by a small rotation of the pole-piece assembly around the neck of the tube. A simple screw adjustment in the yoke mount raises or lowers the yoke with respect to the tube neck. This adjustment controls minor variations of blue raster width caused by yoke and picture-tube manufacturing variations (Fig. 11).

CONCLUSION

The 25 control and functional areas of redesign and development originally listed as design goals for the 25-inch color TV receiver were achieved with the help and cooperation of design engineers in both the Electronic Components and Devices and RCA Victor Home Instruments Divisions. The performance of the 25-inch color instrument, in addition to meeting the design goals peculiar to the CTC-17, meets or surpasses the requirements of previous color instrument designs.

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FEATURES OF THE RCA RECTANGULAR COLOR PICTURE TUBE FAMILY

Rectangular color picture tubes were introduced by RCA to the TV industry in 1964. This paper discusses various features of the rectangular tubes, including physical, electrical, and performance characteristics, and component and circuit considerations. New features, such as Einzel-lens guns and blue-gun-down operation are described.

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SINCE the introduction by RCA of the first rectangular color picture tube in 1964, the family of rectangular color picture tubes has grown to three basic sizes (tube diagonal): 25 inches, 19 inches, and 15 inches. The 25AP22 and 25BP22 tubes (now superseded by 25AP22A and 25BP22A, respectively) were introduced commercially in the fall of 1964; the 19EYP22 and 19EXP22 tubes were introduced commercially in the spring of 1965. The most recent additions to this rectangular picture tube family are the 15LP22 and 15KP22 tubes, to be introduced this year.

These tubes make up the family of RCA *Hi-Lite* color picture tubes, which contain the RCA rare-earth phosphor screen and the RCA-developed, thermally compensated, shadow-mask *Perma-Chrome* assembly. This phosphor screen uses a red-emitting rare-earth phosphor and improved blue-emitting and green-emitting sulfide phosphors. The new group of phosphors is more efficient than the all-sulfide group previously used and is capable of producing significantly brighter pictures in both color and black and white. The thermally compensated construction of the new Perma-Chrome shadow-mask assembly eliminates not only the need for special preheating during assembly line setup, but also a cold-to-hot compromise purity and white-uniformity adjustment of the picture tube and its associated components. These two major design improvements facilitate the design and production of TV receivers having superior brightness and stabilized high-quality purity and white uniformity.

The 15LP22 and 15KP22 are the first color picture tubes to use the *Einzel* or single-voltage electrostatic-focus-lens type of electron gun. This type of gun maintains sharp focus even

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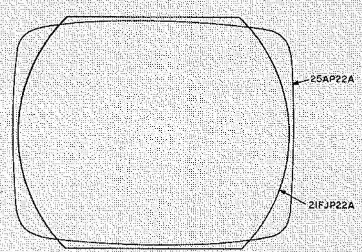


Fig. 1—Comparison of relative screen areas of 21FJP22A and 25AP22A.

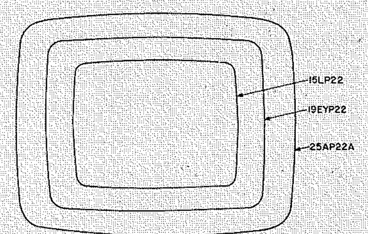


Fig. 2—Relative screen sizes of 15LP22, 19EYP22, and 25AP22A.

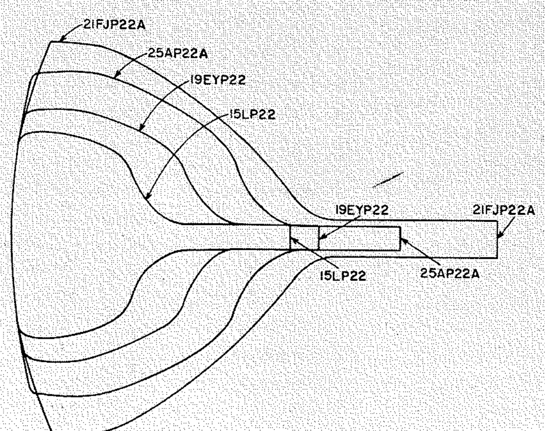


Fig. 3—Profile views of the rectangular 90° color picture tubes in comparison to the 70° round color picture tube.

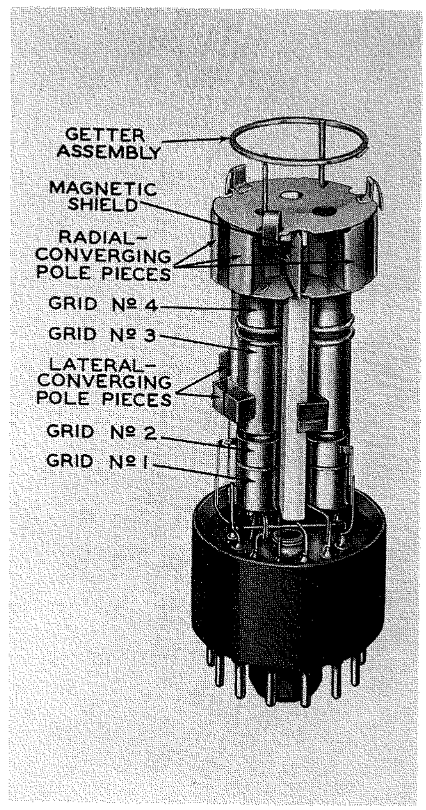


Fig. 4—Electron gun assembly of 21-inch color picture tube, showing internal pole pieces.

TABLE I—Minimum Screen Dimensions of the 21FJP22A and 25AP22A

	21FJP22A	25AP22A
Width (in.)	19.25	19.875
Height (in.)	16	15.575
Area (sq. in.)	267	295

TABLE II—Physical Characteristics of 25AP22A, 19EYP22, and 15LP22 Tubes

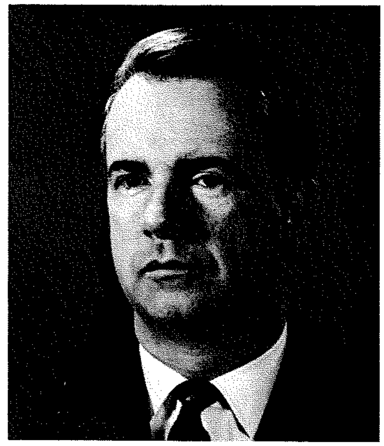
	25AP22A	19EYP22	15LP22
Min. Screen Diag. (in.)	22.995	18.075	13.557
Min. Screen Width (in.)	19.875	15.585	11.689
Min. Screen Height (in.)	15.575	12.185	9.139
Max. Overall Length (in.)	20.924	18.048	15.191
Min. Screen Area (sq. in.)	295	180	102
Weight (lb.)	42	24	12.5

with variations in the high voltage. Receiver costs are reduced because the circuitry is simplified by: 1) the elimination of a high-voltage focus supply, and 2) the lower grid-No. 2 voltage required for the desired beam current.

PHYSICAL CHARACTERISTICS

The shorter tube length, made possible by the wider deflection angle, and the rectangular-shaped screen permit more pleasing cabinet styling and less bulky cabinets. The 25-inch size provides an image size on the TV receiver that is at least equivalent to that obtainable with the mature 21-inch 21FJP22A and 21FBP22A color picture tubes. These tube types are compared in Table I and Fig. 1. The 295-square-inch area of the 25AP22A screen is more than 10% larger than that of the 21FJP22A and has a configuration that corresponds much more closely to the transmitted picture.

The 19- and 15-inch tube sizes provide a uniform sequence of picture sizes that meet the range of consumer demand. These picture sizes are shown in relative scale in Fig. 2. A comparison of the physical characteristics of the three rectangular tubes is given in Table II. The shorter length of these tubes is compared with the 70° round tube in the profile views of Fig. 3. The 25-inch tube is 4.295 inches shorter than the 21-inch tube. The 19-inch tube is 2.876 inches shorter than the



ROBERT W. HAGMANN received his BEE from University of Minnesota in 1948, and has done graduate work at Franklin and Marshall College. He started with RCA in 1948 as a specialized trainee and joined the Electron Tube Div. in 1949 as an engineer in the application areas for color picture tubes. In 1959 he became Manager, Application Engineering Laboratory, Color Picture Tube Engineering. He holds two patents in the area of convergence circuitry and is a senior member of the IEEE. Presently he is Manager, Application and Reliability Laboratory, Color Picture Tube Engineering.

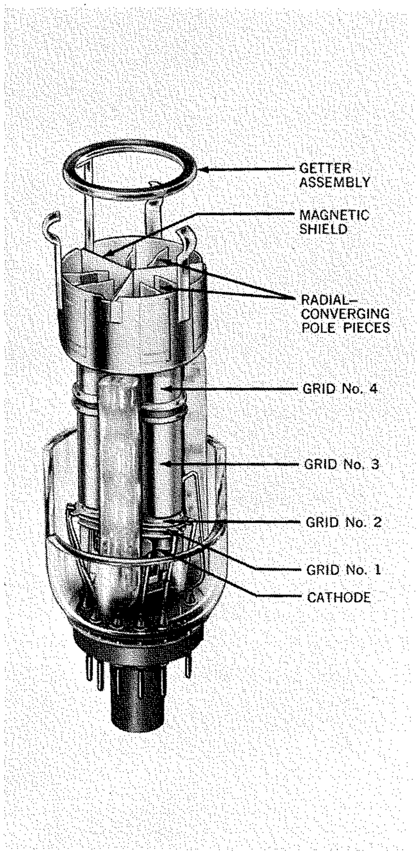


Fig. 5—Electron gun assembly of 19-inch and 25-inch color picture tubes.

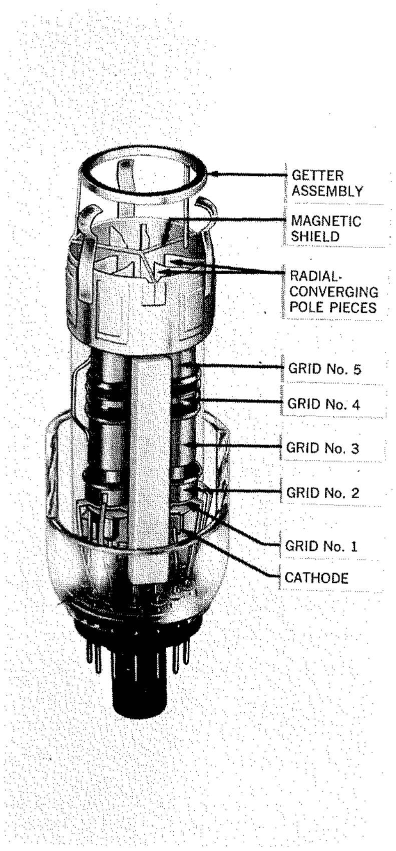


Fig. 6—Electron gun assembly of the 15-inch color picture tube.

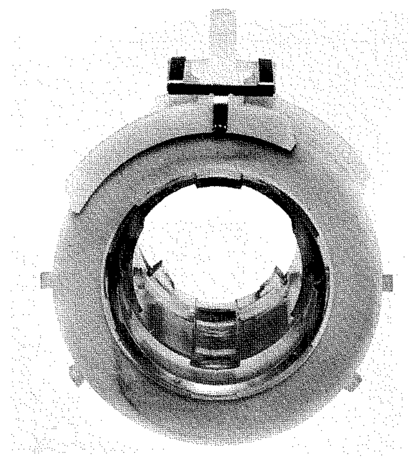
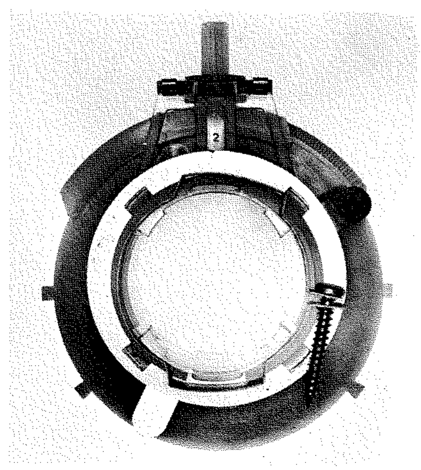


Fig. 7—Purity-lateral converging assembly (front and back views) designed and manufactured by Fastex Div., Illinois Tool Works, Inc.



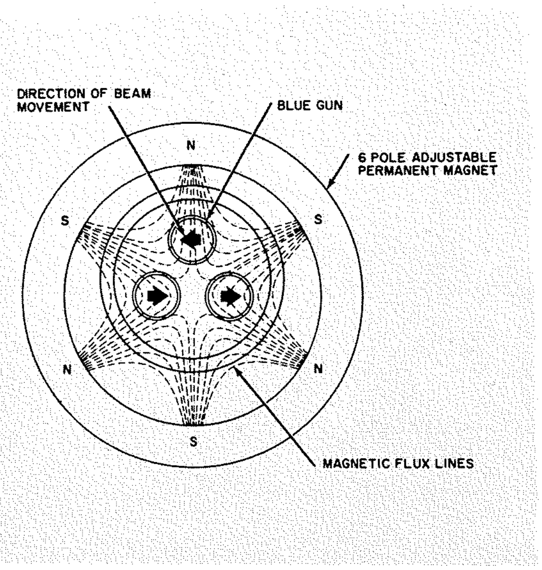


Fig. 8—Flux pattern of 6-pole lateral-converging assembly.

25-inch type, and the 15-inch tube is 2.857 inches shorter than the 19-inch tube.

ELECTRICAL CHARACTERISTICS

Electrical characteristics for the three rectangular color picture tubes are compared in Table III. The 19EYP22 has the same design-maximum anode voltage rating as the 25AP22A; however, it carries a lower rating for the design-maximum total-anode long-term-average current. This lower rating results from the reduced area of the screen. The 15LP22 also carries the reduced anode current rating; however, it also has a lower design-maximum anode-voltage rating. This lower anode-voltage rating, desirable for the Einzel-lens type of gun, maintains satisfactory resolution as a result of the shorter throw distance.

PERFORMANCE CHARACTERISTICS

All tubes in the rectangular color picture tube family are of the Hi-Lite design, which features a phosphor-dot screen using a rare-earth red-emitting phosphor and improved blue and green sulphide phosphors. This new group of color phosphors is more efficient than the all-sulphide group and is capable of producing significantly brighter color and black-and-white pictures. Table IV gives the relative light output characteristics of these tube types. These percentages represent the relative light output when the tubes are operated at their maximum ratings. As the tube size is reduced, the light-output efficiency increases because of the reduced area and the slightly higher transmission of the faceplate which results from the use of thinner glass. The 25AP22A and 19EYP22 tubes both have a faceplate transmission of approximately 41%, whereas the 15LP22 has a transmission of about 44%.

TABLE III—Electrical Characteristics of 25AP22A, 19EYP22, and 15LP22 Tubes

	25AP22A	19EYP22	15LP22
Design-Maximum Anode Voltage (kV)	27.5	27.5	22.5
Design-Maximum Total Anode Current, Long-Term Average (μA)	1000	750	750

TABLE IV—Relative Light Output of Rectangular Color Picture Tubes

Tube Type	Relative Light Output
25AP22A	100%
19EYP22	135%
15LP22	168%

The laminated versions of the 90° color picture tubes all have an integral filter-glass protective window which is sealed to the faceplate of the tube with clear resin. This construction eliminates the need for a separate safety glass window and its companion dust seal on the receiver. Internal reflections are therefore reduced with subsequent improvement in picture contrast and color saturation. The surface of the protective window is etched to minimize specular reflection.

COMPONENT CONSIDERATIONS

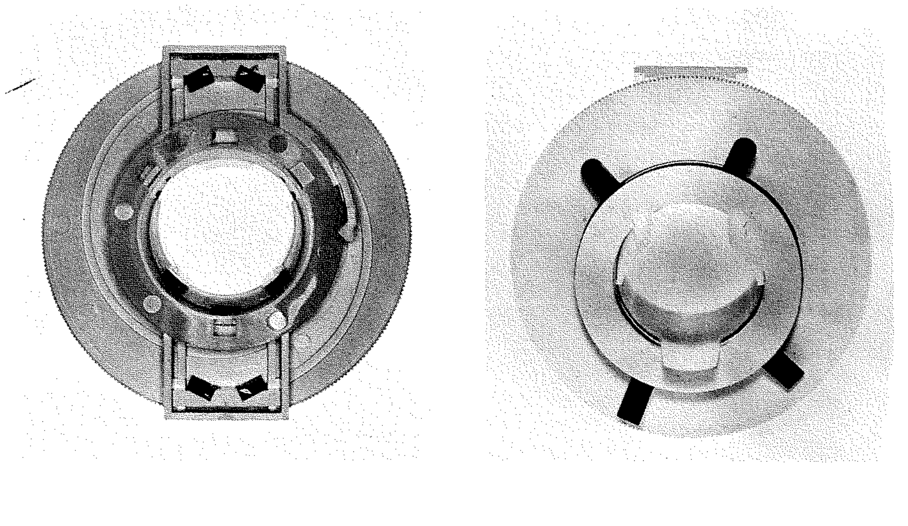
The smaller neck (1$\frac{7}{16}$ inches) of the 90° color picture tubes, as compared to the 2-inch neck of the 70° tubes, permits the use of wider deflection angle tubes with essentially the same deflection power as that required for the older 70° types.

Recent studies indicate that pincushion raster distortion, as observed from a typical viewing position, is reduced when the color picture tube is operated with the blue gun down, as opposed to the previous standard technique of blue-gun-up operation. Blue-gun-down operation is one of the improvements in the 15-inch rectangular color picture tubes.

The differences in the center-to-edge ratio of electron-beam throw distance among the various tube sizes require that a different deflecting yoke be used for each type to obtain optimum performance. The design of wide-angle shadow-mask color picture tubes involves a marriage with the design of the deflecting yoke. The register of the electron beam on its associated phosphor dot is influenced by the deflecting-yoke field. The yoke performance characteristics of raster shape, beam convergence, and register must be balanced to obtain optimum performance and cost. The yoke designs have been tailored to favor the optimizing of beam register and convergence characteristics; raster shape characteristics can be corrected, if needed, by circuit techniques.

The smaller neck diameter necessitated the use of closer spaced electron guns, creating new requirements for the purity and convergence-correction components. The closer spacing of electron beams, in conjunction with the internal magnetic structures of the guns and the external correcting components, can result in considerable interaction between the purity and convergence component adjustments. In the 70° picture tube family, a pair of internal pole pieces assists the coupling of external magnetic fields to the electron beams for the purpose of convergence. These internal lateral-converging pole pieces (Fig. 4) increase the purity-convergence interaction, thus requiring increased setup time on the production line of set manufacturers. Tests indicated that this interaction

Fig. 9—Purity-lateral converging assembly (front and back views) used in RCA receivers.



could be minimized if the pole pieces were not used. Consequently the guns for the new rectangular tubes were designed without internal lateral-converging pole pieces. The electron gun assemblies of the 25- and 19-inch tubes are identical. This electron-gun assembly is shown in Fig. 5. The gun assembly used in the 15-inch color picture tube is shown in Fig. 6, which reveals the differences between the Einzel-lens gun and the bipotential gun structure shown in Fig. 5.

The elimination of the lateral-converging pole pieces required a new design for the lateral-converging component. A component design providing this type of correction is shown in Fig. 7. This component consists of six magnetic poles. A schematic drawing of its flux pattern is shown in Fig. 8. The purity-lateral converging assembly used in RCA receivers is shown in Fig. 9.

To further reduce the purity-convergence interaction, the order of positioning the components on the neck of the tubes was changed. A comparison of component placement on the 70° and 90° tubes is shown in Fig. 10. The relative positions of the purifying magnet and the lateral converging device are interchanged so that the purity device is a maximum distance from the radial-converging magnetic pole pieces. The purifying magnet used for the wide-angle small-neck tubes is basically a scaled-down version of the type used for the 70° large-neck tubes. This device compensates for the effects of uniform extraneous magnetic fields and other factors that could affect register of the electron-beam trio with its respective phosphor-dot trio. This device is placed on the neck of the tube in a plane coinciding with the plane of the three cathodes of the color picture tubes.

CIRCUIT CONSIDERATIONS

The wide deflection angle of the 90° tubes requires that more careful consideration be given to the design of the convergence circuit. In particular, the current waveforms required for correcting the convergence of the three electron beams on the axes of the picture screen require more precise shape and control. The circuits are similar to those used for the 70° tube; however, the horizontal blue dynamic circuit must generate a more precise waveform. Improved current waveforms can be obtained by several techniques, one of which is used by RCA in the CTC-17 receiver design. This system (Fig. 11) provides a waveform that more nearly approximates the shape needed for optimum convergence correction. Another system, which has proven satisfactory for improving the waveshape of the horizontal blue convergence circuit, is shown in Fig. 12. This system provides adequate waveform control and requires less power. To obtain maximum performance from the picture tube, it is desirable to operate it at a high value of cutoff voltage within the picture-tube cutoff rating of 200 volts design maximum. This high cutoff operation provides the optimum in spot size for a given beam current (Fig. 13).

Recommendations for protection of the color picture tube during cascade arcing have been made to the set makers. Tests have indicated that picture tube damage, such as heater burnout, can occur during arc-over between chassis elements that connect sources of high energy by an ionized path to picture-tube elements. Experience has shown that all the B+ and AC sources must be placed at least 1/4 inch away from all connections that are directly connected to picture-tube elements; the minimum DC impedance to any DC source from any of the leads to the picture tube must be at least 1000 ohms.

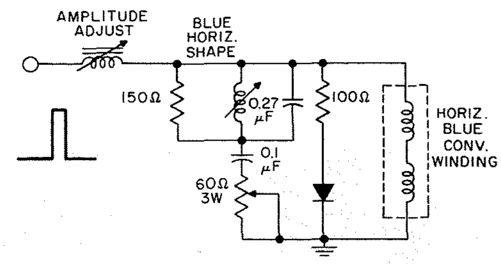


Fig. 11—Basic horizontal blue convergence circuit of RCA CTC-17.

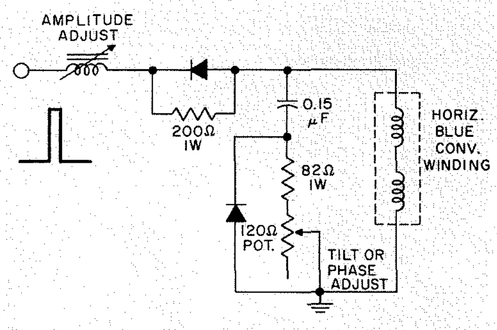


Fig. 12—Horizontal blue convergence circuit.

CONCLUSION

The number of color picture tube types continues to grow as the color TV industry expands. Einzel-lens guns and blue-gun-down operation will probably be introduced into tube sizes other than the 15-inch tube. As the pressures of cost reduction become more intense, new features, such as *bare-face* implosion systems, will undoubtedly be incorporated into the designs of color picture tubes.

Fig. 10—Comparison of component placement.

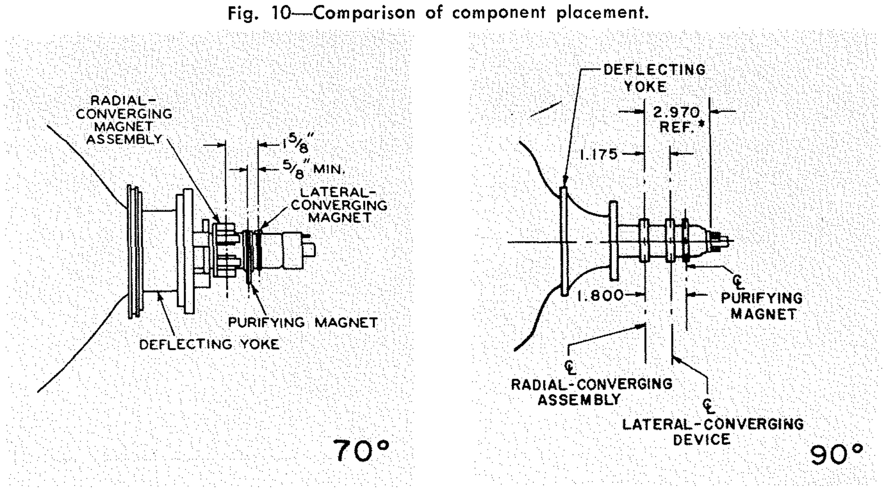
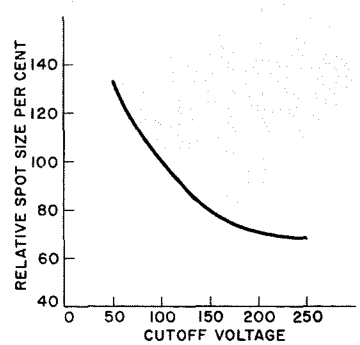


Fig. 13—Relative spot size in percent as a function of cutoff voltage for a beam current of 1000 microamperes.



LUMINESCENCE AND COLOR TV

This paper gives a brief history of luminescence, discusses the physics and chemistry of cathodoluminescent materials, and reviews the development and use of these materials in color television. Editor's Note: Since this paper was written the authors have received the David Sarnoff Outstanding Team Award in Science. (See article this issue.)

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The term *luminescence* was introduced by E. Wiedemann in 1888 to encompass all forms of light radiation, with the exception of those due to heat. A more modern definition of luminescence would be "a process whereby matter generates nonthermal radiation which is characteristic of the particular luminescent material."[1]

Historically, luminescence is one of the oldest solid-state phenomena, the first known synthesis of a luminescent material having been carried out in 1603.[2] (A luminescent material is often referred to as a *phosphor*, from the Greek, "bearer of light".) It was only in the latter half of the 19th century and in the early 20th century that more modern scientific techniques were applied to the synthesis and study of luminescence. Research in luminescence at RCA began in the early 1930's.

By 1938, emission spectra were being studied at low temperature, and the quantum theory of solids was being applied by Mott and Gurney and by Seitz. At this time, Seitz[3] raised four fundamental questions which demonstrate some of the major problems in luminescence, even as they exist today: 1) what is the atomic nature of the luminescence center? 2) what electronic transitions take place during absorption? 3) what transitions give rise to fluorescence? and 4) what transitions give rise to phosphorescence?

We see then that the rate of progress of understanding luminescence as one of the oldest solid-state phenomena has been very slow, due in major part to complexities that are discussed in later sections of this paper.

INCANDESCENCE VS LUMINESCENCE

Incandescence may be defined as the emission of light due to the temperature of a source. The radiation from most incandescent solids approximates that of a black body (Fig. 1) for which there will be more total-flux than is obtainable from any other source operating at the same temperature. In addition, for black-body radiation, Planck's Law applies:

$$W_J d\lambda = C_1 \frac{\lambda^{-5}}{\exp\frac{C_2}{\lambda T} - 1} d\lambda \qquad (1)$$

where W_J = radiated watts/cm^2 of surface/micron, at wavelength λ; λ = wavelength in microns; T = temperature of the black body in °K; $C_1 = 3.738 \times 10^{-5}$ erg cm^2 sec^{-1}; and $C_2 = 1.438$ cm degree.

With increasing temperature, the *peak* wavelength of emission is shifted to shorter wavelength. The value of the peak wavelength, in microns, can be obtained from Wien's Displacement Law:

$$\lambda_{PK} = 2.9 \times 10^3\, T^{-1} \qquad (2)$$

By integrating W_J of equation (1) for values of λ from zero to infinity, we obtain the Stefan-Boltzman Law, that the total radiant power per unit area of a black body varies as the fourth power of the temperature:

$$W = cT^4 \text{ (watts cm}^{-2}\text{)} \qquad (3)$$

where

$$c = 5.679 \times 10^{-5} \text{ erg cm}^{-2} \text{ deg}^{-4} \text{ sec}^{-1}$$

Although the emission from an incandescent solid is the result of a statistically averaged effect over all atoms of the source, the nature of *luminescence* emission is determined by the physical-chemical character of discrete volumes or elements. For luminescent solids these can be far less than 1 percent of the total volume. Thus, in luminescence the basic time constant involved in transient responses is determined primarily by the luminescence center, rather than by the bulk of the containing material.

If, in luminescence emission, P_R is the probability of radiative transitions, P_N is the probability of nonradiative transitions, and η is the luminescence-emission efficiency, then:

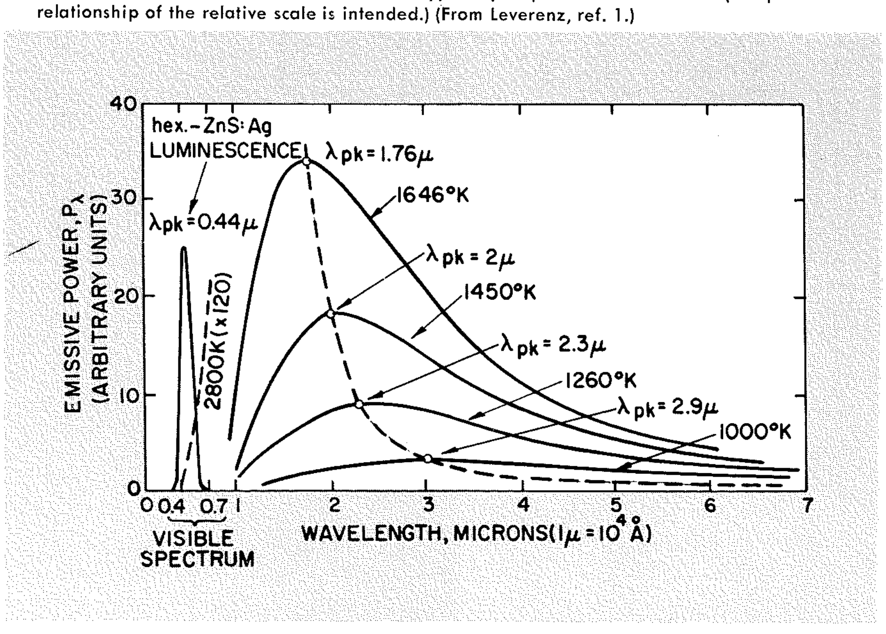
$$\eta = \frac{P_R}{P_R + P_N} \qquad (4)$$

If we assume that P_R is essentially independent of temperature and that $P_N = K \exp(-E/kT)$ then:

$$\eta = \frac{1}{1 + a \exp(-E/kT)} \qquad (5)$$

an expression which fits remarkably well the efficiency vs temperature curves of many luminescent materials.[4]

Fig. 1—Spectral distribution of thermal radiation from a black body at various temperatures. For comparison, the emission of a typical phosphor is also shown. (No particular relationship of the relative scale is intended.) (From Leverenz, ref. 1.)



Final manuscript received January 20, 1966.

PHYSICS OF SOLID CRYSTALLINE PHOSPHORS

Excitation

The first requirement of a phosphor is that it absorb the excitant and that the energy of the excitant is adequate. Fortunately for TV, this is true for cathode-rays if the screen is made with enough phosphor to cover completely the tube face. Radiation (e.g. ultraviolet light) can also be used as an excitant, but the differences are beyond the scope of this article and will not be dwelt on at length.

Energy Transfer

The second requirement is that there be an efficient mechanism for converting the *absorbed* energy from its original form to another that the bulk material can transfer to sites especially adapted to emit photons. (The nature of these sites is discussed briefly in a later section.) Occasionally the site itself has sufficient absorptivity to intercept the energy and emit a photon without direct aid from the bulk which surrounds the special site.

In phosphors of the zinc-sulfide type, the passage of a high-velocity electron through the host lattice creates free electrons and free holes; the holes remain free only for very short times before being captured (trapped) at localized sites. These then become traps for the free electrons. When the electrons are captured (an electronic process), a photon is frequently emitted and the cycle is complete, since there are now no free or trapped holes or electrons.

Band Gap Models

A commonly used pictorial representation (model) of some of the energy changes that have been discussed is shown in Fig. 2. For a crystalline solid, E_{BG} gives the effective upper limit of photon-energy easily transmitted by a perfect lattice. Transition 1 (Part I) represents the absorption of a photon with energy greater than E_{BG}, creating a hole-electron pair. Excess energy (over E_{BG}) possessed by the carriers is quickly lost to the lattice as heat, so that the effective energy approximates E_{BG}. Transitions 2 and 3 are absorptions by lattice defects or impurities with energies less than E_{BG}. The dotted-line *cage* around transition 3 is intended to show that for some impurities, e.g. rare earths, there are transitions which do not directly involve the valence or conduction bands of the host lattices. The exact location, on the energy scale, of the unexcited ion (ground state) is usually unknown and is placed here primarily to show that the exciting photon's energy, $h\nu_3 < E_{BG}$.

Part II shows events that occur subsequent to the absorption process. Transitions 4 point out that, due to lattice readjustments, the energy available for photon emission at sites A or B will be less than that necessary for excitation. Other defects (E) may be able to capture (transition 5) momentarily the electron freed by transitions 1 or 2. Other defects (D or unexcited A or B sites) may capture holes (transitions 6 and 7). This process is usually considered radiationless because of the difficulty of finding radiation arising from such a transition.

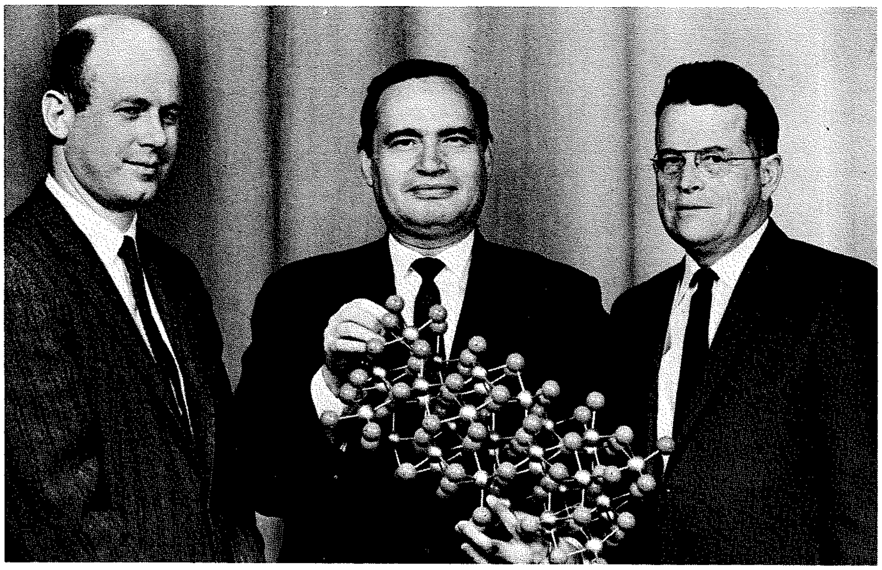
Part III shows the transitions (8, 8′) which give rise to the luminescence for which defects A and B are considered responsible. Transitions 9 and 9′ are the thermal-quenching transitions which reduce luminescence efficiency when the temperature is raised sufficiently high.

It follows that other methods of creating free holes and electrons should give rise to luminescence and, indeed, gamma rays, X-rays, electron beams, ion beams, and even direct-charge injection all lead to photon emission with varying similarities and differences. The chief difference is the efficiency of excitation; the chief similarity is the spectral distribution.

Efficiency

The greatest efficiency in watts absorbed vs. watts generated is achieved usually with direct photon-excitation of the emitting center, with values upward of 70 to 80%. The poorest efficiency is usually with ion beams where the value lies below 0.1%. Electron-beam-excited luminescence efficiency covers a very wide range, being about 20 to 25% at the top.

Since the primary interest in phosphors for color TV is in their characteristics under electron-beam (cathode-ray) excitation, the remainder of this article is confined to this area. There are many specific characteristics, and a few will be discussed to give a working concept



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of the considerations involved in phosphors for cathode-ray TV tubes. Very little is known as to how the energy of a single high-velocity electron is converted to a large number of photons. The broad concepts that must be involved can be stated, but experimental confirmation in any quantitative manner is lacking.

Electron Range

A kindred effect, but one which permits an empirically obtained solution, is the determination of the effective range of an electron in a solid. This range can be measured directly using evaporated films; but for efficient phosphors that are useful only in powder form, the problem is solved in terms of the layer thickness which optimizes the brightness of the layer for the electron velocity under consideration. In TV applications the velocity (in eV) is below 25 keV, dictated by factors not directly concerned with the phosphor. In any case, it is usually found that in the lower range of voltage (below 5 keV), the optimized efficiency drops with decreasing voltage. The effect is very much as if the outer shell of the phosphor particles were inert, or, at least of lower efficiency than the inner portion of the particles, as shown in Fig. 3. The truth of this explanation has yet to be established.

Secondary Emission

The shallow penetration of electron beams in solids leads to yet another characteristic which can be troublesome —secondary emission. As mentioned previously, high-speed electrons generate free electrons and holes. If even one of the slower electrons can reach the surface with a velocity high enough to escape the drag of the bulk ions and electrons, the charge of the particle has been returned to the state, presumably neutral, before the impact of the original primary. However, if more than one electron escapes for each impinging primary, the particle now begins to charge positively, exerting still more drag on further attempts to escape. The process levels off when only one electron escapes for each electron incident. If no electrons escape, the charge builds negatively, repelling, or at least slowing, further primaries. This process, too, levels off when leakage by any means equals the effective influx of the exciting beam. This latter behavior is the more harmful type, and to guard against it, phosphor layers are covered with a very thin evaporated aluminum layer which provides an easy leakage path if needed. Any loss of electron energy in penetrating the aluminum is more than offset, at TV voltages, by the reflective effect which returns toward the viewer much of the light that otherwise would have proceeded back toward the source of the electron beam.

Saturation

Another troublesome characteristic of phosphors is that neither the spectrum nor overall efficiency is completely independent of the current level of the electron beam. This effect, together with the temperature change caused by high currents, requires that phosphors be chosen carefully for each application. Screens used for projection TV require a different type than those used for direct viewing. The requirements of phosphors used for black-and-white (B/W) TV are less strict than are those used for color TV. Generally, all phosphors have an upper limit of current loading per unit area beyond which the spectrum will change and the efficiency will drop. The value of this upper limit and the degree of effect with increasing current differs widely from phosphor to phosphor. In color TV, either change of color or efficiency can lead to significant departures from the desired hues in high-brightness regions of the scene being presented. The red component of color TV screens is the component required to show the least color change, since the usual effect on color is for the hue to move to shorter wave lengths at high excitation levels. This criticality arises from the nature of the human eye, which is especially sensitive to hue and brightness change in the yellow and orange-red portions of the spectrum.

CHEMICAL ASPECTS OF PHOSPHORS

Before going into the specifics of phosphor chemistry as pertaining to TV, certain broad requirements of stability must be recognized. For a material to be used as a phosphor in a TV tube as now manufactured, it must be able to withstand bombardment by high-energy electrons and exposure to a moist, oxygen-containing atmosphere at 400°C for about 1 hour. Of the materials capable of excitation by cathode-rays, many fail under the required tube processing. More inorganic phosphors resist this treatment than do organic phosphors. In addition, the continued bombardment by high-velocity electrons produces a permanent loss of efficiency more quickly with organics than with inorganics. At the present time, there are no known organic phosphors that can compete for application in TV screens.

Classification of Phosphors

The modern classification of cathodoluminescent materials includes defect-semiconductor materials, i.e., the transitions are between defect-induced states of a crystal lattice, or materials containing ions which allow luminescent transitions among their bound electrons. These luminescing ions are transition metal or intertransition metal ions.

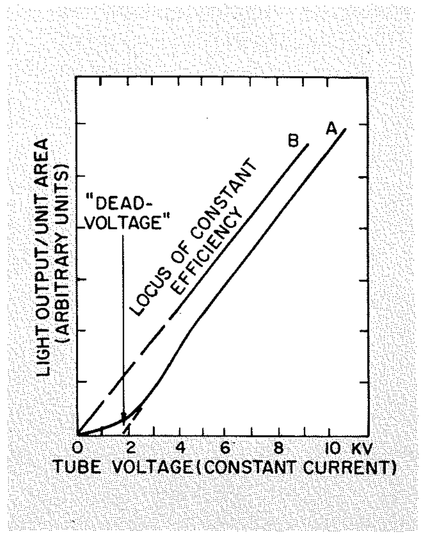
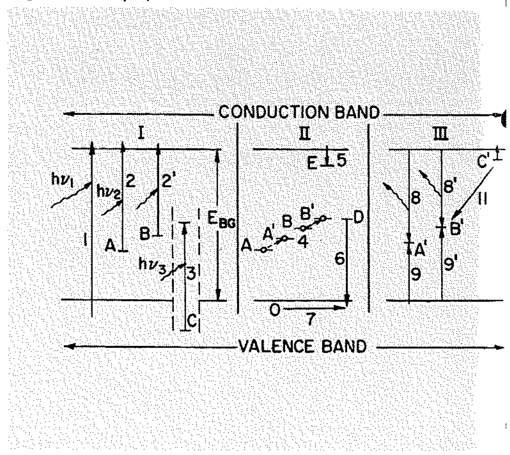


Fig. 3—The effect of voltage on efficiency which leads to the concept of "dead voltage."

The defect-semiconductor type of phosphor is best illustrated by the zinc-cadmium sulfide phosphors in which two types of defects, called donors and acceptors, are induced. These donors and acceptors can arise from deviations in stoichiometry or added impurities; in any event, the electronic distribution of the crystal is altered, with those regions that tend to hold onto electrons being the donors and those that are electron deficient being the acceptors. The manipulation of these defects is to a large extent the solid-state chemistry of this type of phosphor. Since these defects are oppositely charged, they compensate each other and will associate, as is noted later in connection with the associated lumi-

Fig. 2—Band-model representation of electronic energy changes during excitation (I), very shortly after excitation (II), and later during emission (III).



nescent center. The behavior of these defects can be treated by the *law of mass action* in a manner analogous to equilibrium constants or solubility products in aqueous chemistry:

$$\text{donor}^- + \text{acceptor}^+ = \text{donor-acceptor}^\circ \tag{6}$$

$$\frac{[\text{donor}^-]\,[\text{acceptor}^+]}{[\text{donor-acceptor}^\circ]} = K_{constant.} \tag{7}$$

Equation (6) represents an actual displacement in the lattice. In equation (7), the brackets denote (approximately) concentrations. The constant K is for a given temperature and pressure and will change for any change in these quantities. What these equations say is that a piece of matter maintains electrical neutrality so that all positive species equal all negative species. In the simplest case, where only two oppositely charged species are involved, the concentrations are equal. However, it is possible to add intentionally an excess of one species over the other. When this happens, a third species (such as vacancies) appears to compensate the excess. All of these defect species can be related to each other by a series of equilibria expressions.[5]

Charge Compensation

For commercial phosphors, the added impurity types are the most important, and a few examples may be useful. If a perfect zinc sulfide lattice is pictured from which a zinc atom is removed, the region around the missing zinc will be deficient of the two electrons needed to change the divalent zinc ion to the zinc atom removed, as shown in Fig. 4a. If now a silver atom enters the vacant site and is allowed to ionize to its normal monovalent state, it can contribute only one electron. This region of the silver ion is still electron deficient (Fig. 4b) and could accept another electron (therefore being known as an *acceptor*). On the other hand, if a divalent sulfide ion is converted to a sulfur atom, two electrons are donated to the lattice in the process. Now, if a chlorine atom is placed in the sulfur vacancy and allowed to become a monovalent ion by accepting one electron, there is still an electron-rich region in the lattice (Fig. 4c) which is known as a *donor*. The zinc sulfide, silver chloride (ZnS:Ag:Cl) just described is the blue-emitting phosphor used in TV. Referring to Fig. 2, the silver region introduces level A, and the chloride introduces level B.

By following a line of argument similar to the preceding, it can be shown that a trivalent cation substituted on a divalent cation site is a donor (Fig. 4d), and a trivalent anion substituted on a divalent anion site is an acceptor. A glance at a periodic table to ascertain the number of ions which can form donors or acceptors in II-VI compounds shows that a large number of combinations are possible.

Associated Luminescence Centers

The usually accepted picture of the electronic transition responsible for the luminescence of these materials is one in which an electron in the conduction band, or very close to it, combines with a hole trapped in the acceptor center as shown in transition 8 of Fig. 2. This condition holds for donor and acceptor concentrations of the order of 10^{-4} atoms of dopant per gram atom of host. However, if the concentrations of donors and acceptors are both increased to the 10^{-3} level, longer wavelength emissions are observed. These emissions are usually explained as occurring directly between the donor and acceptor levels as shown in transition 11 of Fig. 2. This is the so-called associated center.

Solid-Solution Phosphors

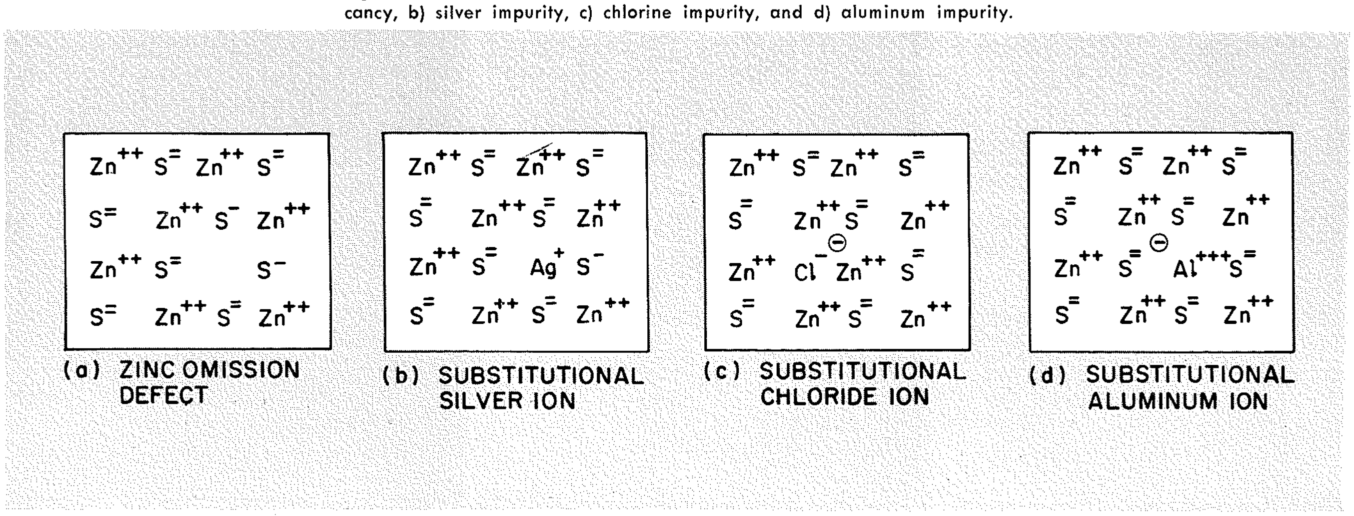
The defect-semiconductor type of phosphor possesses in greater degree the property of allowing emission color to be changed in a continuous manner, in contrast to transition-metal types of phosphors. Since the luminescent transition is across a considerable portion of the band gap, a change in the band gap will change the energy of a transition and thereby its color. If cadmium sulfide is alloyed with zinc sulfide, the emission from silver and cholride activation will progressively move from the blue through the green, yellow, and red, and finally appear in the infrared in pure cadmium sulfide. Fig. 5 shows how the emission shifts with cadmium content. In the all-sulfide color tube, this mixed crystal or alloy system was the basis for all the color phosphors.

Transition-Metal Activators

The other main group of phosphors used in TV is that in which the luminescent transition is confined to the electrons of a specific ion, as in transition 3 of Fig. 2. One of the earliest phosphors used in cathode-ray tubes (CRT) was *willemite*, a zinc orthosilicate activated with divalent manganese; a transition within the manganese is responsible for the light emission. This type of phosphor has two main categories: those containing transition-metal ions, and those containing rare-earth ions.

Of the transition-metal ions, divalent manganese is by far the most important. In these ions the luminescent transition is between states of the outer electrons which are involved in compound formation. For this reason the energy separa-

Fig. 4—Two-dimensional representation of defects in ZnS: a) lattice vacancy, b) silver impurity, c) chlorine impurity, and d) aluminum impurity.



tion of these states is dependent upon the particular state of combination. In more precise terms the energy levels are highly dependent upon the crystal field. As a result of this circumstance, the emission color can be changed, within limits, by changing the crystal field. The tetragonal field of the strength of zinc orthosilicate produces the green willemite emission from divalent manganese, but the octahedral field of zinc orthophosphate produces a red emission from the same ion. Since alloying will also change the crystal field, emission color changes can be brought about by this means, but these changes are less drastic than those in the semiconductor-defect class.

Rare Earths

Rare-earth ions are of major interest in the field of TV phosphors today because they have become available in sufficient purity and quantity at a reasonable price in the last few years. It has been known for many years that they possessed luminescent properties, but their availability at earlier times was limited to gram quantities at enormous cost. The first of these ions to be used in TV is trivalent europium, but it probably will not be the only one.

In rare-earth ions the electrons responsible for light emission are shielded from the bonding electrons; therefore, the transition energy is affected only slightly by the state of combination. Most rare-earth ions have many light-emitting transitions, but most of these emissions are of very limited spectral range, many less than 10 Ångstroms in bandwidth. This is contrasted to previously described phosphors having bandwidths of hundreds of Ångstroms. Although the state of combination does little to change the particular spectral location of a specific transition, it has a

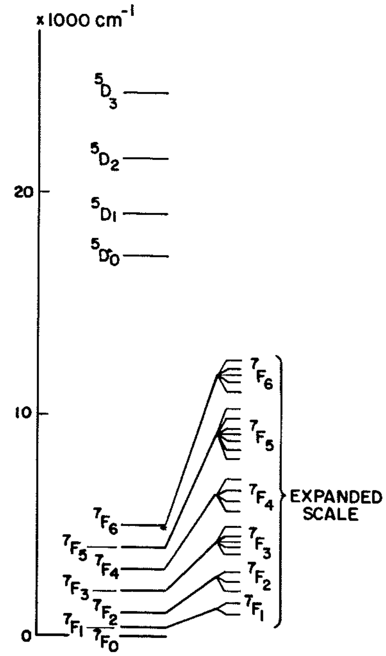


Fig. 6—Generalized energy level diagram for Eu^{+3} ion.

profound effect on which of many transitions is the most intense. For example, Fig. 6 shows a diagrammatic energy level scheme for trivalent europium. In many hosts the transition from $^5D_0$ to $^7F_1$ (5900 Å) is the most intense; but for a good red-emitting TV phosphor, the $^5D_0$ to $^7F_2$ transition (6150 Å) is needed.

PHOSPHORS FOR TELEVISION

Phosphors found their first real uses, other than as a novelty, in the late 1800's, when they were used to detect radiation, such as cathode and X-rays, ultraviolet photons, and gamma-rays. It was not until well into the 1900's that phosphors achieved prominence not only as detectors, but as *converters* of energy, where their emitted light served as a useful source of information. The interesting history of these later developments can be best summarized by examining the development of television.

In the 1930's serious research was begun to improve the cathodoluminescence efficiency of phosphors, in order to make them useful in CRT-types of presentations. New phosphor systems were discovered during this continuing research, many by Leverenz and coworkers at RCA; one of the most efficient of these families was zinc beryllium silicate with manganese activator or, as written in phosphor notation, rbhdl.-$(Zn:Be)_2SiO_4$:Mn. This family yielded phosphors of different emission color, and formed the phosphor basis of one of the earliest B/W TV screens, when mixed with blue-emitting cub.-ZnS:Ag, or rbhdl.-$(Zn:Be)_2SiO_4$:Ti.

In the development of B/W TV in this country, green-emitting rbhdl.-Zn_2SiO_4:Mn and hex.-ZnS:Cu were the first phosphors to be used. In order to produce white, a combination of at least two phosphors, a blue-emitting material and a yellow-emitting material, was utilized. Since colorimetrically a choice of possible complementary pairs existed, the optimum pair to yield the desired white with maximum efficiency had to be evolved. Thus, blue-emitting rbhdl.-Zn_2SiO_4:Ti with the yellow-emitting rbhdl.-$8ZnO{:}BeO{:}5SiO_2$:Mn could be used. Or, as a three-component screen, the rbhdl.-Zn_2SiO_4:Ti with green-emitting rbhdl.-Zn_2SiO_4:Mn and with red-orange-emitting rbhdl.-$8ZnO{:}BeO{:}5SiO_2$:Mn. However, the phosphors that finally won acceptance for B/W TV are derived from a sulfur-dominated family instead of the aforementioned oxygen-dominated materials. Thus, most kinescopes utilized for B/W TV now comprise a mixed screen consisting of blue-emitting cub.-ZnS:Ag and yellow-emitting hex.-(Zn:Cd)S:Ag, with the screen weights of phosphor optimized for maximum efficiency and color under the drive conditions used.

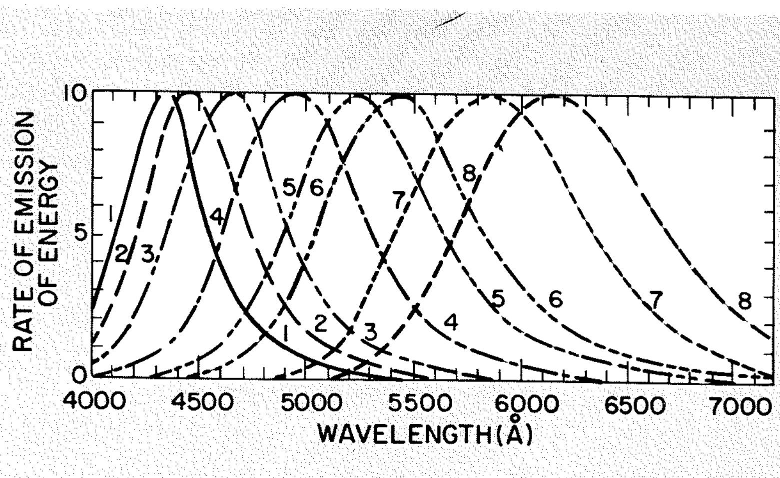
In the development of kinescopes for color TV, at least a dozen phosphors have been investigated to some degree in tubes. Table I, after Hardy,[6] summarizes these materials chronologically and colorimetrically. Since the color kinescope contains separate blue-, green-, and red-emitting phosphors, various combinations of the phosphors listed have been used commercially.

The problems of B/W TV screens, touched on previously, are compounded by the requirements of color TV.

In black-and-white TV, where only a single gun is used, the only requirement is that the brightest *proper* white be obtained for any given current. The mixing of the two colors is accomplished by

Fig. 5—Spectral-distribution curves of the emissions of selected members of the phosphor family (Zn:Cd)S:Ag(0.01), [NaCl(2)], 1200°C (from Leverenz, ref. 1).

Curve		1	2	3	4	5	6	7	8
Mole Proportions	ZnS	10	9.5	9	8	7	6	5	4
	CdS	..	0.5	1	2	3	4	5	6



the screen preparation, and once adjusted to produce *white*, no further changes are possible or needed. In color tubes the color achieved in any particular area of the tube is subject to a variety of factors, many of which involve the signal quality and the set adjustments, as well as the choice of phosphors and the skill of screen preparation. In choosing the phosphor, one aims at a large color gamut and reasonably close values for gun currents for white. If a practical ceiling efficiency (watts out/watts in) is placed on the phosphors, regardless of spectral distribution, then of all possible gamuts, the white brightness (achievable with a given total wattage input) increases as the gamut is reduced. Conversely, striving for the largest possible theoretical gamut reduces the achievable white brightness. In addition, the actual gamut in a TV set, as pointed out by Hardy,[7] is always less than that of which the phosphors themselves are capable. The phosphors chosen at any stage of color TV development have been the result of compromises of many factors. One factor is that most colors of original scenes lie within a smaller gamut than could be achieved by perfecting every aspect of the TV system, including choice of phosphor. Another is that not every phosphor can be easily placed in a commercially feasible tube. This type of problem is beyond the scope of this paper and will not be discussed further.

Phosphors Containing Rare Earths

The most recent change made in color TV screens has been the replacement of the *red sulfide* with the *red rare earth.* The chief difference lies in the spectral distribution of the emission. As used for color TV, the yttrium vanadate:europium phosphor is practically monochromatic in its output. All previous phosphors have had broad emission bands, from 400Å width (at half maximum) upwards to 1000 Å or more. Rare-earth-induced emissions in the proper lattices are seen as many narrow lines, about 1-15 Å wide. Under certain circumstances, most of the energy is emitted in a few lines, which determine the color as being of a definite hue. The fewer, or more closely spaced, the lines, the higher the chroma (or approach to a sensation of monochromaticity). The red rare earth currently used in color TV screens has its most energetic emission line at 6190 Å.

Narrow emission lines have a distinct advantage if they are located in the red portions of the visible spectrum, since the narrow-line emitter concentrates its emission at one spectral point, whereas the broad-band emitter wastes much of its emission at other wavelengths where the eye is relatively insensitive.[8] As a result, the narrow-line emitter can show, for the same radiant output energy, a much higher luminosity than a broadband emitter of the same subjective color, although in most cases the broad emitter is much more efficient on an energy basis. This is illustrated in Fig. 7. The subjective response for any color is described by the values of x and y that lie in the enclosed area. The curved outer line is the locus of all monochromatic radiation. The subjective response, i.e., blue, green, etc., is indicated approximately. On the long-wavelength portion of the monochromatic curve, two curves are superimposed; curve A shows how the visual efficiency of radiation falls off with increasing wavelength, and curve B shows the spectral distribution of the red phosphor used in P22 screens. These curves show that a monochromatic red emitter offers the best compromise among luminosity, radiant output, and achievement of a desired color.

However, calculations have shown that a single-narrow-line emitter in the green region of the spectrum would have to be an extremely intense emitter to compete with a broad-band emitter. A multiple-line emitter whose spectrum leads to chromaticity coordinates identical with a broad-band green would need only to be equal in luminous efficiency to be a valid replacement. A characteristic of line emitters which has no direct bearing on their usefulness and application is the difficulty of determining accurate values for color coordinates. When choices are to be made for the components of color TV screens, there is a definite need for objective evaluation of phosphors. Luminous efficiency is not difficult to measure for any phosphor, nor are the color coordinates of most broad-band emitters; this is true also for line emitters if the lines are few and well separated.[9] However, many experimental phosphors present complex line spectra which cannot be evaluated accurately as to color with currently available equipment. This inadequacy will soon be eliminated by equipment now being assembled.

The characteristic of the eye in the blue region is somewhat similar to that in the red, i.e., the spreading of energy over a broad band need not affect the hue, but for equal watts/watt efficiency, any such spreading reduces the luminous efficiency. If a *rare-earth blue* of the same hue is developed, it could replace the current blue phosphor without necessarily having equal power efficiency.

Another property of rare-earth ions (which would presumably apply to most, if not all other rare-earth phosphors) is the comparative insensitivity of the emis-

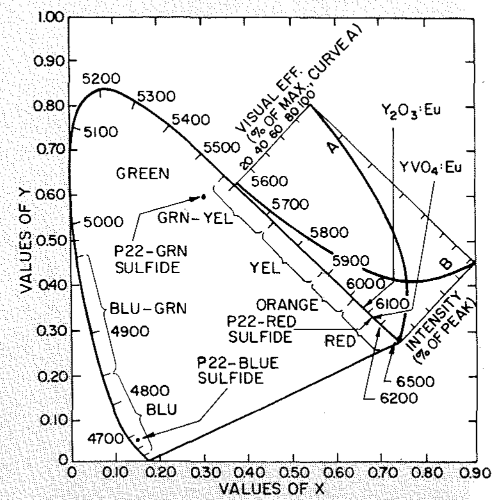


Fig. 7—Location on a chromaticity diagram of the emission colors of phosphors used in color TV. Also shown on the red-orange monochromatic locus are curves showing how the eye sensitivity and the broad-band emission of the red sulfide phosphor are mismatched.

sion spectra to the power density of excitation. As mentioned previously, the average broad-band emitter shows a tendency to color-shift. The rare-earth phosphors also have a greater tolerance to elevated temperatures of operation, but this aspect is not likely to play a role in color TV for some time.

Blue-Emitting Phosphors

During the development of blue phosphors, the brighter calcium-magnesium-silicate:titanium phosphor was displaced because of the smaller gamut the silicate allowed in relation to the zinc sulfide:silver. The zinc sulfide:silver, in hexagonal crystal form, has its emission peak at about 4350 Å. The cubic form of this material, used in most B/W screens and in color screens, has an emission peak of about 4500 Å, which causes it to have a higher luminosity than the hexagonal form.

Green-Emitting Phosphors

The most useful green phosphors have been zinc orthosilicate activated with divalent manganese (willemite) and the solid solutions (alloys) of zinc and cadmium sulfide, in the range of 35 to 40% cadmium sulfide, activated with silver and chloride.

Willemite has a narrower emission band than the green sulfides and peaks at about 5200 Å, circumstances which give it a higher saturation and lower luminosity than the sulfides. The zinc

TABLE I—Progress in the Development of Phosphors for Color TV

Types of Emitters	Phosphor Notation	Colorimetric Data x	y	X	Y(lum/watt)
Blue Emitters					
1. Calcium magnesium silicate:Titanium	$(Ca:Mg)SiO_4:Ti$	0.164	0.120	11.8	8.7
2. Zinc sulfide:Sliver:Chloride	$ZnS:Ag:Cl$	0.154	0.068	16.9	7.5
Green Emitters					
1. Zinc orthosilicate:Manganese (low)	$Zn_2SiO_4:Mn$ (low)	0.218	0.712	9.5	31.1
2. Zinc orthosilicate:Manganese (high)	$Zn_2SiO_4:Mn$ (high)	0.250	0.685	8.7	24.
3. Zinc aluminate:Manganese	$ZnAl_2O_4:Mn$	0.110	0.750	1.5	10.
4. Zinc cadmium sulfide:Silver:Chloride	$(Zn:Cd)S:Ag:Cl$	0.285	0.595	22.	46.
Red Emitters					
1. Cadmium borate:Manganese	$Cd_2B_2O_5:Mn$	0.630	0.370	18.3	10.7
2. Zinc orthophosphate:Manganese	$Zn_3(PO_4)_2:Mn$	0.674	0.326	14.5	7.
3. Zinc selenide:Copper	$ZnSe:Cu$	0.652	0.347	32.	17.
4. Zinc cadmium selenide:Copper	$(Zn:Cd)Se:Cu$	0.662	0.338	21.5	11.
5. Zinc cadmium sulfide:Silver:Chloride	$(Zn:Cd)S:Ag:Cl$	0.663	0.337	16.9	8.6
6. Yttrium vanadate:Europium	$YVO_4:Eu$	0.675	0.325	19.8	9.5

aluminate:manganese has an even narrower emission band than willemite, which accounts in part for its lower brightness. The properties of willemite produce a screen of large gamut but of low white brightness. Greater brightness, with reduced gamut, is achieved by replacing the willemite with zinc-cadmium solid-solution green emitters. The sulfide materials are usually made to have their emission peak near 5400Å. There is a slight shift in the emission peak of these sulfides to longer wavelengths during tube processing, due to some oxidation of the material.

Red-Emitting Phosphors

The synthesizing of red-emitting phosphors for commercial use has been a difficult problem because of the colorimetric considerations discussed above. The earliest red phosphor used by RCA, cadmium borate:manganese, was soon displaced by the redder zinc orthophosphate:manganese, resulting in a decrease in brightness but a larger gamut.

The zinc selenide:copper and zinc-cadmium selenide:copper were not used commercially, in spite of their superior brightness (as shown in Table I), because of screen application problems. Zinc-cadmium sulfide, of about 80% cadmium sulfide, with silver and chloride activators shows improvement in brightness over the phosphate. However, as was mentioned earlier, this red sulfide color shifts toward the yellow with increasing beam current.

Since all of the above red phosphors are of the broad-band-emission type, they suffer from the fact that the eye detects very inefficiently much of their radiation. As early as 1955 Bril and Klasens[8] realized that one way to avoid this problem of red phosphors was to use line emitters. As a result of extensive investigations of many rare-earth emitters in connection with the development of lasers, it was found that the rare-earth vanadates of the zircon structure (gadolinum, yttrium, and lutetium) doped with europium were phosphors of high luminosity under ultraviolet excitation.[10] These materials were then observed to have the necessary properties for a good TV phosphor.[11] Yttrium vanadate was chosen for purely economic reasons, since the other two materials have nearly duplicate properties.

Yttrium vanadate with europium activator has very close to the same luminosity as the zinc-cadmium-sulfide:silver red phosphor; when the intrinsic powder efficiency is measured, however, the sulfide suffers some degradation on tube manufacture resulting in the vanadate having a higher screen luminosity. Although the line emitter is much less efficient on a pure energy basis, the luminosity of the broad band and line emitter are nearly the same. Obviously, if the energy efficiency of the vanadate line emitter could be increased to that of the broad-band emitter, the luminosity would increase greatly.

Efficiencies of Rare-Earth Phosphors

Like the yttrium vanadate with europium activator described above, most phosphors containing rare earth are still considerably less efficient on an energy basis than the conventional broad-band emitters. In particular, the luminosity of the red line emitters would increase by about a factor of three, if this energy conversion were equalized. Increase in luminosity of blue could also be expected from an efficient line emitter, but it must be noted that practically all of the energy must be emitted in the blue region. As mentioned previously, green generated by line emission would have to consist of several appropriately spaced lines, in order to produce a screen of high white brightness.

One of the chief problems encountered in research on rare-earth phosphors is the control of the frequency (wavelength) of the line emission. As previously pointed out, most rare earths emit many lines, and it is most desirable to have emission in only one or a selected few. Results such as these have been observed by changes in crystal field and band gap of the host, and by adding a second doping agent. However, predictability and efficiency are still lacking.

To sum up, phosphors with rare-earth activators hold promise of increased luminosity and good monochromaticity.

FUTURE PHOSPHOR RESEARCH

The future course of phosphor study must of necessity take two paths: a) the empirical search for new or improved phosphors, and b) a basic attack on understanding the processes involved in the luminescence of solids, particularly under cathode-ray excitation. These approaches are exemplified by the intense efforts in the area of phosphors containing rare earths.

Rare earths in phosphors can have at least two functions: a) as efficient activators in new phosphor systems, and b) as probes for investigating the detailed nature of a luminescent center. However, although recent research has emphasized rare earths in luminescent materials, the potentialities of broad-band-emitting phosphors, with conventional (non-rare-earth) activators remain great enough to warrant further exploration.

As mentioned before, the rate of progress of understanding luminescence has been exceedingly slow. It is believed that this rate is actually being increased by using many of the complexities as probes for exploring various aspects of the luminescence process. Consequently, some of the answers to Seitz's questions may be forthcoming in the next decade.

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INTEGRATED CIRCUITS IN TV AND RADIO RECEIVERS

A Technical and Commercial Feasibility Study

A single monolithic-silicon integrated-circuit chip has been developed to perform the functions of amplification, limiting, balanced FM detection, and audio pre-amplification in the 4.5 MHz (Mc/s) intercarrier sound channel of TV receivers. This integrated circuit has been successfully produced by the Special Electronic Components Division and is scheduled for production use in a number of RCA television receivers. Although the principal emphasis has been on consumer product use of integrated circuits, the basic techniques and results described herein are broadly applicable to receivers used in industrial and military communications.

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SHORTLY after the start of the RCA Corporate Program in Integrated Circuits, the Home Instruments Division joined forces with the Special Electronic Components Division in Somerville to investigate the technical and commercial feasibility of using monolithic silicon integrated circuits in radio and television receivers. The driving force behind the program was the desire to exploit the potentially lower cost of integrated circuits in home instruments. The very complexity of TV receivers, particularly color receivers, made the cost-reduction potential attractive, since integrated circuits remove existing restrictions on the number of active devices which can be used economically to carry out a particular circuit function.

A second factor in the program was the possibility of better performance arising from the capability of integrated circuits to use as many transistors, diodes, and resistors as are required to attain ideal performance. The number of transistors and diodes that can be used in conventional circuitry is limited by cost.

A third factor was the expectation that the present high levels of reliability would be even further improved. Extensive military studies have shown that the reliability of a complete silicon integrated circuit can substantially exceed that of a conventional circuit using discrete components.

A fourth consideration was the miniaturization of complex functions made possible by integrated circuits. This is advantageous in personal radios, in small color TV receivers, and in potential new products that require complex electronic functions to be performed in a limited space, with high reliability, and at low cost.

Finally, significance was attached to the reduction in overhead that could be expected from the large reduction in the number of individual components required for a given function.

TEST VEHICLE

From the outset it was clear that for an integrated circuit to compete economically, it would have to perform the function of a complete subsystem. Accordingly, development was started on a suitable test vehicle—one that would encompass a sufficiently large number of functions to be able to compete with relatively inexpensive discrete transistors and with the low-cost associated resistors and capacitors. Such a test vehicle was essential as the means of debugging the entire program, from development concepts through engineering and mass production to use in the end product—the TV and radio receivers.

A survey of radio and TV receiver circuitry showed several areas where integrated circuits might be technically feasible and commercially advantageous. The number of areas was limited because it is not possible to integrate selectivity and, where coils cannot be eliminated by alternative circuitry, the necessary coils must be added externally. Similarly, where values of capacitance in excess of 50 $\mu\mu$f (total per chip) are required, these too must be external. Values of resistance in excess of 20,000 ohms in the current state of the art must be external unless they are eliminated by devising different circuitry. High-voltage and high-power sections of the receiver were considered unsuitable for integration because of the limitations of current integrated circuits.

The area of FM detection was selected as the most promising test vehicle for evaluating integrated circuits in home instruments for the following reasons:

a) It fully exploits the potential ability of an integrated circuit to perform a number of complex functions on a single chip. In this instance, the functions are amplification at high frequency, limiting to remove AM and noise, FM detection to reproduce the modulation, and audio preamplification.

b) It has the potential for lower cost because the functions are performed on a single chip instead of in separate stages using discrete transisors and components.

c) It has the potential for better performance. Conventional FM limiter-detector circuits are limited by cost considerations from using sufficient devices to do an ideal job, whereas an integrated circuit, which is not so limited, can be expected to perform better under fringe-area receiving conditions.

d) It is equally applicable to radio, black-and-white, and color TV receivers.

Consequently, the FM sound channel of TV receivers and the IF amplifier-limiter-detecter channel of FM radio receivers was selected as the first vehicle for exploring the feasibility of monolithic silicon integrated circuits in home instruments.

COUNTER-TYPE FM DETECTOR

The initial approach was an attempt to go all the way in integrating this FM function by using a digital-like counter approach to FM detection. This approach offered the possibility of eliminating the tuned circuits normally used in FM detectors, but it was found that a tuned circuit was still required to provide for the contingency of both the signal and noise momentarily being lost (because of the AGC time constant) during channel switching. In addition, tolerances on the values of R and C required in the multivibrator made the circuit difficult to integrate. These considerations led to discarding the counter in favor of an approach using an inte-

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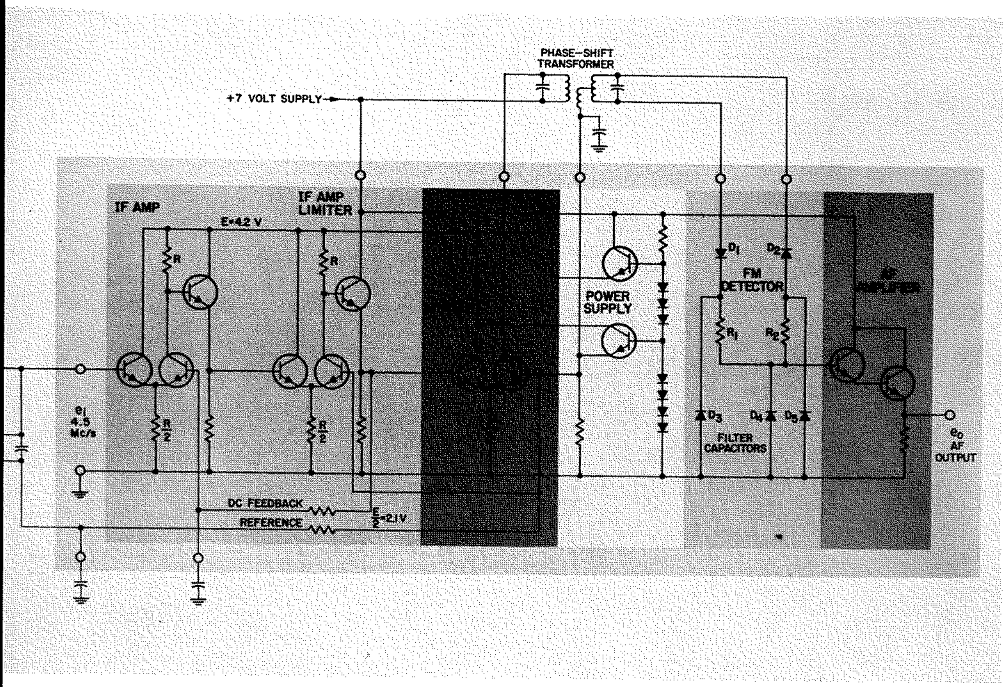


Fig. 1—Schematic diagram of integrated circuit chip used for intercarrier sound function in TV receivers.

grated discriminator network driven from an external tuned transformer. The evolution of this latter approach, which has been successfully developed for application in both radio and TV receivers, is described below.

FM INTERCARRIER SOUND TEST VEHICLE

The integrated circuit developed for the test vehicle role is shown in Figs. 1 and 2. The input signal, e_i, is the 4.5-MHz (Mc/s) FM beat between the 45.75-MHz picture carrier and the 41.25-MHz FM sound carrier produced at the video detector. The input transformer is a high-Q circuit that defines the pass band, eliminates spurious beat components, and improves the threshold sensitivity of the FM system by limiting the effective noise bandwidth ahead of the limiter stage. The phase shift transformer, driven from the output of the third emitter-coupled limiter stage, shifts the phase of the secondary voltage, normally in quadrature with respect to the primary, so that the phase shift follows the FM of the signal. The balanced detector network is followed by an emitter follower which provides the desired audio output signal at a low impedance level. A single-polarity, internally regulated voltage supply furnishes the required voltages for preamplifier, limiter, detector, and amplifier functions. The overall gain at 4.5-MHz is 75 dB.

Amplifier-Limiter

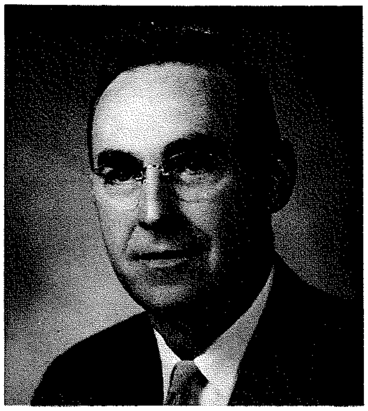
The amplifier (Fig. 1) consists of three direct-coupled cascaded stages. Each stage consists of a triad including an emitter-coupled amplifier and an emitter follower. The operating conditions are selected so that the DC potential at the output of the triad is identical with the DC potential at the input to the triad. This is accomplished by operating the bases at one-half the supply voltage and selecting the common emitter load resistor to be one-half of the collector load resistor. For this condition the voltage drops across the emitter and collector load resistors are equal, and the collector of the emitter-coupled stage operates at one V_{BE} above the common base potential $E/2$, so that the potential at the output of the emitter follower is also $E/2$. Accordingly the triads can be iterated, as shown in Fig. 1.

The operating conditions are selected so that the potential at the output of each triad will continue to be equal to the input potential, despite temperature changes in the integrated transistors and in the integrated resistors. In particular, V_{BE} changes are compensated because the minus unity common-mode gain of the emitter-coupled stage is compensated for by the plus unity gain of the emitter follower. The amplifier gain is independent of the absolute values of the load resistors; this characteristic is desirable since the absolute values of the integrated load resistors cannot be held to better than $\pm$ 20%. However, amplifier operation does depend on maintaining the ratio between the values of the emitter and collector load resistors; fortunately, it is characteristic of the integrated circuit process that resistor ratios can be held to within approximately 3%, which is sufficient. Thus, variations in the resistivity of the integrated resistors have only a negligible effect on the overall high-frequency cutoff of the amplifier, and this does not affect performance, since the cutoff lies beyond the operating frequency.

The emitter-coupled stages in Fig. 1 function particularly well as limiters because each half of the differential amplifier is alternately cut off on the positive and negative half cycles of the input signal. Or looking at it in a different way, the total emitter current, I_o, tends to stay constant, and on the axis of the signal the current is equally divided between the two transistors. On the positive half-cycle the current is steered so that the first transistor carries the full current, I_o, while the second transistor is cut off. Similarly, on the negative half-cycle the current is steered so that the first transistor is cut off and the second transistor carries the full emitter current, I_o. If the collector voltage supply is maintained at 4.2 volts ($6 \times V_{BE}$), the output-voltage collector swing is symmetrical about the zero signal axis, and symmetrical limiting is attained without spurious phase modulation.

The DC operating point is maintained by DC feedback around the first two stages, as shown in Fig. 1. The third stage is automatically held at the proper operating point because the feedback around the first two stages keeps the base of the third stage at $E/2$ volts, so that the third stage is balanced without being in the feedback loop. This is a desirable feature, because the tendency toward oscillation within the feedback loop is minimized when the number of stages within the loop is reduced. Because equal resistors are used in the base return circuit of the first stage, proper bias for the third stage is essentially independent of beta.

An internally regulated power supply provides the desired voltages for both the amplifier and the discriminator circuits. Two emitter followers supply $E =$ 4.2 volts and $E/2$ = 2.1 volts at low impedance; the emitter followers are driven from the voltage drops across the series diode network. This network keeps the gain relatively constant with changes in power-supply voltage. The system characteristics are essentially in-



JACK AVINS received his AB degree with honors from Columbia College and his MSEE from Polytechnic Institute of Brooklyn. Prior to 1941 he was engaged in the design of radio test equipment. From 1941 to 1946 he served as an officer in the Signal Corps. In 1946 he joined the Industry Service Laboratory, RCA Laboratories Division, where he was engaged in research and development of FM and TV receivers. From 1955 to 1956 he was Manager of the RCA Industry Service Laboratory in Zurich. Rejoining RCA Laboratories in 1957, he served as Manager of the Research Applications Laboratory and the Home Instruments Division Affiliated Laboratory. Since 1964 he has been on the staff of the Chief Engineer, Home Instruments Division, responsible for the development of integrated circuits for TV and radio receivers. Mr. Avins is a member of Phi Beta Kappa and a Fellow of the IEEE.

dependent of supply voltage over the range from less than 6 volts to more than 10 volts.

To prevent internal feedback, the first two amplifier stages within the feedback loop are operated from the low-voltage, 4.2-volt, center-tapped supply. The second-stage emitter follower, however, is driven from the unregulated 7-volt supply to prevent a degenerative feedback voltage from being developed across the output impedance of the 4.2-volt supply. The collectors of the balanced output stage are driven from the unregulated supply.

FM Detector Network

The FM detector is of particular interest because all the components associated with this network except the tuned phase-shift transformer are integrated on the same monolithic chip, along with the amplifier-limiter stages. The detector was designed as a sampling detector to eliminate the nonintegrable, large diode load capacitors normally used to obtain peak rectification in balanced phase-shift discriminators and ratio detectors. In the circuit of Fig. 1, average detection is employed with a substantially resistive load. The signal frequency and its harmonics are filtered by the distributed capacitance of the load resistors, and this filtering is further augmented by the capacitance of the small reverse-biased diode junctions, $D3$, $D4$, and $D5$.

By operating the detector into a substantially resistive load, an important advantage is realized: the loading of the discriminator diode load network on the secondary and primary windings of the tuned phase-shift transformer is reduced. In conventional discriminator circuits used for FM detection, the $\pm$ 20% variation in the value of diffused resistors substantially alters the peak-to-peak separation and linearity of the detector. In the circuit of Fig. 1, however, the loading reflected by the diffused load resistors is reduced to such a low level that the loading plays a negligible role in determining the discriminator characteristics; linearity and peak-to-peak separation are maintained over the full range of resistance values. Further, AM suppression and balance are maintained over this full range because the circuit is balanced and the diffused resistors are substantially matched in value, although their absolute values vary widely from wafer to wafer.

In addition to reducing the loading variations, the integrated detector eliminates the high-frequency spikes of RF interference that are characteristic of conventional phase-shift discriminators in FM and TV receivers. These pulses contain harmonics that are picked up by the internal antenna circuits or the IF amplifier input and cause undesirable interference with the incoming signal. In the integrated detector, such interference is greatly reduced because the detector load circuit is essentially resistive.

The frequency demodulation can be analyzed in terms of the switching action, since the junction of resistors R_1 and R_2 is periodically connected to the tertiary signal voltage at the center tap of the transformer secondary.

If E_1 = peak value of the tertiary voltage; E_2 = peak value of the secondary voltage; 2ϕ = angle during which rectifiers D_1 and D_2 are conducting; and θ = phase shift with frequency deviation from the center frequency (i.e. resonant frequency of the secondary tuned circuit), then

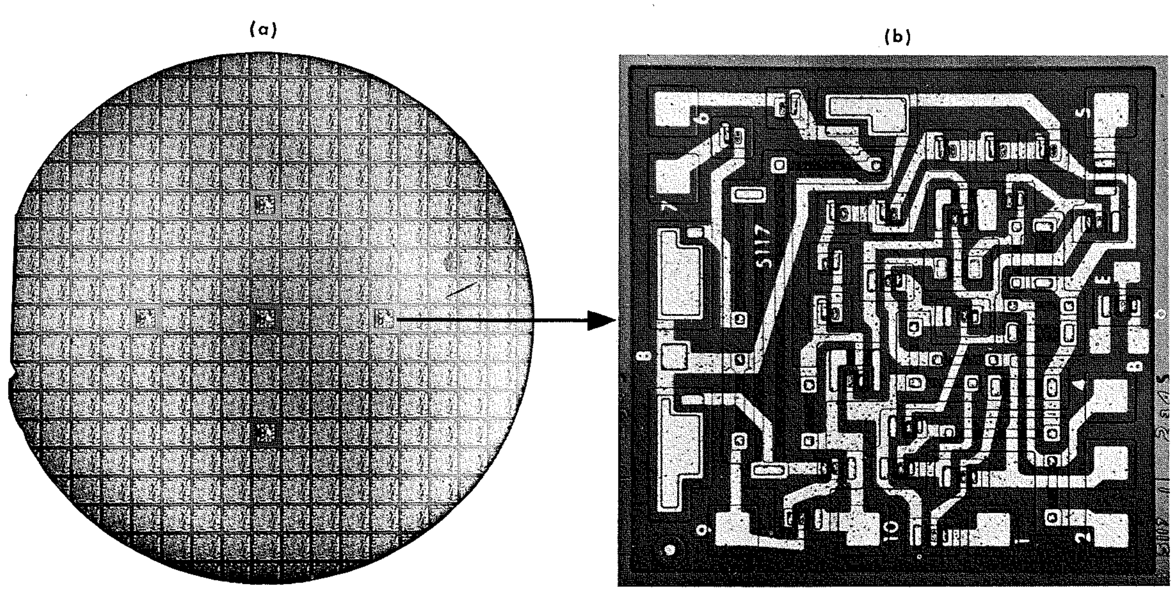


Fig. 2—Integrated circuit chip (a) and complete wafer (b).

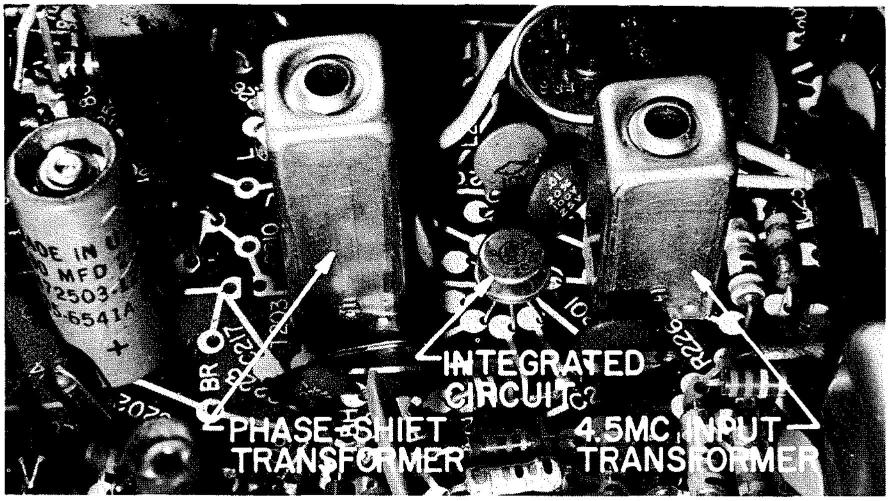


Fig. 3—Typical application of integrated circuit chip used for intercarrier sound function in RCA KCS-153 TV receiver.

$$E_{AF} = \frac{1}{2\phi}\int_{-\phi}^{+\phi} E_1 \sin(\omega t + \theta)\, d(\omega t)$$

$$= \frac{E_1}{2\phi}\Big[-\cos(\omega t + \theta)\Big]_{-\phi}^{+\phi}$$

$$= E_1\left(\frac{\sin\phi}{\phi}\right)\sin\theta$$

If the secondary voltage is large enough with respect to the diode *contact* potential to switch over 180°,

$$\phi = \frac{\pi}{2} \text{ and } \frac{\sin\phi}{\phi} = \frac{2}{\pi}, \text{ so that}$$

$$E_{AF} = \frac{2}{\pi} E_1 \sin\theta$$

Under these conditions, the demodulated output E_{AF} is independent of the amplitude of the secondary voltage.

Where the secondary voltage is not large enough to overcome the diode contact potential, the switching angle is reduced and $\sin\phi/\phi$ increases so that the demodulated output also increases as shown below:

Switching Angle	ϕ	$\sin\phi/\phi$
180°	$\pi/2$	0.64
90°	$\pi/4$	0.90
45°	$\pi/8$	0.98
0°	0	1.0

The effective sensitivity is increased, as compared with the conventional peak rectifier type of discriminator, by virtue of the reduced loading. It can readily be shown that the loading is reduced by a factor of 4.

The 2.1-volt supply voltage is applied through the tertiary winding to effectively bias the input and output diode network, as well as the direct-coupled emitter follower output amplifier. By operating the discriminator network at a positive potential with respect to the grounded substrate, it is also possible for the isolation junctions of the two signal diodes, $D1$ and $D2$, to stay reverse biased with respect to the substrate, even when signal voltage is applied to the circuit.

The 2.1-volt bias voltage applied to the secondary winding results in detector load resistors R_1 and R_2 being clamped at this voltage when a signal (or noise) is being received. In this manner diodes $D3$, $D4$, and $D5$ receive the desired reverse-bias potential and function as small capacitors of approximately 7 $\mu\mu$f each.

At the center frequency, the potential at the junction of R_1 and R_2 is substantially equal to the injected 2.1-volt bias voltage at the secondary winding, and this supplies the necessary positive bias voltage for the emitter-follower output stage. On either side of center frequency the voltage swings positively and negatively about the bias voltage in accordance with the frequency modulation.

APPLICATION TO FM RECEIVERS

The integrated-circuit approach for the intercarrier sound channel of TV receivers is directly applicable to broadcast FM receivers. The same considerations that make the integrated circuit of Fig. 1 desirable in the sound channel of TV receivers also make it attractive for broadcast FM receivers. The chip shown in Figs. 1 and 2 can be used directly at 10.7 MHz, the IF of broadcast FM receivers, to replace an IF amplifier, the limiter, and the FM detector stages of broadcast FM receivers.

This same integrated circuit can be used in communications receivers wherever a wideband limiter characteristic of exceptional flatness and freedom from incidental phase modulation is required. The FM detector network, being substantially resistive, can be used not only at 4.5 and 10.7 MHz, but also at low frequencies such as 455 kHz (kc/s). The same integrated circuit can also be used at frequencies as high as 50 MHz, although the gain of the wideband amplifier falls off, and at the same time the balance of the FM detector becomes more dependent upon an initially accurate balance of the detector load capacitances. However, if the initial design centers are accurate, the matching of component values inherent in the integrated circuit process insures that production circuits will perform satisfactorily.

APPLICATION IN RCA KCS-153 TV RECEIVER

The application of the integrated FM detector system to a TV receiver is illustrated in Figs. 3 and 4. Fig. 3 shows the intercarrier sound section of the RCA KCS-153 TV receiver; Fig. 4 compares functionally equivalent vacuum-tube and integrated-circuit boards. As compared with the original ratio-detector circuit, which used discrete transistors, resistors, and capacitors, a total of 26 discrete components are eliminated by the integrated circuit. The performance of the integrated circuit is at least equal in every characteristic, and superior in most. In particular, the AM-rejection ratio is so large that it cannot be measured with commercial FM-AM signal generators because of the incidental phase modulation of the generators. For input signals between 500 and 200,000 microvolts, the output signal is constant to within better than ± 0.5 dB. For signals between 1000 and 200,000 microvolts, the output variation is less than ± 0.1 dB. Because of the direct coupling in the three-stage amplifier-limiter and the absence of time constants which could charge on impulse noise, this steady-state performance is accompanied by comparably high AM suppression under dynamic conditions of impulse noise interference.

TESTING OF INTEGRATED CIRCUITS

A DC test program has been designed for the integrated circuit of Fig. 1, by which the essential characteristics of the amplifier, limiter, detector, output-amplifier, and power-supply sections are evaluated in a matter of seconds. Units that pass this automated DC test are almost certain to pass the dynamic signal test in which operation in the receiver is simulated by applying an FM signal and observing the flatness of limiting and the demodulated output. The ability to test the integrated circuit as a complete subsystem is of considerable advantage in simplifying the testing and lowering the cost.

It is common practice in discrete transistor technology to subdivide by parameter the total yield of a given generic type to permit the use of the complete distribution. In an integrated circuit, all circuits fabricated within the prescribed broad limits must yield a circuit which provides satisfactory performance. If selection were required, the yield would be reduced and the cost would be raised. This consideration imposes on the integrated circuit designer the requirement of devising circuitry that will result in acceptable system performance for substantially the full gamut of parameter variations characteristic of the integrated circuit process. For example, the circuit of Fig. 1 will exhibit acceptable performance for a beta range between less than 30 and more than 200. Since feedback capacitance is not a factor (the output of each stage is isolated from the input through the common emitter connection), feedback capacitance variations are of no importance.

PRESENT STATUS OF INTEGRATED CIRCUITS IN HOME INSTRUMENTS

The integrated circuit described herein as a vehicle to probe the technical and commercial feasibility of integrated circuits in consumer products has progressed to where it is now being mass produced by the Electronic Components and Devices Division for use in certain models of the 1966 RCA black-and-white and color TV receiver line.

Extensive life and field testing show that the goals of improved performance and reliability have been attained.

Cost is a function of the production yield, and the unusually high yields obtained thus far in production are encouraging. The high yields are attributable to high dependence placed on the matching of component characteristics while being able to tolerate wide variations in the absolute values of the integrated components. Thus, every integrated circuit chip that does not have a catastrophic failure can be expected to yield entirely satisfactory performance as a system. Although it is too early to predict costs accurately, it appears that the expected advantage of lower cost will be realized.

On the basis of our experience with the test-vehicle circuit, it is possible to assess the probable impact of integrated circuits in other areas of TV and radio receivers. Potentially, many of the low-level signal-processing sections of a radio or TV receiver can be replaced with integrated circuits. However, such a replacement on a stage-by-stage basis would be uneconomical and, in addition, would offer no advantage with respect to performance or reliability. Functional blocks must be selected, and *integrable* circuits must be designed to accomplish the desired functions—not necessarily in the same manner as they are performed with discrete components. The success of any given approach is dependent, among other factors, on the ingenuity of the circuit designer in devising integrable circuits capable of competing with efficient low-cost transistor circuits using discrete components.

In this task, a number of difficulties confront the circuit designer: the circuits must work with a limited range of resistance values, essentially without capacitors, preferably without inductors, and with a minimum number of connections external to the chip. On the other hand, the designer has several advantages working for him: transistors and diodes can be used freely, and transistor, diode, and resistor parameters will match to a high degree.

In view of the foregoing considerations, the penetration of integrated circuits into other sections of TV and radio receivers looks promising, and additional applications of integrated circuits, particularly in the low-level signal processing areas, can be expected. In the higher voltage, higher power areas, and in RF amplifier input circuits, it appears more reasonable to predict that discrete transistors and components will continue to represent a better economic solution than integrated circuits, at least for the present state of the art.

TEN YEAR CYCLE IN FM DETECTION SYSTEMS

In each of the past three decades a new FM detector system has been developed, has found almost universal acceptance in radio and TV receivers, and in turn has been displaced by a new detector with more desirable characteristics. This chronicle begins with the balanced FM phase-shift discriminator invented by S. W. Seeley of RCA in 1935.[1] Ten years later, it was largely superseded by the Seeley ratio detector,[2] which displayed improved AM-rejection properties and which is internationally accepted in radio receivers to this day. In 1955 another RCA circuit,[3] the locked-oscillator quadrature-grid FM detector, was developed for TV applications. This circuit has been widely used in the past decade in intercarrier sound TV receivers where its high sensitivity and high audio output (eliminating the audio driver stage) have made possible an efficient high-performance audio system.

Although it is too early to predict accurately how the integrated FM detector system described in this paper will fit into the record of progress in FM detector systems, it appears that the present integrated system will repeat history and displace its predecessors in FM, TV, and communication receivers because of its potentially lower cost, better performance, and greater reliability.

ACKNOWLEDGEMENTS

The author wishes to thank the members of the Integrated Circuits Engineering Department of the Special Electronic Components Division and the members of the Television Engineering Department of the Home Instruments Division who contributed to this project.

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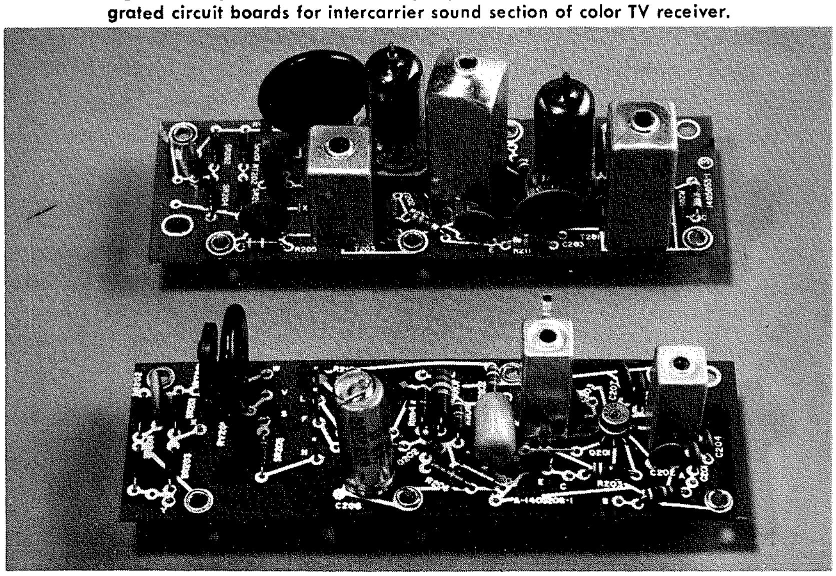
Fig. 4—Comparison of functionally equivalent vacuum-tube and integrated circuit boards for intercarrier sound section of color TV receiver.



A PORTABLE TRANSISTORIZED 12-INCH MONOCHROME TV RECEIVER

Transistorized TV receivers have been in the engineering laboratories for many years. Although some transistorized sets have been marketed, none have attained the performance necessary to meet RCA standards. This paper describes the design of a transistorized receiver that meets all RCA requirements. Good circuit design solutions were found for problems normally associated with transistor TV design. Deluxe performance, good reliability, and future adaptability have been provided. Performance data taken on actual production samples prove that RCA transistorized receivers are of commercially acceptable performance.

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Over the years television receivers have undergone marked changes in size and performance. Significant advances have been made possible by new components and circuit designs. Following this trend, the production of a transistor receiver was inevitable. The 12-inch transistorized portable TV receiver (KCS-153) described herein inaugurates an era of transistorized television by RCA.

Using newly developed transistors and new circuit designs, a 12-inch portable monochrome TV receiver has been designed through the combined efforts of several RCA Victor Home Instruments engineering groups, supported by RCA Electronic Components and Devices engineers. The new receiver, in quantity production for several months, has excellent performance characteristics and is designed to enter the consumer market as a serviceable forerunner of instruments to come. Design concepts are such that circuits developed for this chassis can be incorporated in future large- or small-screen television, as transistorized sets normally evolve.

TUNER CIRCUIT DESIGN

The 12-inch receiver uses solid-state circuits for both UHF and VHF tuners, making available the reception of channels 2 through 83. The KCS-153 major circuit stages are shown in Fig. 1.

The UHF tuner is a smaller version of the KRK-120 tuner used in tube receivers. Many physical changes were incorporated to achieve the desired reduction in size while maintaining high performance. The active elements used are an *npn* silicon transistor for the oscillator and a crystal diode for the mixer.

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The VHF tuner, a four-circuit type, employs three *npn* silicon transistors for the RF amplifier, oscillator, and mixer. The input to the RF amplifier is double-tuned, with a single-tuned circuit coupling the RF to the mixer. The RF amplifier is forward-biased and acts as an AGC amplifier for the IF amplifier. The gain and noise figure of the RF amplifier and mixer are optimized by circuit design. Preset fine tuning consisting of individually tuned coils for each channel is used for the oscillator.

Antenna input to the tuner is chosen by a switch in the back cover which selects either the built-in monopole rod or the two terminals provided for use with a 300-ohm input from an external antenna.

Problems inherent in transistor tuner design, such as cross modulation and overload, were considered and their effects were reduced by circuit innovations. The adequacy of the design is proven by the the fact that the receiver operates with an outdoor antenna without using an attenuator for strong signals.

VIDEO IF SECTION AND SECOND DETECTOR

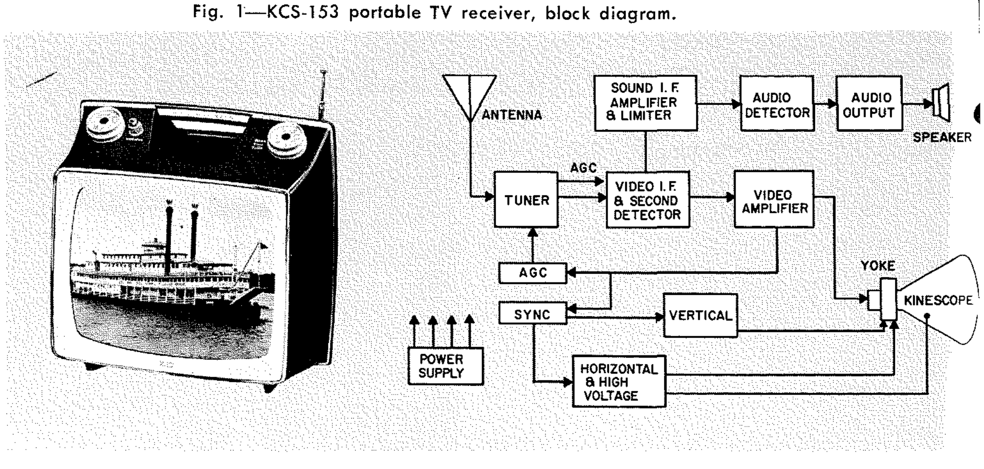
The video IF consists of three stages of amplification at 40 MHz (Mc/s) plus a detector providing 1.5 volts of video at the input to the video amplifier. All three amplification stages use *npn* silicon transistors in the common emitter configuration. The detector is a 1N60 germanium diode.

The tuner is link-coupled to the IF amplifier; the link provides the proper matching between the mixer and the first IF amplifier and a 47.25-MHz adjacent-channel sound trap. In the collector of the first stage, an absorption trap is used for the 41.25-MHz accompanying sound attenuation. The first amplifier is AGC controlled; the last two stages are fixed-biased for maximum gain and signal-handling capabilities. The coupling between the three stages is a form of RX coupling described in another article.[1] The second detector is a conventional tube-receiver detector elevated above ground so that bias can be applied to the first video amplifier. The overall IF and RF are readily aligned. The link circuit between the IF and mixer, including the 47.25-MHz trap, is aligned first. Next, the three IF stages are synchronously tuned to 43.5-MHz and the 41.25-MHz trap is set. This procedure yields an overall bandpass characteristic similar to that of normal TV receiver specifications. The fact that there is no variable neutralizing greatly reduces alignment time.

SOUND IF, AUDIO DETECTOR, AND AMPLIFIER

The 4.5-MHz sound IF is taken off at the first video amplifier, which functions as an emitter-follower to reduce loading on the second detector. Following the first video are two sound-IF amplifier-limiters coupled to a ratio

Fig. 1—KCS-153 portable TV receiver, block diagram.



detector. The resultant audio is fed to a two-stage, DC-coupled, class-A audio amplifier. This circuit provides ample gain with good AM rejection.

The audio output uses a 140-volt, B+ transistor driven by a low-voltage transistor. The two stages are DC coupled with DC feedback for stability and linearity. The voltage breakdown on the output device is 300 volts. Normally the output would require a protective circuit for peak voltage pulses on the primary of the audio output transformer. To keep the peak pulses down, the audio is made to operate with the 3.2-ohm speaker. If the speaker is disconnected, the B+ voltage to the transistor is removed to eliminate peak transients that could destroy the transistor.

Earphone operation with the high-voltage audio output is unique. Since the impedance of an earphone and its cable is greater than 3.2 ohms, a means was devised for connecting the higher impedance, lower power earphone. The final circuit is shown in Fig. 2. When the earphone is inserted it opens the grounded end of the transformer secondary, creating an AC loop consisting of the speaker, the transformer secondary, and R_1. This combination provides sufficient negative feedback to reduce the audio output so that the 50-mW earphone can be used while the speaker is still connected—and it keeps the peak voltage on the audio transistor within limits. The speaker output is negligible, even though it is still connected.

VIDEO AMPLIFIER

The video amplifier consists of two *npn* silicon transistor stages. The first stage is so designed that, in addition to supplying signals from its emitter and collector, it does not load the second detector. The first video collector supplies a positive-going video signal to the sync separator; the emitter is a low-impedance signal source for the video output stage, AGC, and sound IF.

A 150-volt silicon transistor in the video output stage is partially DC coupled to the emitter of the first video amplifier. The 4.5 -MHz sound trap is incorporated in the emitter of the video output stage to minimize 4.5-MHz edge agitation and to prevent the video output stage from loading the sound takeoff circuit.

A high-level contrast control is incorporated in the video output stage with partial DC coupling to the kinescope.

High-voltage arcing from element-to-element down the kinescope gun progressively to the cathode could cause an arc to appear across the video output transistor. Since transistors may be damaged by high-voltage arcs, this situation is adequately controlled with a neon bulb on the kinescope grid.

AGC, NOISE CANCELLATION, AND SYNC

A 28-volt gate pulse is applied to a *pnp* silicon AGC transistor in series with a diode and an arc-protection resistor. The DC output of the gated AGC transistor is applied to the forward-biased RF amplifier which is also a DC amplifier for the IF-AGC.

A zener diode is used as a reference voltage for the AGC-delay control. The adjustment of AGC delay control determines the maximum gain reduction in the reversed-bias first IF stage. This adjustment is, in effect, a change in the delay of the tuner bias, allowing the receiver to be adjusted for maximum protection against undesired picture interferences; at the same time, the receiver is kept free from *background snow* in medium signal areas. This circuit uses only one separate transistor for AGC.

A noise-separator diode is fed by the negative-going signal from the second detector circuit. This diode is biased by the AGC control so that only noise pulses above sync tips cause the diode to conduct. This arrangement eliminates the need for a separate adjustment of the noise-diode threshold and simultaneously tracks (as AGC is adjusted) the conduction of the noise-cancellation diode with the output of the second detector.

The output of the noise-cancellation diode is fed to a normally saturated, noise-cancellation transistor in the emitter circuit of the sync separator. This transistor stays in saturation except when impulse noise pulses are present.

The operation of the transistor sync-separator is similar to that of a tube sync-separator except for the different impedances required. The base of this transistor is driven from a positive-going video signal obtained from the collector of the first video stage.

The output of the sync separator synchronizes the vertical deflection circuit and drives a sync phase-splitter for the horizontal AFC.

HORIZONTAL-OUTPUT SECTION

The AFC diodes are driven from a horizontal-phase-splitter transistor that provides positive- and negative-going sync pulses. The AFC circuit is also supplied with a reference pulse derived from the flyback transformer. The DC from the AFC network is applied to the base of the horizontal-blocking-oscillator transistor.

The horizontal-blocking oscillator is an *npn* silicon unit which incorporates

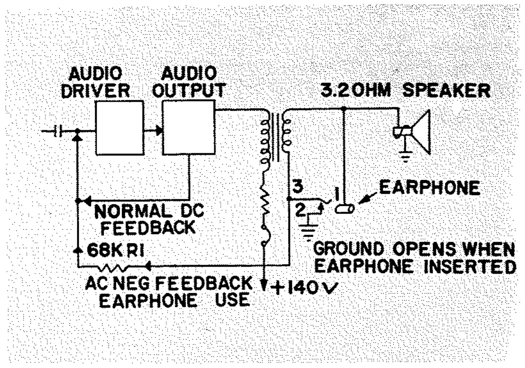


Fig. 2—Audio output with earphone operation, simplified schematic.

LOREN R. WOLTER received his BSEE degree from Iowa State College, Ames, Iowa, May, 1959. He is presently studying for a Master's degree at Purdue University, Indianapolis campus. Mr. Wolter joined RCA in 1959 as an engineering trainee. Following his training assignment, he joined the RCA Home Instruments Division, where he was engaged in the design of IF amplifiers for black and white television receivers. He is now a project coordinator in the design of black and white television receivers. Mr. Wolter also is an Editorial Representative for the RCA Engineer.

L. P. THOMAS received his BSEE from Clemson University, Clemson, South Carolina in 1947. He has since taken courses in business administration at the Wharton School of the University of Pennsylvania. Mr. Thomas joined the RCA television engineering section in 1947 and became responsible for the design and development of the video, sync separator, AGC, and noise cancellation circuitry of black and white television receivers. He also has designed remote-control circuits, automatic contrast and brightness control circuits, and other items related to the television field. Mr. Thomas has 19 issued patents plus patents pending, and he has written a number of technical articles and lectures. He is a senior member of the IEEE and a leader in the television engineering section.

Authors L. P. Thomas (left) and L. R. Wolter with one of the KCS-153 transistorized portables.

a thermistor in the base circuitry and additional stabilization circuitry so that Beta variations of the transistor over a wide range have negligible effect on circuit operation. Overall oscillator stability is generally superior to present tube horizontal oscillators.

The output of the oscillator is transformer coupled to the base of a horizontal-driver transistor. Waveshaping networks in the base and collector circuits of this stage provide the required drive pulse for the horizontal output unit. The horizontal output section is composed of the following major parts:

1) Switch (the horizontal output transistor)
2) Damper diode
3) Horizontal yoke and S-shaping capacitor
4) Width coil
5) Flyback transformer with additional windings for AGC, AFC, and kinescope blanking
6) Suppressors for RF ringing
7) Current-limiter transistor
8) Voltage limiter and kinescope screen DC supply, with protection in the screen circuit for kinescope arcing
9) A special second anode lead incorporating protection against high-voltage rectifier arcing
10) High-voltage rectifier.

Protection against kinescope arcing, screwdriver testing by servicemen, and high-voltage rectifier arcs is provided by protection circuits (items 7, 8, and 9 above). Without these protection circuits, peak power in the germanium horizontal-output unit could exceed 1 kilowatt under fault conditions that can occur in a receiver.

VERTICAL-OUTPUT CIRCUITS

The vertical circuit consists essentially of a switch, or oscillator transistor, followed by two isolation stages driving the vertical-output transistor. The entire circuit can be considered a combination multivibrator and Miller amplifier.

Silicon *pnp* transistors are used in the oscillator and pre-driver stages. The temperature-dependent leakage current of a germanium vertical-power-output transistor in a small receiver generally makes it difficult to deflect a 1⅛-inch, 110° picture tube. For the Miller circuit, this problem was solved by returning the bias network to a high B+ voltage. The vertical-output circuit includes the following features:

1) Anti-hunt circuit to prevent 30-Hz oscillation
2) Correction feedback for improved linearity
3) VDR (voltage-dependent resistor) stabilized bias voltage for the oscillator transistor
4) Thermistor in series with the yoke.
5) Waveshape feedback for improved oscillation stability
6) Anti-lockout circuit
7) VDR vertical output protection.

The AC and DC feedback and certain other features in this circuit are not discussed in this paper.

A person who thoroughly understands the tube art probably will have to spend more time understanding the vertical-output circuit than any other circuit in the receiver. The transistor circuits are by no means similar to tube circuits employing similar principles of operation.

POWER SUPPLY

Three basic supply voltages (+140 VDC, +30 VDC, and 6.3 VAC) are used in the KCS-153. The 140-volt source is used for audio output, video output, and bias voltages. The 30-volt supply is used for the remaining circuits, and the 6.3-volt AC supply is used for the kinescope filament. A circuit breaker protects the 140- and 30-volt supplies; a wire fuse protects the power transformer kinescope filament circuit.

A half-wave rectifier and RC filter make up the 140-volt supply, whereas a bridge rectifier and transistor filter supply the 30 volts. Stacking of the 140- and 30-volt supplies (Fig. 3) allowed the use of one breaker to protect both supplies. The active filter in the 30-volt supply provides a well-filtered supply in a compact space. A ripple voltage appearing at the filter input is coupled to the base of the filter driver; the ripple is amplified, inverted, and coupled to the filter transistor which cancels the effect of the input ripple. The filter provides a 30-volt supply with less than 50 mV of ripple.

The supply is not regulated, since adequate design precautions were taken to allow for the full changes in B+ voltage with line-voltage variations; no adverse effect has been observed on receiver performance. For continued high line-voltage operation, a 128-volt tap is provided on the transformer.

The total power consumption at normal line voltage is 60 watts. Electrolytics in the power supply and throughout the receiver were designed to handle all voltage and temperature changes, with added safety factors for excellent reliability.

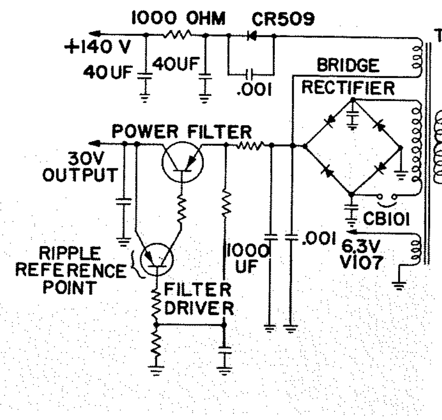


Fig. 3—Power supply, simplified schematic.

PERFORMANCE CHARACTERISTICS

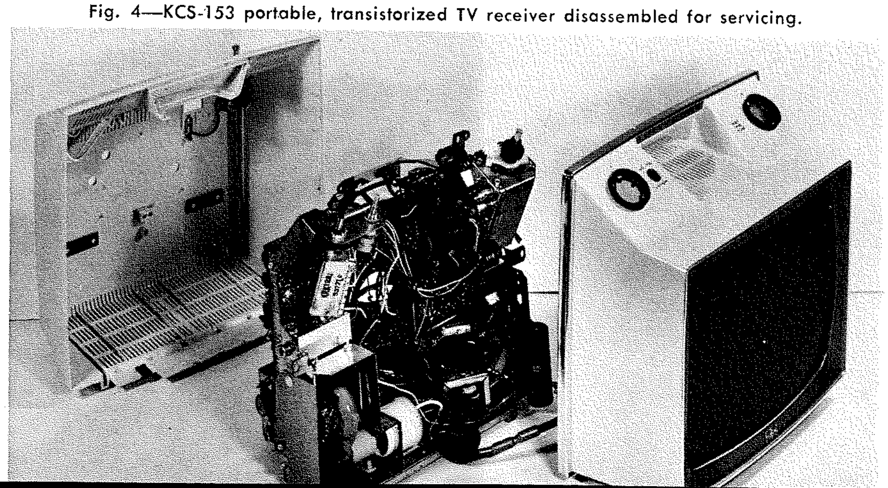
Although the word *deluxe* has been abused in the market place, it is used herein to designate the highest level of performance obtainable.

Considering picture-tube size, the state of the art in available solid-state devices, and industry-wide deluxe tube instrument performance, this 12-inch transistor set is truly a deluxe TV receiver.

Fringe-area performance, or the ability of a receiver to produce reasonable picture and sound from weak signals, is of prime importance in a deluxe receiver. Several of the major factors contributing to a high-performance receiver are:

1) Signal-to-noise ratio in the tuner
2) High sensitivity
3) Second-detector characteristics
4) Video drive with its associated amplitude and phase response
5) Kinescope characteristics and related circuitry and voltages
6) Noise immunity or freedom from undesirable jitter or movement
7) Noise-free sound level
8) Sound-picture tracking, or the ability of the receiver to receive the best picture and noise-free sound at the same fine-tuning adjustment
9) Commercially acceptable interference rejection from undesired stations
10) Minimum receiver generation of harmonics, *spooks*, or other undesired forms of radiation that will interfere with reception.

Fig. 4—KCS-153 portable, transistorized TV receiver disassembled for servicing.



Several of these items are discussed further.

Picture Sensitivity

There are various methods of measuring sensitivity. One method is to tune the receiver to the peak of the response curve, and measure the microvolts of picture-carrier frequency applied to the antenna terminals to produce 1-volt DC above noise across the second-detector load resistor. This method requires very little in the way of test equipment except for a good signal generator, and it is very useful in IF design work. A further advantage is that video frequencies are not required since only a carrier is used.

Another method of measuring sensitivity is to use a video-modulated carrier frequency and measure the microvolts of input signal across the antenna terminals to produce a certain video signal at the kinescope. This method, and the video-drive requirements of the kinescope, determine the overall sensitivity of the receiver that the viewer will observe. This method requires a rather elaborate setup, but it demonstrates the system performance of the receiver.

It should also be evident that in knowing the video gain and the sensitivity to the second detector, the overall gain can be determined fairly accurately.

The division of gain among the various stages is important. Impulse-noise immunity, AGC, sound performance, and the like are affected by the division of the gain among the stages as well as by the value of absolute gain. These factors were all carefully considered in this receiver.

The overall sensitivity of a receiver can be estimated without test equipment. When tuned to an unused channel, a receiver with adequate sensitivity produces a pattern of random clear dots or *snow noise*. A receiver designed with low sensitivity, or one that has dropped in gain due to other deteriorations, will produce a white raster without a snow pattern or will have a washed out appearance.

Kinescope High Voltage

There is an approximate rule of thumb for high-voltage requirements for kinescopes available at the time this set was released for production. That rule is *1 kilovolt per inch,* which generally insures a good spot with reasonable brightness.

Since the 12-inch receiver is portable, it will likely be used in relatively high ambient light conditions. To insure deluxe brightness, this 12-inch receiver has a 13.5-kV voltage supply.

Video Drive

Two stages of video amplification are used. Sync-sound and AGC are taken off the first stage; this allows some sync compression in the video output to obtain maximum picture information for the kinescope.

Approximately 100-volt peak-to-peak video is available in this receiver, a value more than adequate for a 12-inch kinescope. The video circuitry also includes a DC component. In addition to achieving adequate contrast, this receiver also has good phase and frequency response, producing a very good overall picture. Such features are usually missing in low-quality units.

Noise Immunity

Noise immunity is the ability of the receiver to operate with a stable picture in the presence of electrical interference, such as that caused by automobile ignition systems. The term noise immunity should not be confused with the signal-to-noise ratio of the tuner. Noise immunity is a system problem; included in the system requirements are adequate sensitivity, video-amplifier characteristics ahead of the sync and AGC amplifiers, and noise-inversion circuitry.

The 12-inch receiver uses a transistor noise inverter in the emitter of the sync separator to render noise pulses relatively ineffective. This deluxe feature assures better picture stability in fringe or medium-signal areas of reception. Several of the pertinent performance characteristics and features are shown in Table I.

TABLE I—Performance and Feature Summary

Overall sensitivity to 2nd detector	4.3 microvolts
Audio output at 10% distortion	0.5 watt
Video voltage gain	83 times
Peak-to-peak video output	100 volts
High voltage	13.5 kilovolts
Noise inverter	Yes
Gated AGC	Yes
Horizontal and vertical retrace blanking	Yes

SERVICEABILITY

When a new product, such as the 12-inch transistor TV receiver, is introduced into the consumer's home, the product should be easily serviced, and there should be adequate service facilities staffed with qualified personnel.

The RCA Service Company provides the facilities and trained personnel for servicing the KCS-153 TV receivers. To aid in the training of service personnel, the design group held lectures and assisted in the preparation of lesson plans, instruction manuals, and other training material.

To date, service people are well pleased with the serviceability of the receiver; it has proven to be an easy transition from tube sets.

Ease of servicing is mainly a result of good instrument design. The KCS-153 was designed with servicing as a major consideration. It was felt that a new product should be designed to provide an easy transition from present products. With this in mind, components used in the receiver were of known quality, reliability, and availability.

The KCS-153 receiver is made in three sections (Fig. 4); the front holds the kinescope, speaker, and yoke, and the chassis holds all other components except the monopole antenna in the back cover. Since the chassis contains the majority of the components, it can be removed from the cabinet for servicing. High-power transistors are mounted on the removable heat-sink bracket on the lower part of the chassis to obtain good ventilation and heat dissipation. The power devices use sockets for ease of replacement. Service positions for the tuner and high-voltage transformer simplify the servicing of components on the printed-circuit board.

CONCLUSIONS

The design of a 12-inch transistorized TV instrument that meets the demands of present TV receiving has been demonstrated by the actual production and use of the KCS-153 over the past year. Performance is essentially equivalent to the deluxe performance of tube receivers, and servicing is readily accomplished. Circuit design concepts can be applied to the design and production of a variety of transistorized receivers in the future.

ACKNOWLEDGEMENTS

The authors wish to express their appreciation to the engineers in Television Product Design for their comments and assistance in the preparation of this article. This receiver was designed by the Black and White Television Product Design Group and the Television Tuner Design Section. Major contributions were also made by the Indianapolis Components Section and the Electron Components and Devices Division.

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SERVICING THE TRANSISTORIZED 12-INCH PORTABLE TV RECEIVER

The introduction of transistorized television receivers into the consumer market will significantly affect the TV servicing industry. This paper discusses steps that the TV service technician must take to prepare himself for transistorized TV maintenance. Several appropriate trouble-isolation techniques are described.

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The KCS-153 12-inch portable is the first transistorized TV receiver released to the consumer market by RCA. The shift from tubes to transistors represents a significant change in receiver design, and it is expected to have an equally significant effect on the TV servicing industry. For one thing, the high reliability of transistors should result in fewer service calls per year per receiver. Secondly, the service required on the KCS-153 and similar future receivers will probably fall in the defective component category. This means that the service technician who is content with merely changing tubes is likely to find himself unemployed as the changeover to transistors progresses. Transistors and similar solid-state devices are expected to result in an increasing number of new consumer products, all of which will require service. Here again, however, this service will ultimately be performed only by those technicians who have trained and otherwise prepared themselves for this work.

In the case of the transistor receiver, this preparation breaks down into two main steps. The technician whose past experience lies primarily in the tube area must first familiarize himself with basic transistor theory. Once he has gained this basic knowledge, he must become familiar with those service techniques that are applicable to transistor circuits. In this area he will find that several of the procedures used on tube receivers will work equally well on transistor receivers. The basic service sequence is the same in both cases. That is, any service problem must first be localized to a section and then to a particular stage of the receiver; then the defective component within the stage can be isolated and replaced. The key to success in this sequence lies in the second step. If the technician can establish procedures to isolate the different problems to a specific stage, the final step of locating the defective component becomes relatively simple.

Final manuscript received December 30, 1965.

TROUBLE-ISOLATION TECHNIQUES

There are a number of techniques that may be used to isolate a problem to a specific stage. These may be broadly broken down into two groups: 1) techniques used principally by the shop technician, and 2) those used primarily by the technician who provides service in the home. There are, in addition, several procedures equally applicable to both types of technician. The difference between the techniques employed by these two types is essentially a function of the test equipment available. The shop man, for example, can use the voltohmyst, oscilloscope, and sweep generator, all normally part of a well-equipped shop, to localize transistor problems. With this type of test equipment, the experienced shop technician should (after a short indoctrination program) be able to quickly localize most of the problems on the 12-inch KCS-153 and future receivers.

The challenge in servicing this type of receiver lies in developing techniques that can be used in the field. Experience has shown that most customers are reluctant, from both cost and convenience considerations, to have their TV sets removed from the home for service. This customer reaction will remain essentially the same regardless of whether the receiver is of the tube or transistor type. Consequently, emphasis must be placed initially on familiarizing the field technician with localizing procedures that can be employed in the home. The problem is somewhat complicated by the limited amount of test equipment available to the field technician. The average field service technician usually carries only a multimeter, such as the RCA WV-38A, as his basic diagnostic tool. Despite this equipment limitation, there are several signal-injection and signal-tracing techniques that can be used in the field in addition to the basic voltage and resistance checks.

Signal-Tracing Technique

For example, assume that a field technician encounters a transistor receiver with normal sound, but no brightness. A check at the kinescope second anode shows that the *no brightness* condition is the result of no high voltage. This condition can be caused by trouble in any one of the blocks shown in Fig. 1. The multimeter, set to the AC output position, serves as the signal-tracing device in this application. The output position differs from the conventional AC voltage position in that a DC blocking capacitor is in series with one of the test leads.

The meter is first connected to the collector of the horizontal oscillator stage (TP-1, Fig. 1). A reading of approximately 5.0 volts at this point indicates that the oscillator is functioning normally. If no voltage is read, the trouble is probably in the oscillator or its associated control circuits. If a normal indication is obtained at the collector, the meter is moved to the base of the horizontal driver (TP-2). A reading of approximately 1.45 volts indicates that the interstage coupling between the oscillator and the driver is intact. The meter

Fig. 1 — Block diagram of horizontal-deflection and high-voltage section of TV receiver.

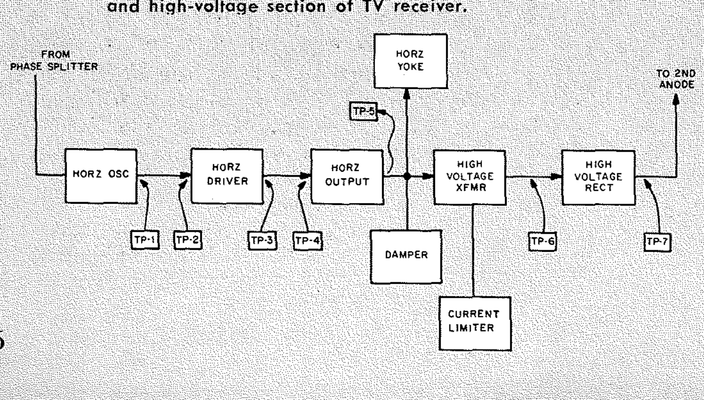
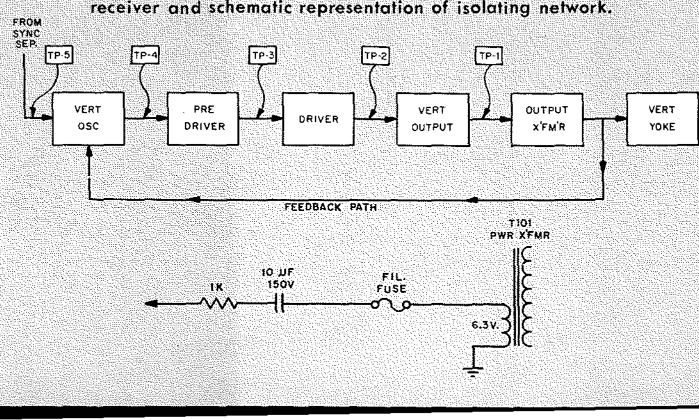


Fig. 2 — Block diagram of vertical-deflection section of TV receiver and schematic representation of isolating network.



H. G. SPENCER completed the basic electronics, basic radar, and advanced radar courses during a 3-year tour of duty with the U. S. Air Force. He joined RCA in 1947 as a Consumer Products field technician. He served as Chief Technician from 1952 until 1956, when he was promoted to District Training Administrator. He was appointed Regional Training Administrator in 1957. Mr. Spencer has been Home Office TV Training Administrator at Cherry Hill, N. J. since 1960. In this capacity he develops the technical programs used nationally to train Consumer Products television technicians.

is then moved in sequence to TP-3, -4, and -5, where readings of approximately 12.0, 85, and 85 volts, respectively, should be obtained. The trouble is localized between the last point at which a normal reading is obtained and the first point at which no reading, or an incorrect reading, is encountered. Should a normal reading be obtained through TP-5, the technician would switch the multimeter to the DC voltage position and, by means of the high-voltage probe, determine if the trouble is in the high-voltage transformer or the high-voltage rectifier stage. Once the trouble has been localized to a stage, the technician would make additional voltage and resistance readings within the defective stage to find the defective component.

Signal-Injection Technique

In the previous example, a signal-tracing technique was employed to isolate the problem. There are sections of the receiver in which a problem can best be localized by means of signal-injection techniques. For this application, the technician can use the 60-cycle 6.3 volts developed in the filament winding of the power transformer as the injection signal. To use this voltage, an isolating network consisting of a 10-microfarad capacitor in series with a 1000-ohm resistor is connected to the ungrounded end of the filament winding, as shown in Fig. 2. The capacitor provides DC isolation, and the resistor serves as a current limiter. The output indicating device is the picture tube or speaker.

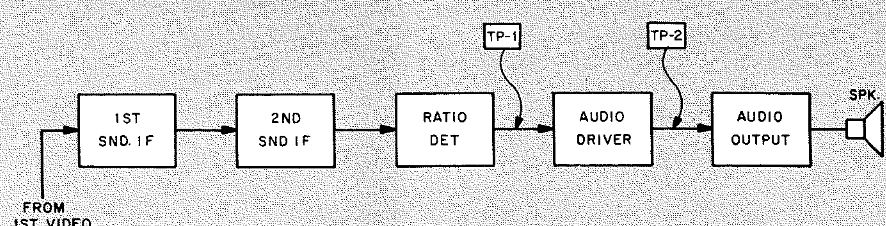
To illustrate this technique, assume that a defective receiver has normal sound and a single horizontal line across the screen. By evaluating the picture and sound information, the technician initially establishes that the trouble is in the vertical-deflection section of the receiver. In the KCS-153, a vertical-deflection problem can occur in any one of the six blocks shown in Fig. 2 or in the feedback path between the vertical output stage and oscillator. To localize the trouble still further, the 60-cycle, 6.3-volt signal is first applied to the primary of the vertical output transformer. If no deflection is noted on the picture tube, there is trouble in the output transformer or in the yoke, and this would be further localized by additional resistance readings. If a small amount of vertical deflection is obtained on the picture tube, the output transformer and yoke are functioning normally, and the 60-cycle signal is then injected at the base of the vertical output stage (TP-2, Fig. 2). If this stage is working normally, a somewhat greater amount of vertical deflection should be observed on the picture tube. No deflection indicates trouble within the stage. If a normal indication is obtained at TP-2, the 60-cycle signal is then injected at the base of the driver (TP-3). No increase in the amount of vertical deflection is normally obtained at this point, since both the driver and pre-driver stages are emitter followers. No vertical deflection at TP-3 indicates trouble in the driver or in the coupling between the driver and the output stage. To further isolate the trouble, the 60-cycle signal is fed to the emitter of the driver stage. If deflection is now obtained, the driver stage is defective. No deflection indicates trouble in the coupling between the driver and the output stage. In a similar manner, the signal is injected in sequence at TP-4 and -5. If the pre-driver and oscillator stages are functioning normally, a substantial amount of deflection will be obtained when the signal is fed to the base of the oscillator (TP-5). This large deflection indicates that all stages from the oscillator through the yoke are functioning normally. The *no deflection* problem in this instance must therefore be the result of a defect in the feedback network between the output and oscillator stage. As in the previous *no high voltage* example, once the problem has been localized to a stage with the 60-cycle injection technique, additional voltage and resistance readings are employed within the stage to determine the defective component.

The 60-cycle injection technique may also be used to localize a *no sound* problem (picture normal). The 60-cycle signal is first fed to the high end of the volume control (TP-1, Fig. 3). If the audio driver, output stages, and speaker are functioning normally, a 60-cycle hum will be heard from the speaker. In this case the trouble would lie in the first or second sound IF or in the ratio detector. If no hum is produced when the signal is fed to TP-1, the trouble is in the audio section and can be further isolated by injecting the 60-cycle signal at the base of the audio output stage (TP-2). Hum at this point indicates trouble in the audio driver stage. The absence of hum points to the audio output or the speaker as the source of trouble.

The 60-cycle injection technique can also be used in the video amplifier stages. The sequence here directly parallels the previous *no sound* example. The signal is initially injected at the base of the first video amplifier. If all stages from this point to the control grid of the picture tube are working normally, a 60-cycle hum bar will appear on the face of the picture. Absence of a hum bar indicates trouble in the video stage. This would be further isolated following a sequence similar to that used in the *no sound* example.

In the previous two examples, the 60-cycle injection technique initially isolates the trouble to the respective IF or output stages. When the trouble is in the IF stages, the technician must use voltage readings to further isolate the problem. Measuring the voltage developed across the emitter resistor in each of the IF stages is an effective means of localizing the inoperative IF stage. This voltage is directly proportional to the current flowing in the stage. A zero voltage reading across a given emitter resistor indicates no conduction within that particular stage. This emitter-resistor-voltage technique is equally applicable in both sound and picture IF stages. In making these and other voltage checks, two precautions must be observed: 1) use only a meter with a high impedance on both the AC and DC positions, and 2) connect the meter to the point to be read before applying power to the receiver. The latter practice will minimize the possibility of destructive transient pulses being introduced by a meter being connected to a circuit with the power applied.

Fig. 3 — Block diagram of sound-IF and audio section of TV receiver.



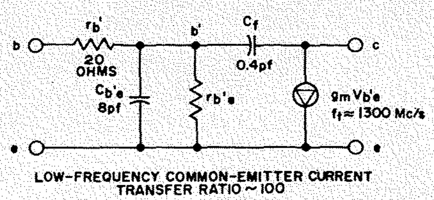


Fig. 1—Equivalent circuit of RCA 40235 silicon transistor.

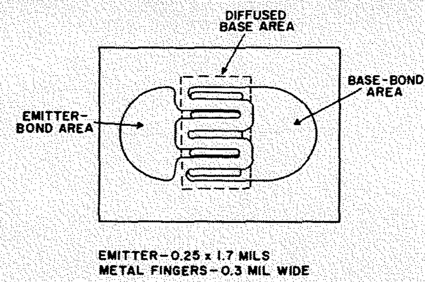


Fig. 2—Geometry of RCA 40235 silicon transistor.

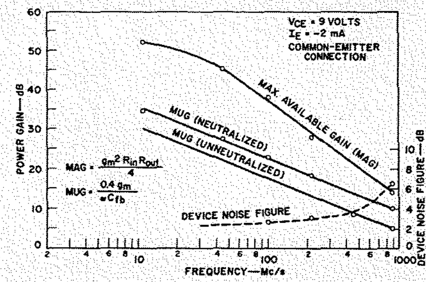


Fig. 3—Gain and noise of the 40235 transistor as functions of frequency.

TRANSISTORS FOR TELEVISION

The application of transistors to black-and-white television has required the development of many new transistor types. The electrical specifications for such types were established to provide desired levels of performance, reliability, circuit economics, or a combination of these factors. This paper describes some of these new transistor types, their application in black-and-white TV receivers, and problem areas in circuit design.

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Examination of the block diagram of a TV receiver reveals three major areas which warrant consideration for transistor applications: the tuner and IF-amplifier, the video, and the deflection portions of the receiver. The associated transistors which require special consideration are tuner types, video-output types, and horizontal- and vertical-output types. Although other functions in the receiver may require transistors with specific characteristics, these characteristics can readily be obtained by minor modification of existing types, or by selection from the types developed for the tuner, video, or deflection stages. For example, transistors for sync, AGC, AFC, and horizontal- and vertical-oscillator stages might be obtained by the first method; transistors for the picture-IF-amplifier section are obtained from tuner types.

TUNER TYPES

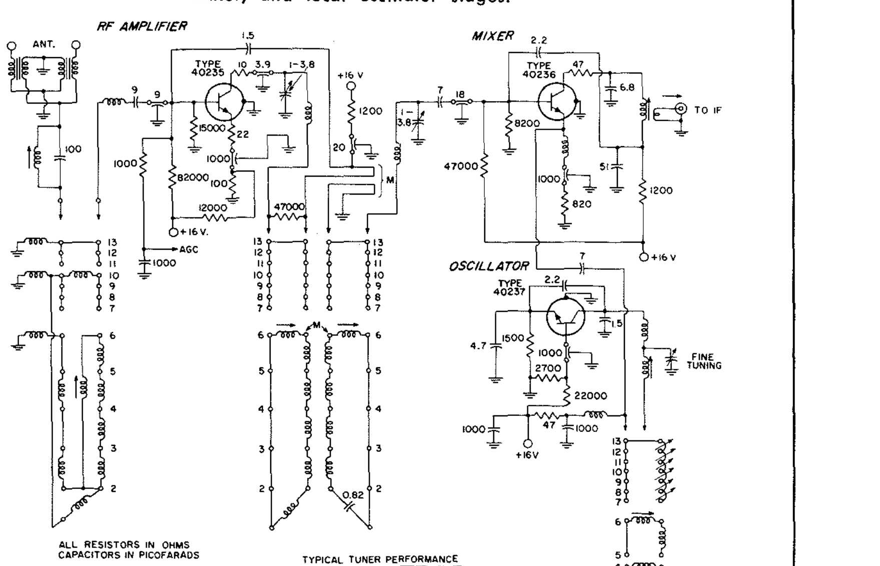
The primary requisites of transistors for TV tuners are high gain, low noise, wide AGC range, and good crossmodulation performance throughout the VHF band. The hybrid-pi equivalent circuit of an RF transistor is shown in Fig. 1. To provide high gain and low noise, the transistor should have low equivalent base resistance $r_{b'}$, high gain-bandwidth product f_T (characterized by low base-to-emitter capacitance $C_{b'e}$), high common-emitter current transfer ratio at low frequencies (β_o), and a low feedback capacitance C_f to minimize feedback. A metal package with a shield (ground) lead is used for tuner and IF types to minimize both collector-to-base feedback capacitance and radiation.

The active device structure used to meet these performance requirements is shown in Fig. 2. This transistor, the RCA-40235, is an *npn* silicon epitaxial planar device consisting of two emitter stripes and three base stripes in an interdigitated array. A metalization-over-oxide technique is used to minimize the active area with good resultant mechanical rigidity. The typical noise and gain performance of the 40235 over the VHF band is shown in Fig. 3. Power gain slightly in excess of 18 dB is achieved at 216 MHz (Mc/s), with a device noise figure of 3.1 dB.

Fig. 4 shows the circuit diagram of a VHF tuner in which three silicon transistors similar to the 40235 are used in the RF-amplifier, mixer, and local-oscillator stages. The typical power gain and noise figure of this tuner are 40 dB and 3 dB, respectively, on channel 2, and 38 dB and 3.5 dB, respectively, on channel 13.

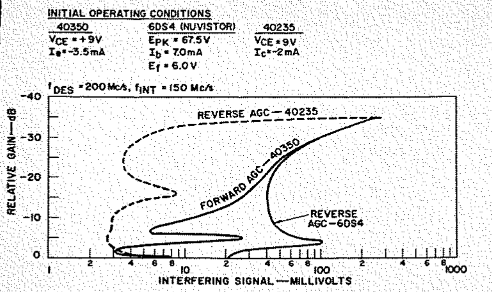
Unfortunately, transistors which have the high f_T and low $r_{b'}$ required to produce high gain and low noise figure produce the poorest crossmodulation performance. The low-pass filter comprised of $r_{b'}$ and $C_{b'e}$ in Fig. 1 provides very little attenuation of interfering signals at frequencies throughout the VHF band, as evidenced by a relatively small decrease of g_m with frequency through

Fig. 4—VHF tuner using silicon npn transistors in RF amplifier, mixer, and local oscillator stages.



TYPICAL TUNER PERFORMANCE	CH.2	CH.13
GAIN	40	38dB
NOISE FIGURE	3.0	3.5dB

Final manuscript received February 7, 1966.



NOTE: ALL INTERFERING VOLTAGES ARE NORMALIZED TO THE INPUT IMPEDANCE LEVEL OF THE TRANSISTOR, WHICH IS APPROX. 130 OHMS AT 200 Mc/s

Fig. 5—Cross-modulation characteristics of 40350 and 40235 silicon transistors and 6DS4 nuvistor tube.

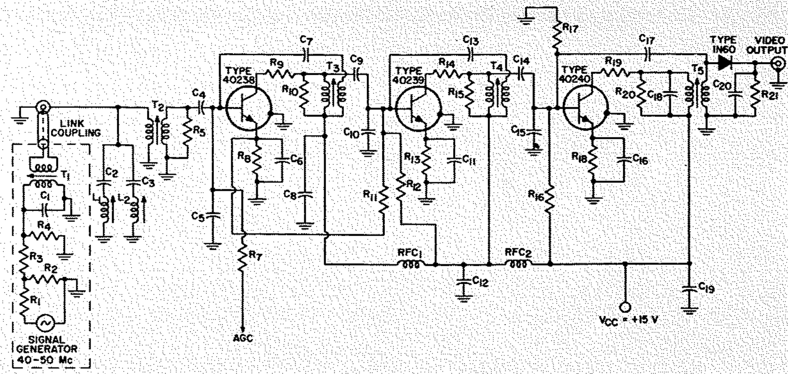


Fig. 6—Schematic diagram of 44-MHz IF amplifier.

this range. Fig. 5 compares the crossmodulation performance of the 40235 using reverse AGC bias with that of the 6DS4 nuvistor tube.

An improvement in the device crossmodulation with AGC can be obtained by use of forward AGC (i.e., increasing the emitter current to reduce gain rather than decreasing the current). Tailoring of the device is required, however, to ensure adequate AGC range. Characteristics of forward-AGC and reverse-AGC tuner types are listed in Table I. The value of maximum usable gain shown for the 40350 forward-AGC version is slightly

TABLE I—Characteristics of RF-Amplifier Transistors

	40350 (Forward AGC)	40235 (Reverse AGC)
Collector-to-Base Voltage, volts	+10	+10
Collector Current, mA	—3	—2
Input Resistance (Re Y_{11e}), ohms	130	200
Output Resistance (Re Y_{22e}), ohms	7,000	7,000
Extrinsic g_m ($\|Y_{21e}\|$), μS	47,500	43,700
Maximum Available Gain, dB	27.1	27.7
Maximum Usable Gain, dB (Single Stage, Neutralized)	18.5	18.1
Typical C_{re} (IMAG. Y_{12e}) 1 MHz, pF	0.5	0.5
Device Noise Figure, dB	3.4	3.1
Typical Gain-Bandwidth Product (f_T), MHz	900	1,300

higher than that shown for the 40235 because the 40350 must be operated in its maximum-gain operating condition at sensitivity (3 milliamperes). The improvement in crossmodulation when compared to reverse AGC is shown in Fig. 5. Further improvement in crossmodulation can be achieved by operating the device from a low source impedance, by using a small unbypassed resistance in conjunction with a low source impedance, or by using greater selectivity in the input circuits. (However, all of these methods sacrifice some gain and noise performance.)

Both forward- and reverse-AGC versions are required in the tuner and picture-IF-amplifier stages of TV receivers. The forward-AGC version is desirable in the RF-amplifier stage to provide good crossmodulation and overload characteristics with AGC. The reverse-AGC devices provide better performance in the mixer, oscillator, and last picture-IF amplifier because these stages are essentially large-signal stages and the g_m decreases rapidly with increasing emitter current (peak signal swings) in the 40350.

At the present time, three IF-amplifier stages are required to produce the necessary gain for acceptable fringe-area performance. The simplest IF amplifier is synchronously tuned with reverse AGC (i.e., all IF transformers are tuned to the same frequency, as shown in Fig. 6, and gain reduction is obtained by reducing the operating current of a transistor). Little or no bandpass shift or tilt occurs with reverse AGC, and performance is adequate for most black-and-white receivers. As in the tuner stages, the use of forward AGC together with a transistor designed specifically for this application increases input-circuit dynamic range capablity and improves crossmodulation (if it is a problem) of the IF-amplifier strip. However, the input and output impedances of a forward-AGC transistor drop drastically as AGC is applied (by as much as 10 to 1), and can severely modify the bandwidth and shape of the IF response. Judicious circuit design can minimize these effects and, in fact, use these changes in device parameters to permit the bandpass to tilt in a predetermined and desirable manner. For example, the IF response can be peaked (the bandwidth and noise reduced) at the picture carrier for fringe-area signals, and the bandwidth permitted to

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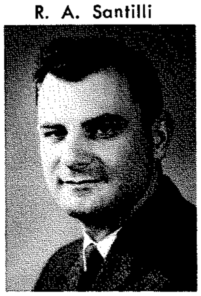
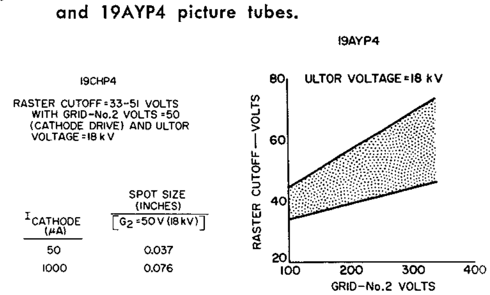
R. A. Santilli



H. Thanos

HARRY THANOS received the B.S. degree in Electrical Engineering from Purdue University in 1950 and the M.S. degree in Electrical Engineering in 1951. After one year at IBM, he joined the Admiral Corporation as head of the transistor department and held that capacity until 1956. He then joined Warwick Manufacturing Company as a senior research engineer working on applications of transistors to television circuits. Since January, 1958, he has been a member of the Commercial Transistor Applications Laboratory in Somerville, where he is currently engaged in applications work on high-frequency circuits utilizing developmental transistors. Mr. Thanos is a member of Eta Kappa Nu, Tau Beta Pi and Sigma Xi. He has published a number of papers on applications of transistors to FM and TV circuits.

Fig. 7—Performance characteristics of 19CHP4 and 19AYP4 picture tubes.



I CATHODE (μA)	SPOT SIZE (INCHES) [G2 = 300 V (18 kV)]	[G2 = 150 V (18 kV)]
50	0.032	0.037
1000	0.056	0.060

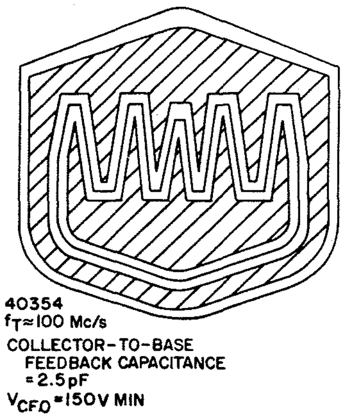


Fig. 8—Structure of 40354 transistor.

TABLE II—Approximate Requirements for 110° Deflection at 18 kV with Unregulated Power Supply. Supply Voltage Shown is at Nominal Line Voltage.

Supply Voltage, volts	12	18	24	36	48
Typical Value of Filter Capacitor, μF	5000	2000	1200	500	300
Peak Collector or Damper Current, amperes	18	12	9	6	4.5
Peak Permissible Collector Saturation Voltage or Damper Forward Voltage (based on linearity), volts	0.5	0.75	1	1.5	2
Damper Forward Voltage Required to Compare with 0.6 volt at 36-volt Supply, volt	0.2	0.3	0.4	0.6	0.8
Beta Required for 400-mA Drive	45	30	22	15	11
Peak Collector or Diode Voltage, volts	100	150	200	300	400
Yoke Inductance, μH	22	50	90	200	350
Capacitance for "S" Shaping, μF	36	16	9	4	2
Voltage Drop (L di/dt) in 1 inch of wire if peak collector current is changed in ½ μsec, volts	0.9	0.6	0.45	0.3	0.21
Relative Length of Wire Required to Produce a Given Ground-Loop or Lead-Length Effect	0.11	0.25	0.45	1	1.8

increase gradually with increasing signal and AGC. Also, a raised sound *plateau* in the IF response in the fringe-area signals increases the sound sensitivity to produce more audio quieting. Forward AGC requires more AGC power and produces more variations in bandpass shift and tilt than reverse AGC; as a result, it should be used only when advantage can be taken of its improved crossmodulation and overload characteristics.

VIDEO-OUTPUT TYPES

The specification of a video-output device depends to a great extent on the picture-tube drive requirements. Picture tubes presently used in tube receivers can be divided into two classes: 1) high-grid-No. 2-voltage types such as the 19AYP4, and 2) low-grid-No. 2-voltage types such as the 19CHP4. The raster-cutoff conditions (cathode-drive) for these tubes are shown in Fig. 7. The maximum raster cutoff of the 19CHP4 with an anode (ultor) voltage of 18 kilovolts is 51 volts at the recommended grid-No. 2 voltage of 50 volts, as compared to 77 volts for the 19AYP4 when operated at a grid-No. 2 voltage of 300 volts. Both of these tubes are 110°-deflection types with 1⅛-inch neck diameters. Larger-screen tubes rarely exceed these drive requirements; smaller-screen tubes with smaller neck diameters could require considerably less. As a result, designing for the 19AYP4, the most difficult approach, should be the goal.

Because the video-output device should be capable of supplying approximately twice the video drive required for the average picture tube (and one and one-half times that for the limit tube) and because the video content at best is only 75 percent of the video signal, a device with a breakdown of at least 150 volts is required. The drive requirements of the 19AYP4 may be reduced somewhat by operation at a somewhat lower grid-No. 2 voltage, as shown in Fig. 7. (The deterioration in spot size under this condition is no worse than that for the 19CHP4.) However, this possibility is not justification for reducing the breakdown requirements of the transistor, but rather should be used to improve the ratio of video drive capability to picture-tube drive requirement.

The load resistance for the video-output transistor is dictated by the video response desired, and values of the order of 4000 to 5000 ohms are typical. Application of DC voltage of 130 volts to the video-output transistor with this order of load resistance requires a device linear to 30 milliamperes and packaged so that it is capable of dissipating at least 1.3 watts with a reasonable heat sink at an ambient temperature of 55° C. The gain of the stage is primarily a function of the gain-bandwidth product f_T, the low-frequency common-emitter current transfer ratio β_o, and the feedback capacitance C_f; f_T and β_o should be as high as possible, and C_f as low as possible.

The structure selected for the video-output transistor is shown in Fig. 8. The device is a silicon *npn* transistor of high emitter periphery mounted on the header for good heat transfer from the collector junction to the heat sink. A radiating area of approximately 4 square inches (10-mil copper, 1 inch by 2 inches) is required to satisfy the thermal requirements up to 55° C.

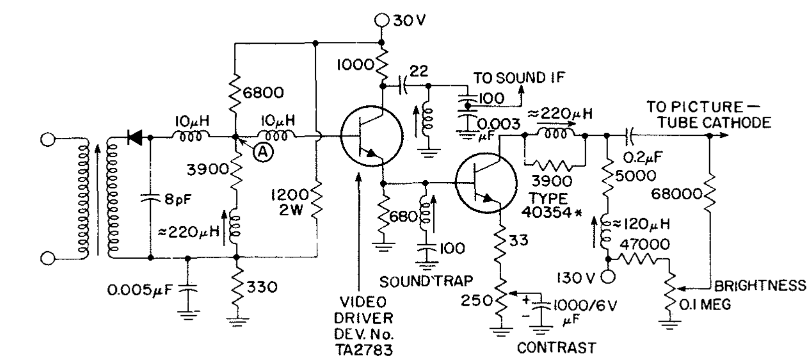
The circuit diagram of a typical video-output circuit driving a 19AYP4 is shown in Fig. 9. Low-level contrast is used and the sound and sync are obtained from the video driver to improve the drive capability of the output stage. Sync clipping can then be permitted in the video-output stage to provide more actual video drive. (The low-level contrast also eliminates the need for compensation of the contrast control for video-bandwidth change as the contrast control is varied, as required in high-level-contrast systems.) The voltage gain of the video circuit is approximately 100 with a 3-dB bandwidth of 3.2 MHz.

DEFLECTION TYPES

In the deflection portions of the receiver, reliability with special emphasis on economics and ease of manufacture are the primary considerations. Economics in this portion of the receiver is related to the power supply; for constant input-drive conditions to the output-deflection device and output-deflection energy, high supply voltages provide cost and fabrication advantages with respect to the output transistor, the damper diode, and the yoke. The disadvantage of high supply voltages is increased transistor breakdown voltage. Table II shows the effect of power-supply voltage on a horizontal-deflection circuit that uses a common-emitter transistor operated in a conventional trace-driven manner. It is obvious that higher supply voltages are desirable. In addition, circuit efficiency is generally improved and circuit layout becomes less critical with increased supply voltage.

If the power-supply voltage is increased to 150 volts, some cost reduction

Fig. 9—Schematic diagram of video amplifier circuit.



INPUT IMPEDANCE AT A≈3900 OHMS
BANDWIDTH≈3.2 Mc/s @ −3dB
VOLTAGE GAIN≈100

*HEAT SINK REQUIRED (4 SQ. IN. RADIATING AREA, 10—MIL COPPER) may be realized by elimination of the power-supply transformer (hot chassis); however, severe breakdown requirements are placed on the semiconductor device. Until solid-state devices having breakdown voltages of 1 to 2 kilovolts are commercially available, the advantages of *hot-chassis* operation cannot be realized. In many lower-deflection-energy systems, operation from a 12-volt supply may be desirable for mobile applications. B-boost techniques may be employed to take advantage of a higher supply voltage.

The horizontal-deflection transistor recommended for black-and-white TV receivers is a diffused-collector, graded-base *pnp* germanium power transistor. Fig. 10 shows the circuit diagram of a typical horizontal circuit using this device, the RCA-2N3731. The circuit consists of a blocking-oscillator stage (2N2614), a driver stage (2N3732), and a horizontal-output stage (2N3731). (The collector-to-base voltage rating of this 10-ampere horizontal-output transistor is 320 volts minimum; with a retrace time of approximately 12 microseconds, therefore, the power-supply voltage is limited to 30 to 35 volts.) This circuit is capable of conservatively providing the yoke-deflection current and high voltage for a 19-inch (19AYP4), 114° receiver from an unregulated power supply. The peak collector current is 5.5 amperes, and the typical flyback voltage (third-harmonic tuned) is 250 volts with a 12-microsecond retrace time. The collector-current turn-off time, typically 0.5 microsecond, results in low turn-off-time losses.

During development of the 2N3731, devices with appreciably higher breakdown voltages were fabricated and evaluated with the goal of a higher supply voltage. However, the higher breakdown voltage was accompanied by longer collector-current turn-off times and poorer high-current performance. Since the longer turn-off times produced higher turn-off-time losses, decreased efficiency, increased device heating, and poorer reliability, a compromise on breakdown voltage was made.

Early experiences with transistor deflection systems indicated a lack of reliability even though devices were apparently operated well within ratings. It was found that abnormal operation of the deflection circuit caused instantaneous operation of the transistor in excess of its ratings, which could result in destruction of the transistor. Typical circuit malfunctions are second-anode arcing (internal picture-tube arcing), high-voltage rectifier arcing, and corona or arcing of the secondary winding of the flyback transformer to ground. In normal circuit operation, the transistor operates as a switch; there is very little voltage across the device when current is flowing through it, and vice versa. In general, abnormal operation of the circuit permits excessively high currents to flow in the presence of appreciable collector voltage for an unpredictable period of time. Peak instantaneous dissipations of the order of 2000 watts are readily demonstrable. The total energy the device is subjected to is important, and the time of operation in these high-dissipation states is unpredictable. Destruction is possible well within the voltage and current ratings of the device if the length of time is sufficient. Since the transistor may be destroyed by operation outside its capabilities, even by one such operation, methods of protection must be employed to assure that the operating conditions during abnormal operation do not deviate appreciably from normal conditions. (Although the subject of protective circuit techniques is too lengthy for discussion in this paper, a more complete description of the destructive mechanisms and corrective techniques is available in the literature.[1]) Protective techniques may be very subtle and need not be particularly obvious in a circuit design. The circuit shown in Fig. 9 incorporates protective circuitry. The flyback transformer is isolated from the device through the small coupling capacitor *C*. This decoupling capacitance permits the forward transfer of high-frequency energy for high-voltage generation and minimizes the reflection of lower-frequency components as a result of abnormal conditions in the secondary circuit of the flyback transformer. This capacitor, in conjunction with the unregulated power supply which permits the power-supply voltage to collapse when a malfunction demands appreciably more current from the power supply, provides one method of protection.

Although the horizontal device functions primarily as a switch, a vertical-output type may be obtained from the same family, provided the basic device linearity and breakdown are adequate. The typical vertical-output circuit shown in Fig. 11 is a class AB amplifier which is driven with a shaped sawtooth signal. The 2N3730 output device is obtained from the horizontal-deflection family, but it is specified for vertical-output service. The sawtooth signal is generated by an RC charging circuit from a 220-volt supply. The 40231 oscillator acts as a switch, triggered by the flyback pulse to discharge the capacitor for the next cycle. The sawtooth signal is shaped and fed to the 40232 amplifier. The high input impedance of the 40232 permits the use of a high-impedance timing-and-shaping circuit that eliminates the need for large values of timing capacitors. This circuit complements the horizontal-deflection circuit shown in Fig. 10 to provide a complete deflection package for a 19-inch, 114°, 18-kilovolt receiver.

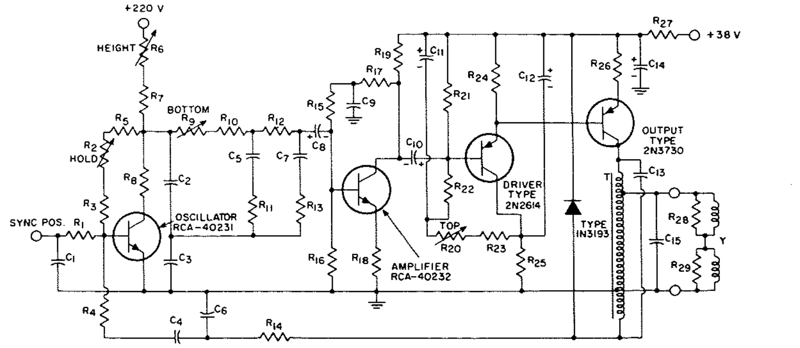
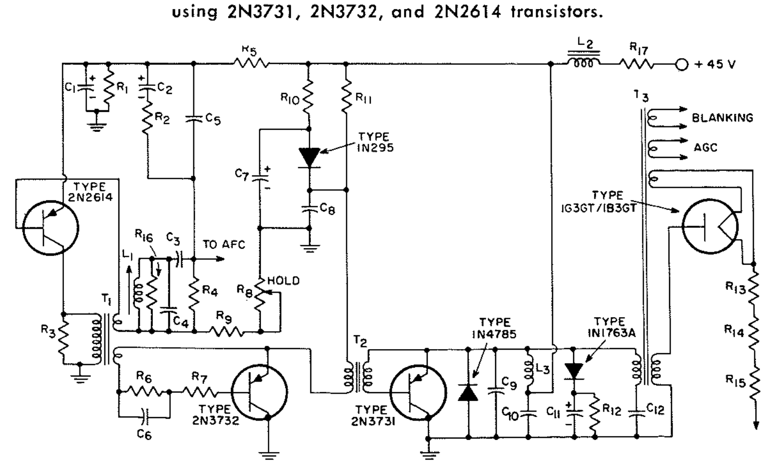


Fig. 11—Typical vertical-deflection circuit using 2N3730 transistor.

Fig. 10—Typical horizontal-deflection and high-voltage circuit using 2N3731, 2N3732, and 2N2614 transistors.



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THE 2N4012 OVERLAY TRANSISTOR

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THE RCA-2N4012 is the first commercial microwave power transistor that can provide a minimum power output of 2.5 watts as a frequency tripler at an output frequency of 1 GHz (Gc/s) and a collector efficiency of 25 percent. This transistor is one of the RCA overlay transistor family and is designed to operate in military and industrial communications equipment as a frequency multiplier in the UHF or L-band range. The 2N4012 transistor can be operated as a doubler, tripler, or quadrupler with watts of output power at frequencies in the low-giga-hertz-per-second range. This device has made possible the use of a single transistor to replace both the transistor power amplifier and the varactor-diode frequency multiplier previously required.

In frequency-multiplier circuits, this overlay transistor operates simultaneously as a power amplifier to provide gain at the fundamental or drive frequency, and as a varactor diode to generate harmonics of the drive frequency. The capacitance of the collector-to-base junction in an overlay transistor varies nonlinearly with collector voltage, much as varactor junction capacitance varies with the diode junction voltage. It is this nonlinear capacitance that permits harmonic generation in overlay transistors.

Fig. 1 shows the power output capabilities as a function of output frequency for a typical 2N4012 transistor used in the common-emitter circuit configuration for frequency doubling, tripling, and quadrupling. In a common-emitter doubler circuit, the transistor delivers 3 watts of output power at 800 MHz (Mc/s) with a conversion gain of 4.8 dB. In a common-emitter tripler circuit, it delivers 2.7 watts of output power at 1 GHz with a conversion gain of 4.3 dB. In a common-emitter quadrupler circuit, it delivers 1.7 watts of output power at 1.2 GHz with a conversion gain of 2.3 dB.

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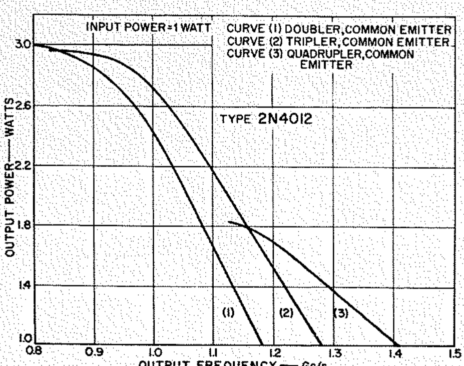
Fig. 1—Power output vs. frequency for RCA-2N4012 operated in doubler, tripler, and quadrupler circuits.



OPERATION

The mechanism of amplification and frequency multiplication in overlay transistors can be considered in two sections: one that is capable of gain at the fundamental frequency, and another (in which the collector-base capacitance serves a varactor) that is capable of multiplication. Transistors suitable for multiplier applications must be capable of delivering power with gain at the fundamental frequency and of converting the power from the fundamental frequency to a harmonic frequency. A good multiplier transistor, therefore, must first be a good UHF transistor capable of high power output, gain, and efficiency. Second, its varactor section should also have minimum losses in order to provide maximum conversion efficiency.

The figure of merit for the amplifier portion of the transistor is given by f_{max}, the maximum frequency of oscillation, as follows:

$$f_{max} = (PG)^{1/2} f = \frac{1}{4\pi}\left[\frac{1}{r_{bb}' C_c \tau_{ec}}\right]^{1/2} \quad (1)$$

where PG is the power gain, f is the fundamental frequency of operation, r_{bb}' is the base spreading resistance, C_c is the collector capacitance, and τ_{ec} is the emitter-to-collector transit or signal delay time.[1]

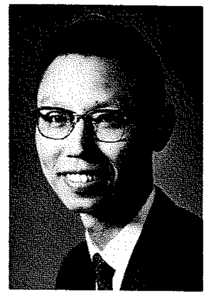
The efficiency of the varactor portion formed by the collector-base junction is determined by f_{VCB}, the cutoff frequency, as follows:

$$f_{VCB} = \frac{1}{\pi C_{min} (r_b' + r_s)} \quad (2)$$

where C_{min} is the minimum collector-base capacitance and r_s is the collector series resistance.

The 2N4012 overlay and internal structures are shown in Figs. 2 and 3; Fig. 4 is a cross-sectional view on which the varactor elements are indicated schematically. The minimum collector-base capacitance, C_{min}, consists of two parts. The major part, which comprises the active portion of the varactor, is made up of the capacitance formed by the collector-base junction not opposite emitter sites. This part of the capacitance is called the outer collector capacitance, C_o. The second part of C_{min} consists of that part of the collector-base junction which is opposite the emitter-base junction. This part is called the inner collector capacitance, C_i. C_o is a much more efficient varactor than C_i, because C_i has to charge and discharge through r_{bb}' and r_b', as well as through r_{si}, whereas C_o has to charge and discharge only through r_b' and r_{so}. Since the intrinsic base spreading resistance, r_{bb}', is much greater than the extrinsic base spreading resistance, r_b', there is a larger difference in the cutoff frequency, f_{VCB}, for the two parts of C_{min}. The large difference in r_b' and r_{bb}' arises from the difference in sheet resistance in the two

G. J. Gilbert



H. C. Lee

GEORGE J. GILBERT received his B.S. degree in Physics from Georgia Institute of Technology in 1958. He joined RCA Electronic Components and Devices the same year, and since then has worked on the design and development of high-frequency transistors, including the 2N3375 and the 2N4012. He is currently engaged in design and development of a 10-watt, 1.2-GHz overlay transistor.

H. C. LEE received his BSEE from Massachusetts Institute of Technology in 1952, and the M.S. degree in Electrical Engineering from Stanford University in 1953. Since joining RCA in 1958, he has had extensive experience in circuit design for UHF power and small-signal transistors and in microwave power generation using varactor diodes. He is responsible for the design of direct frequency-multiplier circuits using overlay transistors, and for the development of amplifier and frequency-multiplier circuits at S-band using overlay transistors.

Fig. 2—Cross-section showing overlay structure.

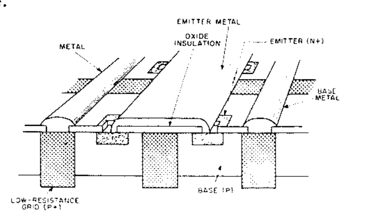
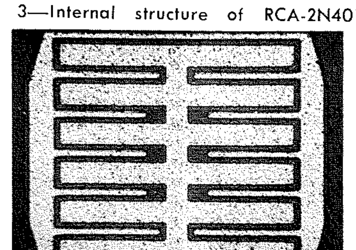


Fig. 3—Internal structure of RCA-2N4012.



areas. The sheet resistance under the emitter, which forms r_{bb}', is several thousand ohms per square; the sheet resistance between the emitter and base contacts varies from 5 to 100 ohms per square. Fig. 5 shows an equivalent circuit for the varactor portion of the 2N4012. In this circuit, r_{so} and r_{si} comprise the collector series resistance associated with the inner and outer collector capacitances, C_i and C_o.

Because of the unique features of the overlay transistor, the emitter area is one-tenth the base area and, hence, $C_i = 0.1\ C_o$. The inefficient portion of the varactor formed by the collector-base junction opposite the emitter sites is almost negligible because of the reduced emitter area. This reduction is accomplished without sacrifice of the current-handling capability of the transistor because of the large emitter periphery. Previous interdigitated structures had the emitter area almost one-half of the base area and, as a result, exhibited poor varactor performance.

The varactor cutoff frequency, f_{VCB}, is also maximized by minimizing the collector series resistance, r_s. This resistance is kept to a minimum by the NN^+ epitaxial structure used for the collector region. The N epitaxial layer forms the dominant part of the collector series resistance. The thickness of this layer is kept to the minimum value which will provide the required collector-base breakdown voltage. Lowering the resistivity of the epitaxial layer would also lower r_s, but it would *raise* C_{min}, and, therefore, yield no improvement in f_{VCB}. Measurements made on the 2N4012 at a V_{CB} of 65 volts give the following typical values[2]: $r_b' = 0.1$ ohm; $r_{so} = 1.8$ ohms; and $C_o = 3.5$ picofarads. The resulting varactor cutoff frequency, as calculated from these values, is 24 GHz. The combination of an overlay transistor having an f_{max} of 800 MHz and a varactor having an f_{VCB} of 24 GHz results in device operation above the f_{max} of the transistor.

APPLICATION AND PERFORMANCE

The diagram of Fig. 6 shows a circuit that triples from 340 MHz to 1.02 GHz, using an RCA 2N4012. This circuit uses lumped-element input and idler circuits and a coaxial-cavity output circuit. The transistor is inside the cavity and its emitter is grounded to the chassis. A pi section in the input (C_1, C_2, L_1, L_2, and C_3) provides impedance matching at 340 MHz between the driving source and the base-emitter junction of the transistor. L_2 and C_3 provide the necessary ground return for the nonlinear capacitance of the transistor. L_3 and C_4 form the idler loop for the collector at 340 MHz. L_4 and C_6 form the second-harmonic idler circuit for the collector at 680 MHz. The output circuit consists of a foreshortened coaxial cavity, 1¼ by 1¼ inches square. A lumped capacitance, C_5 (Johanson Type JMC 2954), in series with a ¼-inch hollow-center conductor of the cavity near the open end, permits adjustment of the electrical length. Output power at 1.02 GHz is obtained by direct coupling at a point near the shorted end of the cavity.

The graph in Fig. 7 shows the output power of this tripler at 1.02 GHz as a function of the input power at 340 MHz taken with a typical 2N4012 operated at a collector supply voltage of 28 volts. The solid-line curve is the power output obtained when the circuit is retuned at each input power level. The dashed-line curve is obtained with the circuit tuned at an output level of 2.9 watts. An output power of 2.9 watts at 1.02 GHz is obtained with 1 watt of drive at 340 MHz. The 3-dB bandwidth measured at this power level is 2.3 percent. Spurious frequency components measured at the output are better than 20 dB down. The variation of output power and collector efficiency with collector supply voltage at an input drive level of 1 watt is shown in Fig. 8. These curves are obtained with the circuit tuned when the collector voltage is 28 volts and the power output is 2.9 watts. With 1 watt of drive, the 2N4012 transistor in this tripler circuit provides a minimum output of 2.5 watts and a maximum of 3.5 watts. The median value is 2.8 watts.

A 340-MHz amplifier using the same circuit configuration and components as those of Fig. 6 was constructed to compare amplifier and tripler performance. The conversion efficiency for a large number of 2N4012's was measured. The conversion efficiency is defined as the power at 1.02 GHz obtained from the tripler divided by the power at 340 MHz obtained from the amplifier with the same input power (1 watt). The efficiency varies between 60 and 75 percent and averages at 65 percent, comparable to a good varactor in this frequency range.

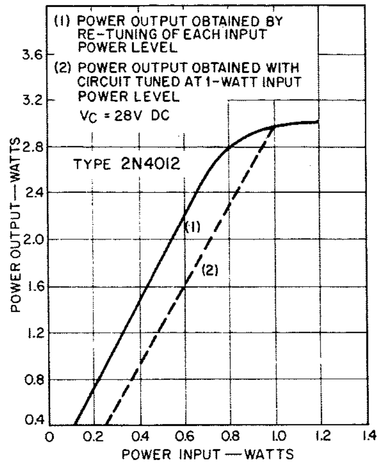


Fig. 7—Power output vs. power input; RCA-2N4012 tripler circuit (340 MHz to 1.02 GHz).

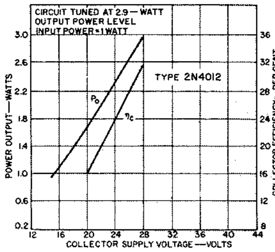


Fig. 8—Power output and collector efficiency vs. collector voltage; RCA-2N4012 tripler circuit (340 MHz to 1.02 GHz).

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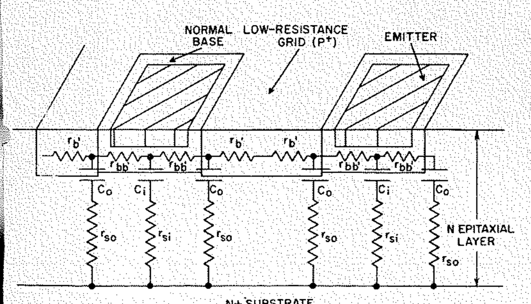

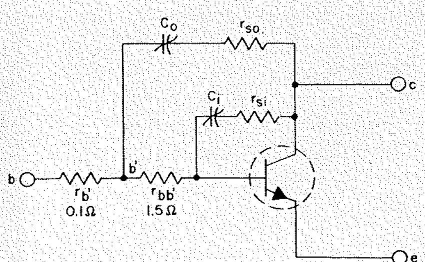

Fig. 5—Simplified circuit for varactor portion of RCA-2N4012.

Fig. 4—Cross-sectional view of RCA-2N4012 with varactor elements shown schematically.

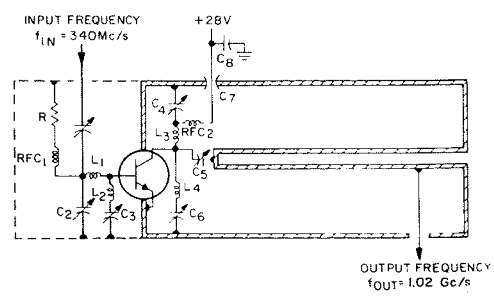

Fig. 6—RCA-2N4012 tripler circuit (340 MHz to 1.02 GHz).

DESIGN OF A TRANSISTORIZED TV PICTURE IF AMPLIFIER

This paper describes the design of a transistorized TV picture IF amplifier. Transistor and tube circuits are compared, practical design equations for determining the R-X and coupling transformers are presented, and AGC considerations and IF design are discussed.

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Ever since the first transistor radio receiver appeared on the counters of retail stores, increasing effort has been expended to produce transistorized television receivers. Such transistorized instruments have been confined mainly to laboratory development, primarily because of continuing tube improvements, both in performance and cost.

In recent years, however, there has been an increased improvement in the transistor itself as well as a greater understanding of the theory behind the transistor. Thus, competitively priced, transistorized TV receivers can now be designed to take their rightful place in the market. Because of the different characteristics of transistors and tubes, new design techniques were developed to achieve optimum performance of all transistorized circuits. The picture IF amplifier circuit is no exception, and its design procedure is described herein.

COMPARISON BETWEEN TUBE AND TRANSISTOR

To establish the best design approach for the IF amplifier, the pertinent transistor parameters must be first understood. Relative merits of transistors may best be established by comparing the transistor directly with pentode tubes currently used in IF designs. Typical comparisons between an RCA-4JD6 tube and an RCA-TA2562 transistor are given in Table I.

TABLE I — Tube Vs Transistor Comparison

	4JD6	*TA2562*
Input resistance (R_i)	5 kΩ	0.55 kΩ
Output resistance (R_o)	100 kΩ	40 kΩ
Input capacitance (C_i)	10 pF	5 pF
Output capacitance (C_o)	3 pF	1.0 pF
Transconductance (G_m)	14 mA/V	85 mA/V
Maximum feedback capacitance (C_f)	0.018 pF	0.55 pF
Max. available gain (MAG) $\text{MAG} = \frac{g_m^2 R_i R_o}{4}$	44 dB	45 dB
Max. usable gain (MUG) for single-stage amplifier $\text{MUG} = \frac{0.4\, g_m}{\omega c_f}$ @ 45 MHz	30.3 dB	22.1 dB
Gain-bandwidth[1] $(\text{GB}) = \frac{g_m}{4\pi\sqrt{C_i C_o}}$ *(matched capacitance transformer)*	204 MHz	——
GB Product $= f_t$	——	900 MHz

Final manuscript received January 12, 1966.

From Table I, it is evident that in a tube high-frequency IF amplifier, the gain per stage is limited primarily by feedback and gain-bandwidth product (GB); this is especially true with many other types of tubes where GB can be substantially less than 200 MHz (Mc/s). The current design practice is to use neutralizing and double-tuned circuits, thus increasing the design potential of the tube.

In transistor design the first apparent limiting factor is feedback.[2] Typical gain-per-stage figures for TA2562 are given in Table II.

TABLE II — Gain Per Stage at 43 MHz (2 mA Bias) for TA2562

	Gain per Stage in dB		
Condition	*Stage 1*	*Stage 2*	*Stage 3*
Unneutralized	22.7	20.5	18.7
Fixed neutralizing	27.8	25.6	23.8
Unilateralized MAG	43	43	43

Here again, the gain per stage may be increased up to MAG by using more complex neutralizing circuitry. Since the GB product from Table I is much higher than that of an equivalent pentode, the comparison may suggest some superiority of the transistor in the high-gain, medium-bandwidth amplifier circuits. However, the additional peculiarity of semiconductor devices must also be considered. In the present state of the art, transistors exhibit great variations of input and output resistance. Typical spread (n_i) between R_{max} and R_{min} is 4:1 for input, and 6:1 for output resistance. When fixed-turns-ratio transformers are used for coupling, perfect matching for gain cannot be consistently obtained; furthermore, the effect on bandwidth would render the maximum-gain amplifier unusable.

To eliminate such effects on the bandwidth, sufficient loss must be inserted into the coupling circuit. For producing designs similar to those found in present tube receivers, the effects of transistor parameters on bandwidth were assumed to be ±15% for television IF amplifiers, i.e. ±7.5% for input and ±7.5% for output impedance variations (Appendix 1).

To determine the amount of external loading, the equivalent circuit in Fig. 1 will be considered; circuit elements are designated as follows:

R_i = Mean input resistance of transistor T_1

R_o = Mean output resistance of transistor T_1

R_s = Driving resistance

R_L = Load resistance

$$n_i = \frac{R_{i_{max.}}}{R_{i_{min.}}}$$
maximum spread of input resistance

$$n_o = \frac{R_{o_{max.}}}{R_{o_{min.}}}$$
maximum spread of output resistance

$$\delta_\Delta = \frac{\text{Fractional Bandwidth Spread}}{2};$$
in this case = 0.15

$$R_1 = \frac{R_s \times R_i}{R_s + R_i} = \quad (1)$$
total input circuit resistance

$$R_2 = \frac{R_L \times R_o}{R_L + R_o} = \quad (2)$$
total output circuit resistance

From Appendix 2, the following relationship may be derived:

$$\frac{R_i}{R_1} = \frac{n_i - \frac{1}{n_i}}{2\delta_\Delta} \quad (3)$$

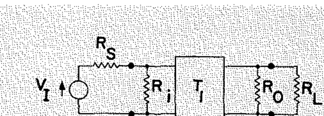

Fig. 1—Transistor input and output resistance.

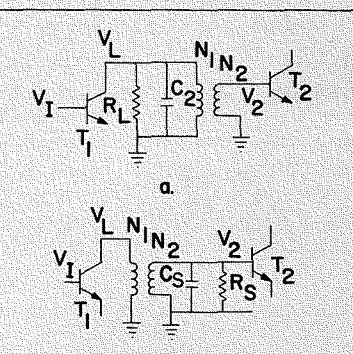

Fig. 2—The coupling transformer.

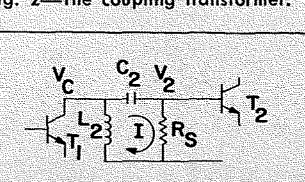

Fig. 3—The RX transformer.

$$\frac{R_o}{R_2} = \frac{n_o - \frac{1}{n_o}}{2\delta_\Delta} \quad (4)$$

From equations 3 and 4, input and output stability factors may be determined.[3]

$$IS = \frac{R_i}{2R_1} = \frac{n_i - \frac{1}{n_i}}{4\delta_\Delta} \quad (5)$$

$$OS = \frac{R_o}{2R_2} = \frac{n_o - \frac{1}{n_o}}{4\delta_\Delta} \quad (6)$$

It should be remembered that R_i and R_o are mean input and output resistances derived from the maximum-resistance-parameter spread limits; these values need not be identical with typical values published in manufacturers' data. Both sets of values should be investigated in the final design.

Insertion loss is called P_L and may be derived from:

$$P_L = OS \times IS \quad (7)$$

From equations 5 and 6,

$$P_L = \frac{\left(n_i - \frac{1}{n_i}\right)\left(n_o - \frac{1}{n_o}\right)}{16(\delta_\Delta)^2} \quad (8)$$

From equations 7 and 8, an alternative figure for maximum usable gain may be defined (MUG_Δ) where bandwidth spread is the limiting factor:

$$\text{MUG}_\Delta \text{ in decibels} = (\text{MAG} - 10 \log P_L) \quad (9)$$

Example 1

In the design where $\delta_\Delta = 0.15$, the input and output resistance spreads are $n_i = 4$ and $n_o = 6$, respectively.

From equations 3 and 4,

$$R_1 = \frac{2\delta_\Delta}{n_i - \frac{1}{n_i}} \quad R_i = \frac{R_i}{12.5} \quad (10)$$

$$R_2 = \frac{R_o}{19.5} \quad (11)$$

$$IS = 6.25, \text{ and } OS = 9.8 \quad (12)$$

$$\text{MUG}_\Delta = 45 \text{ dB} - 10 \log P_L = 45 \text{ dB} - 18 \text{ dB} = 27 \text{ dB} \quad (13)$$

From the example shown above, and by comparison with Table II, it will be evident that MUG_Δ should be considered in the design whenever MUG_Δ is in the same order of magnitude as MUG. Further, since parameter spread between various transistor types is in the same order of magnitude, the P_L values are similar for a given design quality, and the MAG may be considered as an approximate figure of merit for the device employed, at least in preliminary sample comparisons. From equations 10 and 11, the numerical values for the denominator » 1; hence, referring back to Fig. 1, equations 1 and 2 can be approximated as follows:

$$R_s \approx R_1 \quad (14)$$

$$R_L \approx R_2 \quad (15)$$

COUPLING TRANSFORMER DESIGN

The simple single-tuned interstage coupling transformer, shown in Fig. 2a, consists of a transformer with a turns ratio $N = \frac{N_1}{N_2}$; it is tuned with C_2 and has total circuit damping which is presented by R_L; Fig. 2b shows the same transformer with pertinent components referred to the secondary winding.

The usual requirements for the stage are:

gain $= G$

$$\text{3-dB bandwidth} = B = \frac{f_o}{Q} = \frac{\omega_o}{2\pi Q}$$

The preliminary design procedure for the coupling transformer follows the theoretical considerations, and is given here:

For optimum mismatch from equations 3 and 4,

$$N = \sqrt{\frac{R_2}{R_1}} = \sqrt{\frac{R_o\left(n_i - \frac{1}{n_i}\right)}{R_i\left(n_o - \frac{1}{n_o}\right)}} \quad (16)$$

For the transformer shown in Fig. 2,

$G = \frac{g_m R_2}{N} \approx \frac{g_m R_L}{N}$, using the approximations from equations 15 and 16 (17)

$$B = \frac{\omega_o}{2\pi Q} \quad \text{where } Q = R_L \times \omega_o C_2$$

$$\text{Thus } B = \frac{1}{2\pi R_L C_2} \quad (18)$$

Hence the design factors of N, R_L, and C_2 are determined by equations 16, 17, and 18 respectively, giving complete transformer design description. A similar procedure is used with double-tuned transformers, modifying the procedure for well-established, double-tuned requirements.[4]

DESIGN COMPROMISE

The preceding theory and design procedure applies mainly to the picture IF amplifier design, where bandwidth limitation and repeatability are of primary importance. Factors such as noise figure, cross modulation, and variations of operating conditions with AGC will affect the final choice of IS and OS. In the case of noise considerations, the power loss in the transformer coupling circuit will degrade the system noise figure. In overall TV receiver system design, the IF amplifier is preceded by RF and mixer stages;[5]

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thus, considerable amounts of input mismatch may be tolerated without degrading overall system performance.

Where the gain requirements are less than MUG_Δ (by examination of equations 5, 6, 7, and 8), greater design flexibility will be achieved.

With the high GB products available, it is also possible to make bandwidth of the critical stages wider, hence contributing less to overall bandwidth variations. Here, again, if this method is carried to extremes, more tuned transformers would be required in other stages when good skirt selectivity is required, thus affecting the cost of the design. Further, with reduced stage stability at high gain, the bandwidth and gain will be adversely affected by feedback variations.

RX TRANSFORMER DESIGN

In a medium-bandwidth transistor amplifier, the RX transformer can be used for interstage coupling. This transformer, shown in Fig. 3, consists of a simple tuned circuit, L_2 and C_2, with series damping resistor R_s providing an input mismatch to transistor T_2, a series damping for $L_2 - C_2$, and, hence, an output mismatch for T_1. From the basic transistor parameters (Table I) and from equation 10, it will be seen that $R_s \ll Z_{in}$; thus, for high circuit Q's,

$$I \approx \frac{V_c}{X_{c2}}, \quad X_{c2} = \text{reactance of } C_2 \gg R_s$$

$$\text{and } V_2 = IR_s \approx \frac{V_c}{X_{C2}} \times R_s \quad (19)$$

$$\text{hence, } N = \frac{V_c}{V_2} \approx \frac{1}{R_2 \omega_o C_2}$$

The name *RX transformer* is derived from a reactance-to-resistance voltage transfer.

$$\text{Also: } Q = \frac{1}{R_2 \omega_o C_2} = N \quad (20)$$

$$\text{Equivalent } R_L = Q^2 R_s = N^2 R_s \quad (21)$$

By comparing the above results with those of the transformer in Fig. 2, the

RX transformer is treated as a special case wherein $Q = N$ numerically, with other parameters being equivalent. Because of this equity, the use of the RX transformer is greatly limited, since the design flexibility is restricted. However, the advantage of the RX over the conventional transformer will be apparent by considering its performance in the presence of a capacitive feedback, represented as C_F in Fig. 4a. The amplitude and phase of the feedback voltage at the base of transistor T_2 are compared with input voltage V_2 in Figs. 4b and 4c. The dotted lines in both curves outline the feedback effects on the transformer

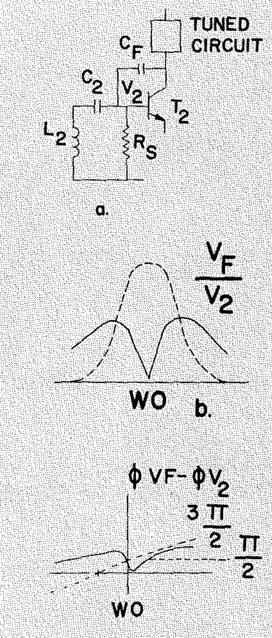


Fig. 4—Collector-to-base feedback in the RX transformer.

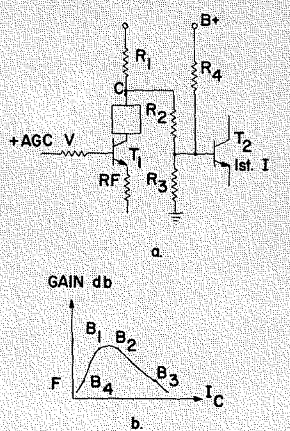


Fig. 5—AGC system for KCS-153.

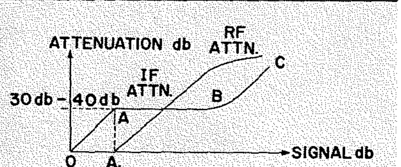


Fig. 6—Idealized AGC characteristic for TV receiver.

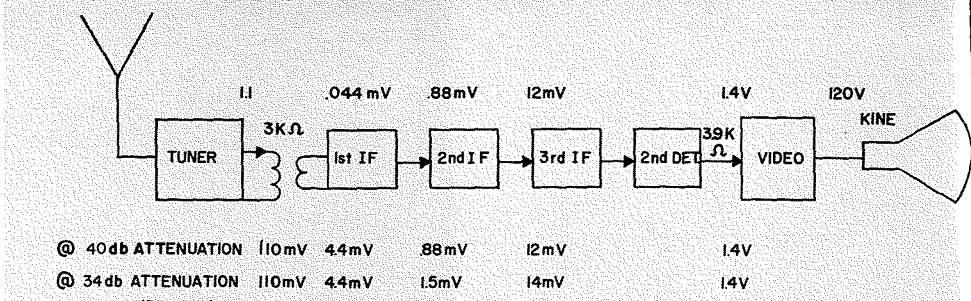


Fig. 7—Picture signal amplifier, block diagram.

shown in Fig. 2; from this feedback characteristic, a good degree of self-neutralizing can be expected from the RX transformer when in synchronous tuning condition. In stagger-tuned circuit applications, the stability is retained with a correct choice of stagger frequencies.

By again looking at Table I, with $R_i = 0.55$ kilohm, and $R_o = 40$ kilohms, and using example 1 plus equations 10 and 11, then N becomes,

$$N = \sqrt{\frac{\frac{40}{19.5}}{\frac{0.55}{12.5}}} = 6.8 \qquad (22)$$

Typical values of loaded Q's, between 5 and 10, are used in the picture television IF amplifier, depending on the design. Thus, in transistor applications, the RX transformer can be used easily, making available its advantages of simplicity and improved stability.

The design of the RX transformer follows procedures that are similar to those of any other transformer, except for this special case since $Q = N$, the stage gain (G) is

$$G = \frac{g_m}{\omega_o C} \qquad (23)$$

Thus tuning capacitance is first determined, and then the value R_s is selected for best compromise where Q equals N.

AGC CONSIDERATIONS

The AGC requirements for television receivers are described in other papers.[6] In general, the forward-bias method has been preferred because this method provides better linearity and slower slope characteristics at cutoff, as may be seen in Fig. 5b. A further argument in favor of forward bias, especially with older germanium transistors, is the difficulty of controlling I_{cbo} with temperature, especially near cutoff with reverse bias.

However, in recent silicon transistors I_{cbo} is kept very small; thus, when a satisfactory cross-modulation characteristic is available, the reverse bias can be used successfully. The advantages here are the higher efficiency provided for a given AGC control power available, and more consistent gain at cutoff. Further, with the latest transistors being designed for reverse-bias operation, adequate gain reduction may be consistently obtained with a single stage; hence, simplicity is enhanced and the cost of the design is reduced, or performance is improved for similar costs.

Thus, the reverse bias method cannot be rejected without good reason; in optimum design, both approaches should be considered and the best one applied to the overall system design.

Cross-modulation characteristics generally favor the use of a forward-biased AGC in tuner RF stages. By choosing filters with adequate selectivity to precede the first IF stage, the IF cross modulation can be eliminated; consequently, reverse-bias AGC would be preferred. One such reverse-bias AGC system is shown in Fig. 5. A typical gain-versus-collector-current curve of the transistor is shown in Fig. 5b. The RF tuner stage, at a maximum gain condition, is biased to point B_1 on the curve of Fig. 5b; this point is coincident also with the optimum RF noise figure. As signal increases, the forward-bias voltage is applied to the RF stage, moving the operating bias to point B_2. This has negligible effect on the gain of the RF stage; however, voltage at the collector (point C) drops due to the increased bias current, thus reducing the IF stage base bias. In this operation, IF gain reduction is first obtained. Since the amount of attenuation in the IF (T_2) is limited by current through resistor R_4, which is connected between the base of T_2 and the fixed B+ voltage point, further increasing the bias at T_1 provides attenuation in the RF stage only, as shown by condition $B_2 - B_3$ in Fig. 5b. The limiting resistor (R_4) can be replaced by more complex diode circuitry, resulting in a better clamping performance whenever required.

A typically ideal AGC requirement is shown in Fig. 6; with increasing signal, IF gain is reduced initially to improve the signal-to-noise ratio at the mixer and first IF. After the noise effect has been eliminated, most of the gain reduction should be obtainable in the RF stage to

Fig. 8—Stage responses for KCS-153: a) for input transformer; b) for remaining three transformers.

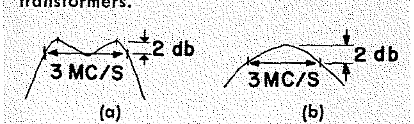
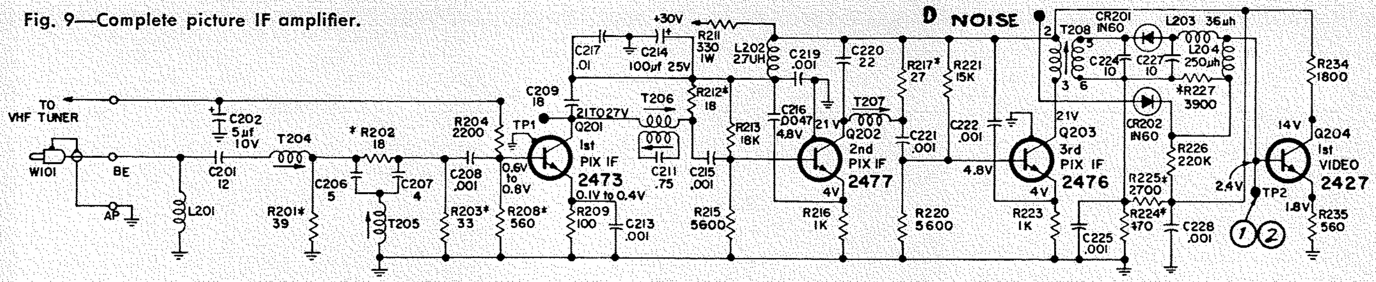


Fig. 9—Complete picture IF amplifier.



give the following stages adequate protection against cross modulation and beat effects. A small slope on curve *ABC* may often be required to provide additional gain reduction (points *B-C*), after the gain control limit of the RF device is reached.

KCS-153 IF DESIGN

The gain requirements for the picture IF amplifier and the total gain distribution that may be derived from following considerations are shown in Fig. 7; the gain quoted is the peak gain of overall IF response.

TABLE III — Total Gain Requirement

Basic overall gain of IF	63 dB
Transformation gain present from 3-kilohm input to 3.9 kilohms	—1.3 dB
Second detector efficiency 65%, loss	3.5 dB
Bridge T trap expected to be used, loss	3.0 dB
Total Gain	68.2 dB

From Table III, three stages at 23 dB gain are required; in the initial design, the RX transformers are used to obtain neutralized operation.

Considering bandwidth selectivity and phase response, five tuned transformers are acceptable in the design; this is based on standard practice in current tube design. To obtain maximum cross-modulation protection in the first IF stage, a double-tuned input and three single-tuned coupling transformers were chosen. The initial response of individual stages is shown in Fig. 8. The parameters realizing both responses (a and b) will follow normal practice in tube receiver design, with additional care being taken to correct impedance matching.

The only factor now remaining is to insure adequate cross-modulation performance. The device to be used is measured for 1% cross modulation, and bias conditions are carefully selected. Typical readings for unwanted signal level at the base of TA2562 in reverse bias for 1% cross modulation are given in Table IV.

TABLE IV — Typical Readings of Gain, Bias, and Signal to Produce 1% Cross Modulation.

Gain	*Bias*	*Signal*
Max.	4 mA	4.5 mV
—1 dB	2 mA	3.5 mV
—34 dB	—	3.5 mV
—40 dB	—	2.7 mV

The selectivity of the input coupling circuit is made sufficient for adjacent signal cross modulation at IF; for inband cross-modulation protection, Fig. 7 should be examined and the following steps considered:

1) With the first IF at maximum attenuation of 34 dB, the double-tuned input selectivity is made such that the sound carrier is at least $\frac{4.4}{3.5} = 1.25$; i.e., 2.5 dB below picture carrier.
2) Second IF—no problems.
3) Third IF—biased to 4 mA and no problems.

It will be seen that in the stages following the first IF, an accompanying sound rejection increases to insure proper second-detector operation, thus improving the cross-modulation safety factor.

CONCLUSION

The complete schematic of the picture IF amplifier is shown in Fig. 9; selectivity and bandwidth response are shown in the graph of Fig. 10. The field performance and cost of the amplifier are comparable to current tube receivers. The added advantages of low-voltage operation and small size promise a wider use of transistors in future TV receivers.

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2. D. D. Holmes and T. O. Stanley, "Stability Consideration in Transistor Intermediate Frequency Amplifiers," *Transistors I*, p. 403-421.
3. D. D. Holmes and T. O. Stanley, *Transistors I*, p. 407.
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5. "Transistor Circuit Design" by Texas Instruments Inc.
6. *Funk Technic 1964*, Nr. 15, p. 538-540.

APPENDIX 1

In current TV design practice, the spread of tube capacitances is approximately ± 20%. Damping resistors of 5% tolerance are used to control operating circuit bandwidth; thus, total possible bandwidth variation is ±25%.

Since the GB product is high in transistor design, bandwidth will be controlled mainly by external tuning capacitors with ±5% tolerance; resistors with a 5% tolerance will also be used. For an absolute maximum bandwidth spread of ±25%, ±15% may be allowed for the spread of input and output resistance of the transistors used.

APPENDIX 2

The parameter spread for transistor input is considered below. By definition:

$$R_i = \frac{R_{i\,max.} + R_{i\,min.}}{2}, \text{ and } \frac{R_{i\,max.}}{R_{i\,min.}} = n$$

$$\text{Also } R_1 = \frac{R_s \times R_i}{R_s + R_i}$$

For fractional bandwidth spread of $\pm\delta_\Delta$, because circuit Q's are proportional to R_1, and considering the admittances in the limit case, the equations become

$$\left(\frac{1}{R_s} + \frac{1}{R_{i\,min.}}\right) - \left(\frac{1}{R_s} + \frac{1}{R_{i\,max.}}\right) = \frac{\delta_\Delta}{R_1}$$

$$\frac{\delta_\Delta}{R_1} = \frac{1}{R_{i\,min.}} - \frac{1}{R_{i\,max.}}$$

$$\frac{\delta_\Delta}{R_1} = \left(\frac{1}{\frac{2R_i}{1+n}}\right) - \left(\frac{1}{\frac{2R_i}{1+\frac{1}{n}}}\right)$$

$$\frac{\delta_\Delta}{R_1} = \frac{1 + n - 1 - \frac{1}{n}}{2R_i}$$

$$\frac{2R_i}{R_1} = \frac{n - \frac{1}{n}}{\delta_\Delta}$$

Fig. 10—Overall picture channel amplitude response.

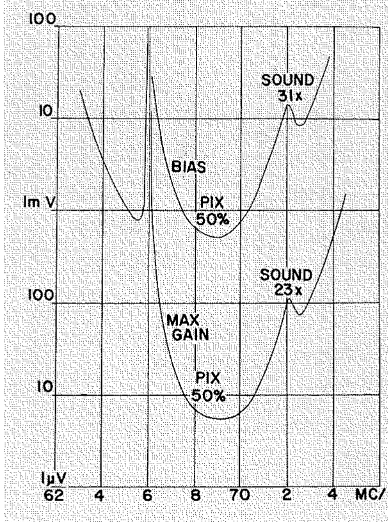
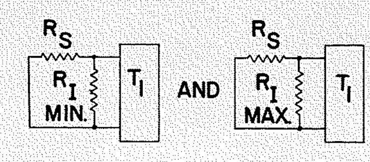


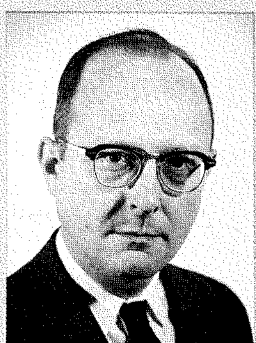
Fig. 11—Transistor input configurations for R_i min. and R_i max.



The David Sarnoff

The 1966 Individual Awards for
Science and Engineering


Dr. C. W. Mueller


S. H. Buder

CHARLES W. MUELLER, Fellow of the Technical Staff, RCA Laboratories, Princeton, N.J., recipient of the 1966 David Sarnoff Outstanding Achievement Award in Science *"for outstanding contributions to the technology of semiconductor devices and circuits."*

DR. MUELLER, winner of the individual science award, has been one of the pioneers in semiconductor technology. On the staff of RCA Laboratories since 1942, much of his early work was with microwave receiving tubes, secondary electron emission, and semiconductor devices. In the field of junction transistors, he was responsible for development of the alloy-junction technique used in many commercial types and for the first alloy-junction transistor specifically designed for higher frequency operation. He developed the thyristor switching transistor and supervised the work leading to a parametric diode with a cutoff frequency above 200 GHz and developed the low-inductance ceramic package required for such high frequencies. He also participated in the development of very-high-speed tunnel diodes. Recently he has led the development of a technique for growing single-crystal silicon films on sapphire and the use of these films in integrated circuits.

SPURGEON H. BUDER, Missile and Surface Radar Division, RCA Defense Electronic Products, Moorestown, N.J., recipient of the 1966 David Sarnoff Outstanding Achievement Award in Engineering *"for development and application of novel techniques of system analysis and synthesis to advanced defensive systems."*

MR. BUDER, recipient of the individual engineering award, has gained considerable recognition for his technical contributions to and his comprehensive knowledge of overall air defense systems. He has been with RCA since 1941. He was lead systems engineer in development of the Navy Mk 2 target-designation equipment, one of the first advanced fire-control systems for missiles. He also made significant contributions to RCA's Range Instrumentation Program, in development of novel radar applications, designation systems, and data-processing techniques. In 1960 he was RCA's key technical representative at Vandenberg Air Force Base for the integration of the control and checkout subsystems of the Atlas missile with the overall missile system. In 1963 he was instrumental in RCA's successful analysis of the Terrier missile system for improved performance. His work during 1964 and 1965 in the field of system analysis and synthesis of tactical air defense systems has contributed heavily to the effectiveness of this nation's future-generation air defense systems.

. . . About the Awards

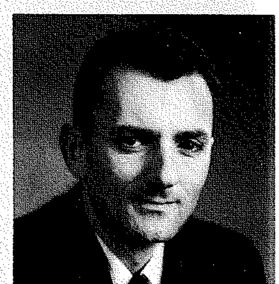
RCA's highest technical honors, the four annual David Sarnoff Outstanding Achievement Awards, have been announced for 1966 by Dr. George H. Brown, Vice President, Research and Engineering. David Sarnoff, RCA Board Chairman, and Dr. Elmer W. Engstrom, Chairman of the Executive Committee and Chief Executive Officer, will present the awards, which consist of a gold medal, a bronze replica citation, and a cash prize for each.

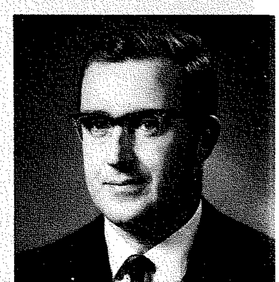
The Awards for individual accomplishment in science and in engineering were established in 1956 to commemorate the fiftieth anniversary in radio, television, and electronics of David Sarnoff, RCA Board Chairman. The two awards for team performance were initiated in 1961. All engineering activities of RCA divisions and subsidiary companies are eligible for the Engineering Awards; the Chief Engineers in each location present nominations annually. Members of both the RCA engineering and research staffs are eligible for the Science Awards. Final selections are made by a committee of RCA executives, of which the Vice President, Research and Engineering, serves as Chairman.


Dr. S. Larach


Dr. R. E. Shrader

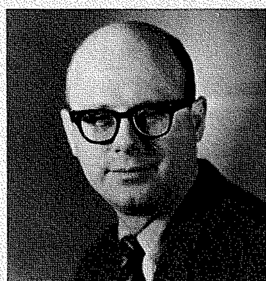
The 196


R. S. Hartz


F. P. Jones

...utstanding Achievement Awards

Dr. S. M. Thomsen



Dr. P. N. Yocom

Dr. Simon Larach, Dr. Ross E. Shrader, Dr. S. Milton Thomsen, and **Dr. P. Niel Yocom,** Materials Research Laboratory, RCA Laboratories, Princeton, N.J., recipients of the 1966 David Sarnoff Outstanding Team Award in Science *"for team performance in making substantial advances in the science, art, and practical application of luminescent and photoconductive materials."*

Drs. Larach, Shrader, Thomsen, and **Yocom,** winners of the team award in science, have made important individual and collective contributions in developing special luminescent and photoconductive materials for use in manufacturing photocells and kinescopes. Dr. Shrader, a physicist, in order to evaluate the thousands of experimental photoconductors and phosphors synthesized by his three fellow chemists, designed original spectroradiometric equipment. Dr. Thomsen, in his research on phosphors and photoconductors, developed a technique for making practical sintered cadmium sulfide photocells, a process now used to manufacture such cells at RCA's Mountaintop plant. Dr. Larach made personal contributions both to the development of cathodoluminescent phosphors for kinescopes and electroluminescent phosphors for solid-state information displays. In all these studies, Dr. Larach coordinated the laboratory work with that of other technical personnel in Lancaster and Mountaintop. Dr. Yocom played an important part in the development of a new red-emitting phosphor for color kinescopes and an intense blue-emitting phosphor. He has also been creative in conceiving and developing new luminescent materials for masers and lasers.

...eam Awards for Science and Engineering

Messrs. Ralph S. Hartz, F. Peter Jones, Henry M. Kleinman, and **Joel Ollendorf,** Commercial Receiving Tube & Semiconductor Division, RCA Electronic Components & Devices, Somerville, N.J., recipients of the 1966 David Sarnoff Outstanding Team Award in Engineering *"for team performance in developing the first line-operated audio output transistor for the economical transistorization of home radio and television receivers."*

H. M. Kleinman

J. Ollendorf

Messrs. Hartz, Jones, Kleinman, and **Ollendorf,** winners of the team award in engineering, have successfully developed a high-voltage off-the-line transistor for the practical transistorization of television deflection systems and video output circuits. In January 1963, Messrs. Kleinman and Hartz set up complete objective specifications for the new transistor; Messrs. Ollendorf and Jones began design of a device and processes to meet these specifications. An initial problem was development of a high-resistivity crystal for fabrication of a prototype device. Further studies indicated that achieving and maintaining high voltages was just as dependent upon proper surface preparation as on use of a high-resistivity crystal. A prototype device, the TA2301, was introduced in January 1964. As a result of Messrs. Kleinman and Hartz' successful development of circuitry for reliable operation under normal or abnormal operating conditions, the device was announced commercially in November 1964. Messrs. Jones and Ollendorf, as a last phase of the project, assisted in establishing proper manufacturing processes for the new high-voltage silicon transistor. In 1965 the new device, known commercially as the 40264, was selected by *Industrial Research* magazine as one of the 100 most significant new technical products of the year.

A TRANSISTORIZED HORIZONTAL DEFLECTION CIRCUIT FOR TELEVISION

This paper describes a transistorized horizontal system developed for use in TV receivers. The system consists of an output stage, a driver stage, an oscillator, and AFC circuitry. The output stage uses two power transistors, one as a switch and the other as a protective device. The driver stage is a one-transistor power-gain stage with special wave-shaping circuitry. A transistorized blocking oscillator is used with refinements added for stability purposes. The AFC circuit is a conventional phase splitter with a diode-gate phase detector.

L. G. SMITHWICK

TV Product Development

RCA Victor Home Instruments Div., Indianapolis, Ind.

THE primary function of the horizontal deflection circuit in a television receiver is to drive a sawtooth current through a yoke to provide electromagnetic deflection of the electron beam in the kinescope. The deflection system also performs these secondary functions: 1) develops the second-anode high voltage, 2) accelerates the anode voltage, and 3) produces blanking pulses for the kinescope. Deflection is performed at a rate of 15.75 kHz (kc/s) to provide a 525-line raster.

Although many of the concepts and principles of the circuit discussed herein are common to both vacuum-tube and transistor circuits, special consideration is given to the unique characteristics of the transistor.

OUTPUT STAGE

Horizontal deflection of the electron beam in a television receiver is accomplished by switching techniques instead of the large-signal amplification used to drive the vertical yoke. Unlike the vertical yoke, which is treated as a resistive load at the vertical sweep rate, the horizontal yoke must be considered a purely inductive load at 15.75 kHz.

Circuit Operation

The output stage of the KCS-153 horizontal deflection circuit is illustrated in Fig. 1. The driver switches transistor Q_s from an *off* condition to an *on*, or saturated, state at the appropriate times. For simplicity, this transistor will be considered an ideal switch. In this circuit $C_s \gg C_t$ and $L_f \gg L_y$.

Assume that initially the following conditions exist:

1) Capacitor C_s is positively charged.
2) Transistor Q_s is saturated.
3) Current is flowing in L_y and the flyback transformer, as shown in Fig. 1.

Then, at time t_0 (Fig. 2) transistor Q_s is switched *off*, but the currents present in the yoke and flyback transformer cannot cease flowing immediately. The only available path for current flow is through C_t; since these currents drop rapidly, C_t is charged to a high voltage (about 250 volts for the KCS-153). The value of this voltage is, of course, determined by the magnitude of currents I_f and I_y and the capacitance of C_t (represented in Fig. 2 from t_0 to t_1).

Now that capacitor C_t is charged to 250 volts and I_y has dropped to zero, this capacitor will begin to discharge through L_y, creating a yoke current in the opposite direction (t_1 to t_2, Fig. 2). Note that Q_s is still in the *off* state.

Perhaps a more appropriate way of looking at the circuit operation from t_0 to t_2 would be to note that C_t and L_y form a tuned circuit having a resonant frequency with a period equal to twice the time from t_0 to t_2, or about 20 μsec.

At time t_2 the voltage across C_t has dropped to zero and would ring negative except for the presence of diode D_D. The large current present in L_y opens this diode at t_2 and current flows through D_D and L_y in the form of a ramp (Fig. 2) until time t_4.

Transistor Q_s is switched to *on* at t_3 (Fig. 2); however, this has no bearing on the conduction of D_D until time t_4, when the diode becomes back-biased. At t_4, the transistor begins to conduct yoke current in the forward direction and continues to do so until time t_5, when it is switched *off* and the cycle begins anew.

Up to this point little has been said about the function of capacitor C_s. This capacitor is necessary to block DC current flow in the yoke and prevent a degree of static deflection which would place the raster off center. A further result is that without C_s the AC current in the yoke would follow the rule,

$$di = \frac{Edt}{L}$$

and yoke current would be a linear ramp function (neglecting yoke resistance). A linear current ramp is not desirable because the curvature of the face of the CRT has a radius greater than the distance from the electron gun to this face, and S shaping must be applied to the yoke current to obtain a linear visual sweep. This S shaping is shown in Fig. 2 (t_2 to t_5). The voltage across the capacitor resembles a rectified sine wave, and this waveform forces the yoke current to vary in the fashion shown in Fig. 2.

It would appear (Figs. 1 and 2) that the voltage at the emitter of transistor Q_s should be the positive portion of a sine wave whose frequency is determined by L_y and C_t, since $L_f \gg L_y$ and $C_s \gg C_t$. Such is not the case, and the actual waveform is a combination of two sine waves (Fig. 3). This waveshaping is referred to as third-harmonic tuning, and the method used to obtain it is described in a later section of this paper.

This tuning, while not essential, greatly improves circuit operation and can be explained in the following manner. Since the flyback transformer will always have some leakage inductance, the circuit will always exhibit some high-frequency ringing. If this frequency is controlled, it can be used to improve circuit operation and to decrease losses. As shown in Fig. 3, it is used to reduce the peak pulse on the output transistor. Further, since the phase of the third harmonic is reversed on the leakage inductance of the flyback transformer, it adds to the peak pulse at the high-voltage rectifier, thereby increasing the second-anode voltage.

In this circuit it is necessary to protect transistor Q_s when fault conditions occur, particularly high-voltage arcing. If arcing occurs in the high-voltage rectifier or the kinescope, much more energy is required of the flyback primary; this means that a considerably higher current must flow through Q_s when the transistor is turned *on*. Since the base current in Q_s is constant during its *on* period, the transistor is forced out of saturation and a voltage develops between collector and emitter. This in-

Final manuscript received January 12, 1966.

LUKE G. SMITHWICK received his BEE degree from the Georgia Institute of Technology, Atlanta, Georgia in 1964. Upon graduation he was employed by RCA Home Instruments Division where he is currently working with transistorized deflection systems in the Black and White Product Design Group. He is a member of Tau Beta Pi and Eta Kappa Nu.

creased V_{CE} and the high collector current result in very high peak power in the device (typically 400 watts or more) and it fails.

To protect transistor Q_s, a current-limiting circuit is added to the output stage (Fig. 4). The base of Q_L in the limiter is tied to ground through a resistor, and since this base is always essentially at B+ voltage, a constant current flows through the base-emitter junction. This establishes a maximum, but not a minimum, collector current.

Now, obviously the maximum current through Q_L is determined not only by the value of the base resistor, but by the β of the transistor as well. Thus, a close tolerance must be maintained on this parameter to insure uniform current limiting in each receiver.

Another failure condition exists when the high-voltage rectifier arcs and places a very high (possibly 400 volts) reverse voltage across limiting transistor Q_L. Since such a voltage would destroy Q_L, diode D_L is provided as a protective device to prevent the appearance of a reverse voltage across this transistor.

In reactive circuits where large currents are switched at high speeds, it is difficult to avoid radiation caused by transients. In a horizontal circuit, radiation is suppressed by small feedthrough capacitors to keep it from appearing in the video information.

The fast switching of high currents further requires that AC isolation be maintained between the circuit and the power supply to prevent the propagation of horizontal information to other circuits; isolation is provided by a simple *L-C* filter (Fig. 4).

Design Considerations

The equivalent circuit of the output stage (Fig. 5) is a convenient tool for examining the design methods used for this circuit. In the equivalent circuit the battery and switch represent the supply voltage and the active element (transistor in this case) used for switching. L_1 represents the inductance of the yoke and the flyback transformer in parallel. But since the flyback transformer inductance is an order of magnitude greater than the yoke inductance, it is reasonably accurate to assign L_1 the value of the yoke inductance (200 μH for the KCS-153).

Tuning capacitor C_t in parallel with L_1 is chosen for resonance at the proper flyback frequency. L_2 represents the leakage inductance of the flyback transformer referred to the primary.

Capacitor C_3 represents stray capacitance between windings of the flyback transformer referred to the primary. Capacitor C_2 represents the capacitance of the high-voltage rectifier plus the cap lead and the capacitance between the transformer and the high-voltage box. Again, these are referred to the primary.

A Practical Design Example

In designing a horizontal output circuit, practical as well as theoretical approaches are used, as shown in the following general example. These calculations are not intended to provide accurate design information, but are a rough approximation for initial design.

1) *Determination of Required Deflection Energy*

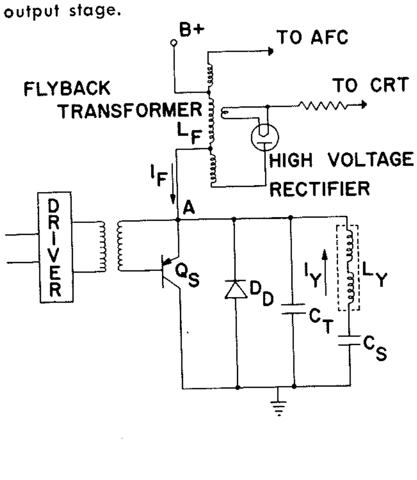
Although it is possible to estimate theoretically the energy required for beam deflection, a simpler approach is more practical. Given the high-voltage requirement, the proper kinescope, and a suitable yoke, the DC current required for deflection from the center of the tube to the side is measured. Energy is then calculated from the following equation:

$$E = \tfrac{1}{2} I^2 L, \text{ where } I = \text{DC current required for full deflection.}$$

2) *Determination of Maximum Flyback Voltage*

Given a transistor to act as a switch which has an established V_{CEX}, the maximum flyback voltage is determined as follows:

Fig. 1—Simplified schematic of the KCS-153 output stage.



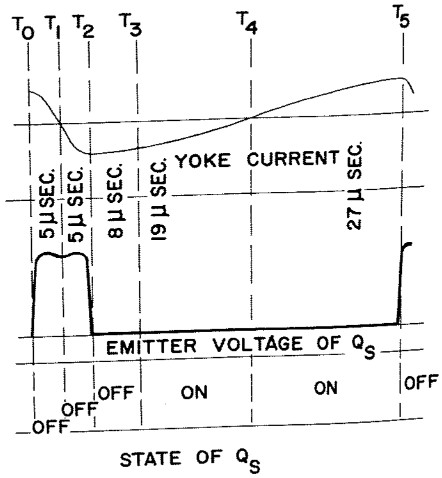


Fig. 2—Output-stage currents and voltages.

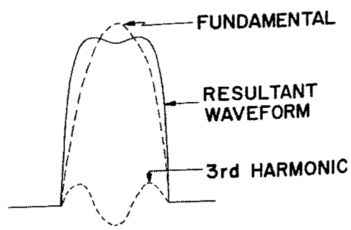


Fig. 3—Third-harmonic tuning technique.

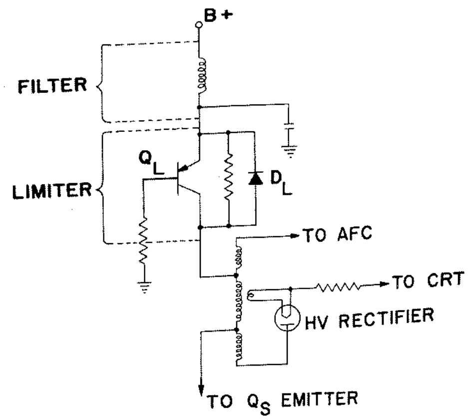


Fig. 4—Current-limiting circuit for output transistor protection.

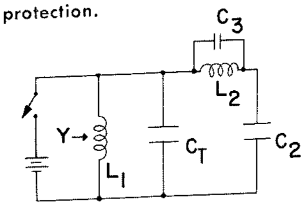


Fig. 5—Equivalent circuit for horizontal output stage.

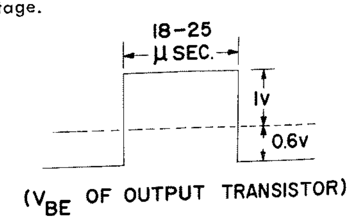


Fig. 6—Simplified base waveform for output transistor.

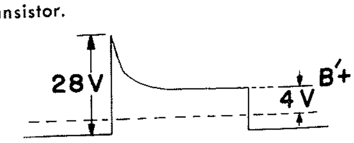


Fig. 7—Waveshaping of driver output.

$V_p = 0.75\ V_{CEX}$, where the 0.75 is simply a safety factor.

3) *Supply Voltage*

Once the flyback voltage, V_p, is determined, it is possible to specify the supply voltage required.

Supply voltage $V_b = \frac{1}{8}\ V_p$ (approximate relationship)

4) *Required Yoke Inductance*

Given the energy for full deflection (step 1), the inductance of the yoke can be established from the following equation.

$$\hat{E} = \frac{1}{2} I^2 L$$

$$\hat{E} = \frac{1}{2} \left(\frac{V_b t}{L}\right)^2 L$$

$$L = \frac{V_b^2 t^2}{2\hat{E}} \quad \text{where } t = 26\ \mu\text{sec}$$

5) *Flyback Tuning Capacitor*

Assuming a flyback pulse width of 10 μsec, the sine wave required to produce such a pulse would have a period of 20 μsec, corresponding to a frequency of 50 kHz. The C_t in Fig. 5 is chosen for resonance with the yoke at this frequency.

$$\text{then: } C_t = \frac{1}{\omega^2 L_1} \cong 10^{-11}/L_1$$

$$\text{or: } L_1 C_t \cong 10^{-11}$$

6) *Flyback Transformer Turns Ratio*

The flyback turns ratio can be determined from the following equation, where KV = high voltage, and V_p = flyback pulse peak:

$$N \cong \frac{KV}{1.4\ V_p}$$

7) *Evaluation of* C$_2$

For the average receiver, the values of rectifier capacitance, cap-lead capacitance, and capacitance of flyback transformer to ground total about 9 pF as a sum. Then C_2 is determined as follows:

$$C_2 = (9 \times 10^{-12})\ (N^2)$$

8) *Determination of* C$_3$

The value of winding capacitance is measured on the actual transformer.

Then,

$$C_3 = (\text{measured capacitance})\ (N_2)$$

9) *Calculation of* L$_2$

To calculate the required leakage inductance, it is necessary to write the admittance equation for the equivalent circuit of Fig. 5. Solution of this equation leads to a numerical value for L_2.

DRIVER STAGE

When design of the output stage is completed, its drive requirements are already established.

First, the driver must be capable of delivering sufficient current to maintain a minimum β output transistor in saturation at peak output current.

In the case of the KCS-153, the specified minimum β of the output unit at maximum collector current (6 amperes) is 15; therefore, it appears that a minimum of 400 mA is required for proper drive. However, if 400 mA driving current is used, high β output transistors exhibit a longer cutoff time because of storage in the base-emitter and base-collector junctions. Therefore a compromise was made, and the drive current was set at 360 mA. While this current allowed lower β limit transistors to travel slightly out of saturation, the associated dissipation and scan nonlinearity was not intolerable and good cutoff time was maintained.

The driving waveform on the base of the output transistor must drive this transistor *off* for 18 to 25 μsec. This period, covering the retrace time and part of the damper diode conduction time, is:

$$t_R < T_o < (t_R + t_D)$$

where t_R = retrace time, T_o = *off* time, t_D = diode drive time.

$$t_R \cong 10\ \mu\text{sec.}$$

$$t_D \cong 25\ \mu\text{sec.}$$

The next consideration is the driver transformer turns ratio. Since the voltage and current in the secondary are dictated by the output transistor, the turns ratio must be a compromise that will insure that neither the voltage nor the current rating of the driver transistor is exceeded. A 4:1 turns ratio was used in the KCS-153.

Given the above considerations, the waveform shown in Fig. 6 normally would be used for switching; however, this waveform does not provide an adequately fast cutoff and must be altered. The voltage waveform of Fig. 7, when developed on the *primary* of the driver transformer, will provide the proper fast cutoff time.

This waveform is produced by the waveshaping circuit shown in Fig. 8. A description of circuit operation follows. Before driver turn-on, $C583$ is charged to almost full B+ voltage (about 28 volts), and in the series combination of driver transformer primary and driver transistor (these being in parallel with $C583$) all of this voltage is developed acoss the transistor. Then, as driver turn-on occurs, 28 volts is immediately developed across the transformer primary because the transistor is saturated. This provides the fast turn-off voltage required. As $C583$ discharges, the waveform exhibits a rapid fall time and stabilizes at a new voltage (say B′+) provided for by $CR507$, $R591$, and $C580$, a voltage clamp. This condition prevails until driver turn-off occurs.

Fig. 8—Complete driver stage.

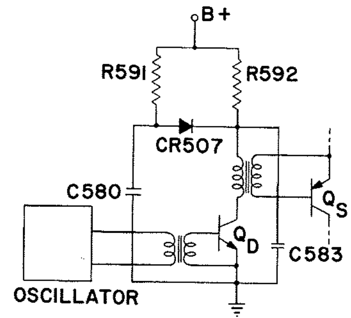


Fig. 9—Base-waveshaping circuit of driver.

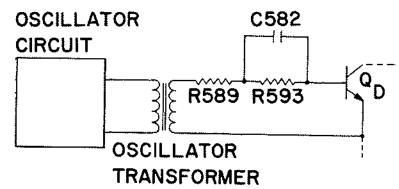


Fig. 10—State of driver-base circuit at oscillator turn-off.

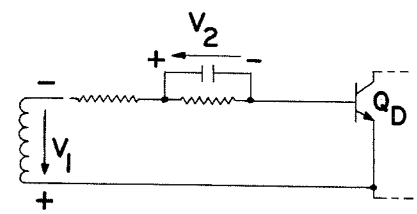


Fig. 11—Final driver output waveform.

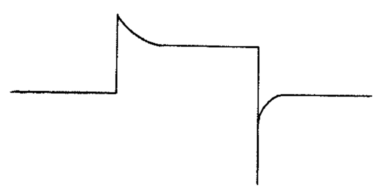


Fig. 12—Basic transistor blocking oscillator.

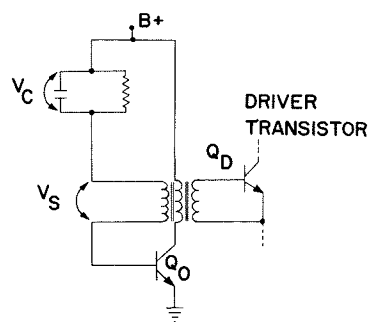


Fig. 13—Base voltage associated with basic blocking oscillator.

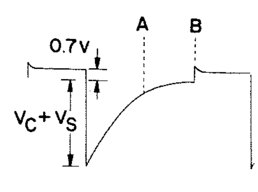


Fig. 14—Reshaped base voltage of oscillator transistor.

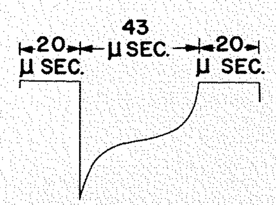


Fig. 15—Base off-time waveshaping network.

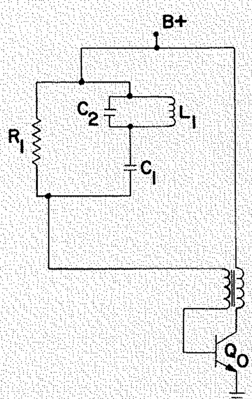


Fig. 16—Base-voltage waveshaping technique.

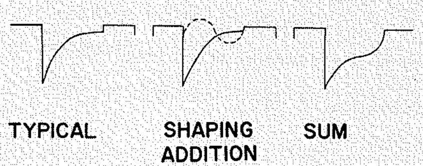


Fig. 17—Shaped-oscillator base current.

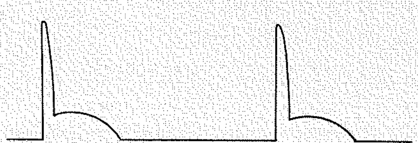
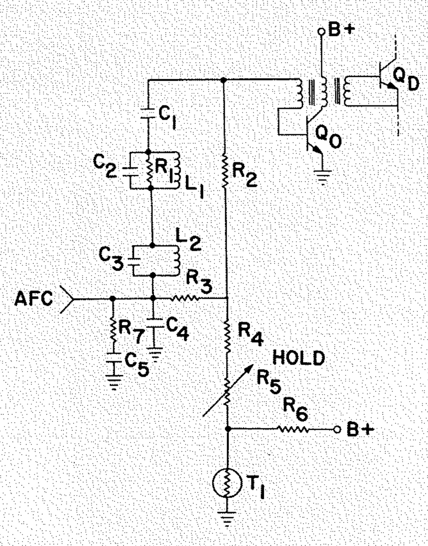


Fig. 18—Complete transistor oscillator.



Further consideration must be given to the base circuit of the driver transistor (Fig. 9). When a rectangular voltage is developed across the primary of the oscillator transformer, *C*582 acts as a short circuit at turn-on and the base current into the driver transistor is limited by *R*589. The current into the base then falls off at a rate prescribed by (*C*582 × *R*593). As the turn-off portion of the oscillator waveform occurs, the circuit is in the state shown in Fig. 10, and the resultant turn-off voltage has a sharp drop as indicated in Fig. 11. This voltage waveform on the base of the driver transistor provides both fast turn-on and fast turn-off.

HORIZONTAL OSCILLATOR

In the KCS-153 the rectangular waveform required for the driver transistor input is provided by a blocking oscillator, chosen for its ability to produce a low-impedance output waveform that is somewhat rectangular. Although the basic transistorized oscillator (Fig. 12) is conventional, it has several limitations and must be altered to make it sufficiently stable.

Several problems were solved to insure that the frequency of the oscillator would remain constant not only on a one-cycle basis but during long-term operation and temperature changes.

The basic oscillator has an unstable *off* time due to the shape of the base voltage waveform (Fig. 13). With this waveform, firing or turn-on should occure at point B; however, noise or circuit variations can cause a slight rise in the curve before point B and cause firing anywhere between A and B.

To eliminate this problem, waveshaping was employed to obtain the waveform shown in Fig. 14. Note that any slight change in amplitude in the waveform between A and B will not cause premature firing, and turn-on will occur only at point B. The circuit alteration shown in Fig. 15 provides this new waveshape. Note the addition of C_2 and L_1. The resonant-frequency period of L_1 and C_2 is related to the *off* time of the oscillator and adds the dotted waveform (Fig. 16) to the original waveform to provide the final base voltage shape in the *off*-time portion.

The *on* time is affected by the β of the transistor and the primary inductance of the transformer. The basic oscillator (Fig. 12) and the improved version (Fig. 15) are both dependent upon β as a factor in determining *on* time. Since cutoff occurs when I_C/I_B approaches β, cutoff could be controlled quite accurately if I_B could be made to approach zero at some predetermined time. With this ideal type of base current (Fig. 17), I_C/I_B will approach β at about the same time without regard to the value of this parameter.

This waveform can be achieved by further altering the base circuit as shown in Fig. 18; this circuit represents the one actually used in the KCS-153. The fixed tuned circuit composed of C_1 and L_1 comprises a series resonant circuit which changes the *on*-time base current to the waveform shown in Fig. 17.

AUTOMATIC FREQUENCY CONTROL

The incoming video information contains horizontal sync pulses which are separated from the video to regulate an automatic frequency control circuit. Such a circuit controls the frequency of the horizontal oscillator so that each horizontal scan line begins at the exact time the corresponding video is placed on the kinescope.

In the KCS-153 the flyback transformer contains a winding which feeds a pulse back to the AFC circuitry, where it is integrated to form a sawtooth. This sawtooth is fed to a diode gate which is opened for about 3 μsec during retrace time when the diodes are forward biased by a sync pulse. This sampling of the sawtooth waveform produces a pulse output proportional in amplitude to the degree of phase error. These pulses are passed through a low-pass filter having a DC output voltage whose magnitude varies with oscillator frequency. This DC voltage varies the bias on the oscillator, locking it to the proper frequency.

It may be asked why it is preferable to use AFC instead of the incoming sync pulses to trigger an oscillator. If noise should enter the system, and it invariably will, it would cause false triggering for such an oscillator. On the other hand, although a noise pulse will open the AFC gate and allow false information to be admitted, all information, both correct and false, is passed through the AFC filter. Consequently, the incorrect sampling is averaged with all the correct information, and the resultant error is greatly reduced, making the set much more immune to pulse noise.

ACKNOWLEDGEMENTS

The basic product design of this system was by Chi-Sheng Liu, who is presently studying for his Doctor's degree through a David Sarnoff Fellowship. The author worked with Mr. Liu in the final phases of the design. Contributions to this system were also made by L. R. Kirkwood, J. M. C. Tucker, J. H. Nevin, and T. Burris.

A TRANSISTORIZED TV VERTICAL DEFLECTION CIRCUIT

A TV vertical deflection circuit using four transistors arranged as a multivibrator and amplifier is described herein. The amplifier is connected as a Miller integrator in which feedback makes circuit operation more independent of transistor parameters. The circuit incorporates features that improve the thermal stability and linearity of the raster and the frequency stability of the multivibrator. The combination of multivibrator and Miller integrator causes system oscillations at 30 Hz; removal of these oscillations requires modification of the Miller feedback loop. Technical advantages of the final design are discussed.

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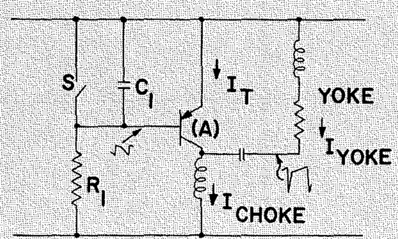

Fig. 1—Simple vertical deflection circuit.

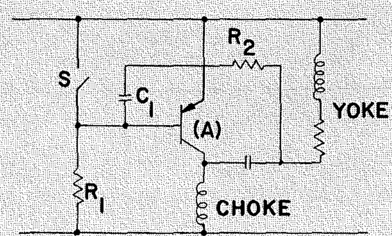

Fig. 2—Basic Miller amplifier deflection circuit.

THE purpose of the vertical deflection circuit is to drive a 60-Hz (c/s) sawtooth current through the vertical yoke windings. In most transistor circuits this is done using a voltage sawtooth generator consisting of C_1, R_1, and switch S (Fig. 1). Switch S is open during trace and closed during retrace; the resultant voltage is amplified by the transistor to provide the required current in the yoke.

The circuit described in this paper was designed for a 12-inch kinescope (12 BNP4) having a 1⅛-inch neck diameter, a 110° deflection angle, and a second anode voltage of 13.5 kV. The deflection current required for this combination is 500 mA peak to peak on a 47-ohm saddle yoke. In a 12-inch portable receiver, such scan requirements can cause the germanium output transistor to become quite hot, causing leakage current and thermal runaway.

OPEN-LOOP CIRCUIT

The open-loop circuit shown in Fig. 1 contains nothing to reduce the effects of component or device changes; for example, the electrolytic capacitor, C_1, can change value causing an alteration in the raster height. Also, it is difficult to compensate for output devices with limit leakage currents. To reduce these difficulties and to accommodate a wide range of transistor parameters, the decision was made to use a feedback circuit. The Miller integrator seemed to be the most suitable because it can easily be adapted to produce a linear sawtooth output.

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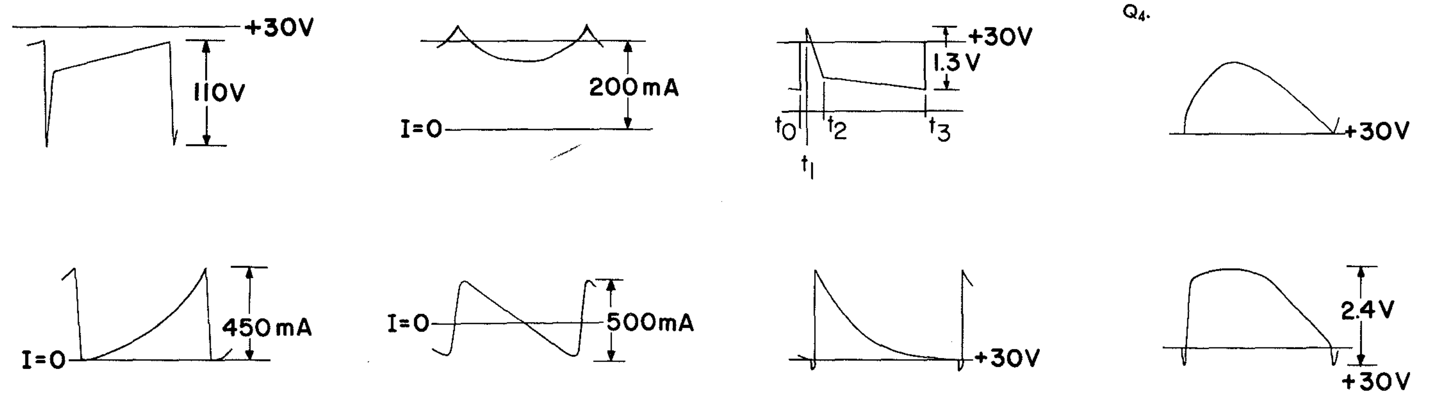
Fig. 3—Voltage and current waveforms.

(a)—Yoke voltage. (c)—Choke or B+ current. (e)—Voltage at the base of transistor Q_2. (g)—Waveshape derived from the collector waveshape of transistor Q_4.



(b)—Output transistor collector current. (d)—Yoke current. (f)—Typical waveshape at the base of a vertical switch transistor. (h)—Waveshape at the base of transistor Q_1.

THE MILLER CIRCUIT

The basic Miller circuit is shown in Fig. 2. During trace time, switch S is open and amplifier A has a high gain. The sawtooth or trace portion of the output voltage waveshape has an amplitude of 30 volts; thus the voltage waveshape at the input of the amplifier has an amplitude of 30 ÷ gain. When the input is 20 volts and the gain is sufficient, the voltage across R_1 is almost constant, giving a constant current through R_1. If the input impedance of amplifier A is high enough, all of this current will flow through capacitor C_1, resulting in a linear change of voltage on C_1; hence, the voltage across the yoke is a linear ramp.

During retrace, switch S is closed and the amplifier is, therefore, biased *off*. The stored energy in the yoke inductance creates a negative peak pulse which recharges C_1 through switch S and R_2. Resistor R_2 must be chosen so that the charge received during retrace is sufficient to restore C_1 to the correct initial condition at the start of trace.

This simple circuit imposes a voltage across the yoke as shown in Fig. 3a. The yoke impedance is 47 ohms in series with 53 mH; at 60 Hz the inductance has little effect during trace, so that this voltage waveform results in a linear sawooth current.

PRINCIPLE OF OPERATION

In the complete circuit (Fig. 4) the principle of operation is the same as in the circuit of Fig. 2. Transistors Q_2, Q_3, and Q_4 replace amplifier A. Transistors Q_2 and Q_3 are emitter followers and provide additional current gain. Switch S is replaced by transistor Q_1. There is a type of multivibrator connection between Q_1 and the amplifier stages Q_2, Q_3, and Q_4. At the end of trace, Q_1 starts to conduct; this tends to cut off transistors Q_2, Q_3, and Q_4, thereby reducing the current in the yoke. This current change through the yoke inductance causes a negative voltage spike which is fed through circuitry H (Fig. 4) to the base of Q_1, increasing the conductance of the transistor. This regenerative feedback causes Q_4 to be biased *off* and Q_1 to be pulsed into saturation for the duration of retrace. The charge in capacitor C_4 holds Q_1 *off* during trace, as in a multivibrator.

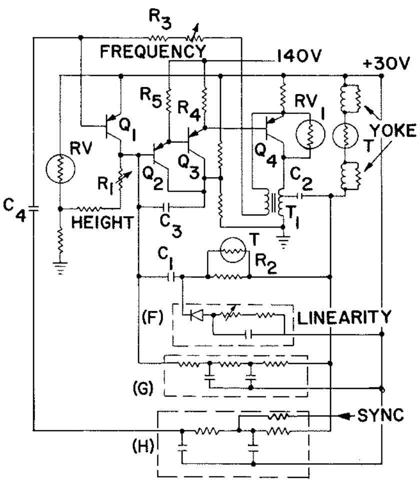
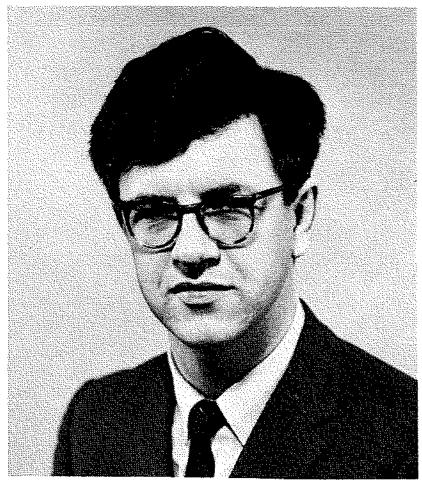


Fig. 4—Schematic diagram of complete TV vertical deflection circuit.



JAMES A. McDONALD received his B.Sc. (honors) from the University of Glasgow in 1955. He is currently working for an MSEE at Purdue University. From 1955 to 1962 he worked in the Electronics Division of the Associated Electrical Industries Ltd, Leicester, England. His assignments included the design of IF amplifiers, AFC systems, and sea clutter circuits for marine radar. He also worked on AFC systems for classified radar-missile programs. Mr. McDonald joined the RCA Victor Home Instruments Division in 1962. He is presently working on television deflection circuits.

The Miller capacitor is C_1 and R_2 is the resistor which affects the charge received by C_1 during retrace. Network F and network G (Fig. 4), which affect stability and linearity, are discussed later. The frequency is synchronized with a pulse to the base of Q_1, via some integration in network H.

OUTPUT STAGE

The design of the output stage is almost independent of the rest of the circuit. In the production of horizontal output transistors some units fail to meet the specifications, but they satisfy the less stringent requirements of the vertical output. For economy these units are used and they determine the type of vertical output transistor (similar to the 2N3730). Since the peak pulse on the horizontal output is a fixed multiple of the B+ voltage, the voltage rating of the horizontal output transistor determines the maximum B+. In the KCS-153 chassis the B+ is 30 volts.

A tapped choke is used to remove the DC current from the yoke, and the yoke is connected back to B+ through an electrolytic capacitor. This arrangement puts the choke in series with the power supply and reduces the ripple on the B+ voltage due to the current of the vertical output. Fig. 3b shows the current that would be drawn from B+ with a one-to-one isolation transformer. Fig. 3d shows the actual supply current using the choke and electrolytic.

The design of the output choke or transformer can be better understood by examining the current waveshapes of the output stage (Figs. 3b, 3c, and 3d). The choke current can be considered as the algebraic sum of the transistor and yoke currents. After retrace, the output transistor is *off* and stays *off* for a significant time. When the output transistor is *off*, the rate of current change in the series yoke-choke circuit must be the correct amount to maintain linearity. The current decays exponentially with a time constant τ given by

$$\tau = \frac{L\text{ choke} + L\text{ yoke}}{R\text{ choke} + R\text{ yoke}} \cong \frac{L\text{ choke}}{R\text{ choke} + R\text{ yoke}}$$

As can be seen from the shape of the yoke current at the beginning of trace,

$$\tau \cong 0.6 \times \frac{1}{60}\text{ sec.} = \frac{1}{100}\text{ sec.}$$

If the choke resistance is approx 10Ω,

$$\text{then } L\text{ choke} = \frac{47 + 10}{100} = 0.57\text{ H}$$

If the choke inductance is too large, the current change is too small, but the amplifier can be made to operate sooner and thus provide good linearity. With this type of circuit when the inductance is too small, the current change is too great and nothing can correct the linearity. In open-loop circuits (Fig. 1), the choke impedance and the electrolytic C_2 have an effect on linearity during all of the trace. In the feedback circuit both these components have little effect because the Miller capacitor is connected directly to the yoke.

The 47-ohm saddle yoke used was less efficient than a toroidal yoke, but the $L{:}R$ ratio is considerably less, resulting in a smaller peak pulse during retrace. Even with the smaller peak pulse, it is still necessary to protect transistor Q_4 during retrace with a voltage-dependent resistor, RV_1.

SYSTEM OSCILLATIONS

When the circuit was first built it had a number of defects, including a 30-Hz oscillation which caused a separation of the two vertical fields.

The waveshape on the base of transistor Q_2 is shown in Fig. 3e at time t_o, when transistor switch Q_1 is closed by multivibrator action and the yoke peak pulse charges capacitor C_1 through Q_1 and R_2. When transistor Q_1 conduction stops, the base voltage of transistor Q_2 rises above the B+ voltage by an amount dependent upon the total charge received by capacitor C_1 during retrace. From time t_1 to t_2, the amplifier is still biased *off*. When the current through R_1 has made the base voltage of transistor Q_2 sufficiently negative, the amplifier becomes operative and the Miller rundown starts. At the end of trace the amplitudes of the voltages on the base of Q_2 and on the yoke are dependent on the time interval $t_3 - t_2$. If time t_2 is large, the Miller rundown time $t_3 - t_2$ is small, resulting in a smaller final current. The amplitude of the retrace peak pulse is dependent upon the yoke current, and t_2 is dependent upon the peak pulse. In summary it can be said that a high-peak pulse causes a low-peak pulse in the next cycle; this is an unstable condition and the system can sustain an oscillation at 30 Hz.

Another way of examining this circuit is to consider it as a pulse-sampling feedback system. This particular pulse sampling introduces an effective phase delay which causes the system to oscillate at half the pulse repetition rate, if the loop gain is high enough. The diode clamp of network F (Fig. 4) reduces the loop gain at 30 Hz and thus prevents oscillation. Time t_2 can be controlled by varying the load on the clamp. This arrangement is used for linearity control of the top of the raster. The clamp allows steady-state changes to be partially transmitted so that it is not necessary to readjust the linearity control when the height control is moved.

High-frequency attenuation must be incorporated in the amplifier circuit to suppress the violent oscillation that occurs in the kHz region. There are several methods of suppressing the oscillation. The most economical method is to connect capacitor C_3 between the base of transistor Q_2 and the collector of transistor Q_3, thus completing a secondary Miller feedback loop within the amplified circuit.

FREQUENCY STABILITY

Frequency stability was not adequate originally; during trace, transistor Q_1 was biased *off* and the waveshape on the base of Q_1 was as shown in Fig. 3f. Retrace started when the voltage on the base of Q_1 changed polarity; this was determined mainly by the $R_3 \times C_4$ time constant. At high temperature, leakage current and the change in base-emitter voltage caused the frequency to increase rapidly. The leakage problem was eliminated by changing to a silicon switch. When the output collector waveshape is inverted and fed through an RC integrator, it looks like Fig. 3g. When this signal is added to the original base waveshape, the desired waveform (Fig. 3h) is produced. This waveform was obtained by simply connecting the frequency-determining resistor, R_3, to a winding on output choke T_1. The resultant waveshape on the base of transistor Q_1 (Fig. 3h) has a fast rate of change at the end of trace, yielding a stable frequency and some noise immunity.

DC BIAS

The specification of I_{CBO} for the output transistor is 2.5 mA at 65°C. Therefore, under worse conditions of ambient temperature and high line voltages, I_{CBO} can rise to 8 mA. This condition can result in a collector current of the order of 500 mA due to leakage alone. Normal DC current at 25°C is 190 mA. Routine methods of obtaining DC stability were not much help. When the output stage became hot, leakage current prevented this stage from being cut off during retrace. The Miller circuit still maintained the correct AC waveshape of yoke current but there was nothing to prevent the transistor DC current from increasing until thermal runaway occurred.

This problem was solved by connecting R_4 to a higher voltage than the B+ voltage. This voltage source supplies sufficient current during retrace to cut off an output transistor with high leakage current. Resistor R_5 was similarly connected to ensure that transistor Q_2 would control the biasing of Q_3 so that Q_3 could always drive Q_4 to cutoff. At present the only effects of leakage current are slight changes in the operating points of transistors Q_2 and Q_3. There is no leakage problem with Q_2 because it is a silicon transistor. With this modification, the current in the output stage is not changed by leakage current.

The resistance of the yoke increases with temperature but this is offset with a thermistor in series with the yoke, as is done in tube receivers.

The time, t_2 (Fig. 3e), that the amplifier starts operating depends partly on the sum of the base-emitter voltages (V_{BE}) of transistors Q_2, Q_3, and Q_4. When the temperature rises, V_{BE} decreases thus making $(t_2 - t_1)$ smaller. To compensate for this, R_2 has a thermistor in parallel with it. As the thermistor resistance decreases, capacitor C_1 receives more charge during retrace, which restores $(t_2 - t_1)$ to the correct value.

RASTER STABILITY WITH AC LINE CHANGE

The raster height is determined by current in the height-control developing a voltage across C_1. As the line voltage increases, the raster height tends to increase due to the different tracking of the vertical circuit and the second anode voltage with B+ change. By connecting the height control to a voltage-dependent resistor, the voltage change across the height control is reduced, and a constant raster height is maintained with varying AC line voltage.

S-SHAPING

The Miller amplifier produces a linear sawtooth of current in the yoke; this results in a stretch at the top and bottom of the raster because of the curvature of the kinescope faceplate. The voltage waveform on the yoke is processed by network G (Fig. 4), which provides an approximately parabolic waveshape output. Network G consists of two RC integrators in cascade. When this parabolic waveshape is added to the constant input voltage, the Miller circuit integrates once more to provide the correct S-shaped sawtooth.

INTERLACE

The multivibrator action is synchronized by a negative pulse from the sync separator. A horizontal voltage is developed on the vertical winding of the yoke due to the cross coupling between the horizontal and vertical windings. If this horizontal voltage is sufficient to affect the turning *on* or *off* of the switch transistor Q_1, then the interlace will be spoiled. Network H (Fig. 4) is part of the multivibrator loop and consists of a double RC filter which reduces the horizontal voltage on the base of transistor Q_1. Network H is also used to integrate the incoming sync signal to remove the horizontal sync pulses.

SUMMARY

The Miller circuit has several advantages over the more common open-loop circuits. Changes in circuit performance due to variations in the active device have been considerably reduced. Long-term stability and reliability of the circuit has been improved by the use of paper capacitors and an additional transistor instead of the electrolytic capacitors in conventional circuits. The Miller circuit is capable of providing vertical deflection for any kinescope and second anode voltage in current use.

ACKNOWLEDGEMENTS

The author acknowledges the contributions made by J. B. Beck, T. J. Christopher, and R. N. Rhodes.

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TECHNOLOGICAL IMPROVEMENTS IN HORIZONTAL-DEFLECTION TUBES

This paper describes the improvements made in horizontal-deflection tubes from the introduction of monochrome TV receivers in 1946 to the present. The trend toward higher output, design considerations for high-current tubes, and the importance of high plate-to-screen-grid current ratios are discussed. Advantages of multi-fin plate structure are noted, and the problem of snivet interference is examined.

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PRESENT-DAY horizontal-deflection systems require a mass-produced horizontal-output tube which has high output, high dissipation ratings, and low screen-grid current, and which is free of spurious high-frequency radiation. New materials, new designs, production and testing improvements, and refinements in circuitry have responded to a continued trend toward stricter requirements, and have made possible the development of substantially improved tubes in smaller glass envelopes.

Final manuscript received February 14, 1966

RISING TREND TOWARD HIGHER OUTPUT

When the first monochrome TV receivers were introduced in 1946, a horizontal-output tube having a zero-bias knee current of 170 milliamperes was suitable for the application. The use of wider deflection angles and higher anode voltages for bright, high-definition pictures has required increasingly higher current from horizontal output tubes. This trend has been intensified by color TV and low B^+ applications, both of which have placed stricter demands on the output characteristics. Today tubes are being developed and manufactured which produce more than 1.0 ampere of current and which are housed in a smaller glass envelope than was previously used.

The evolution of the bulb envelope of horizontal-deflection types is shown in Fig. 1. In 1946 a typical tube, such as the 6BG6G, originally used a large ST16 bulb with an octal base. Improved materials and better mechanical design made it possible to reduce bulb size and increase ratings to meet the demands for higher output. The 1965 tube, type 6JE6A, shown in Fig. 1 is housed in a comparatively small T-12 bulb with a Novar base, and has ratings substantially higher than those of its predecessors. The bar chart identified with each tube type in Fig. 1 indicates the total permissible input power dissipated by the tube. This value, the summation of heater power, grid-No. 2 input power, and maximum plate dissipation, is included to emphasize the successful achievement of higher rated tubes, despite the use of a smaller glass envelope. This development has been made possible through the use of lead-glass bulbs, which withstand higher temperatures without electrolysis during life. New plate materials, including sandwiches of aluminum, steel, and copper, have resulted in more

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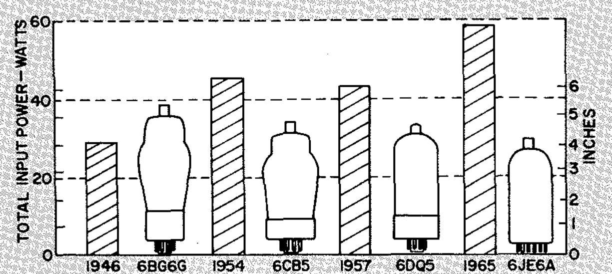


Fig. 1—Evolution of size and shape of horizontal-output tubes, together with increase in total power input from 1946 to 1965.

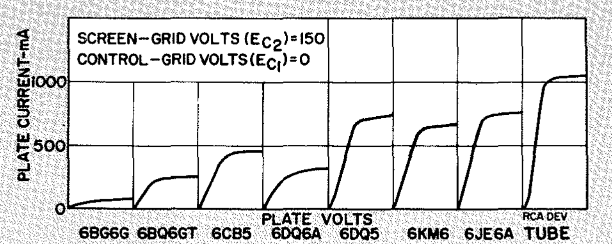


Fig. 2—Plate current as a function of plate voltage for typical horizontal-output tubes developed from 1946 to present, arranged in order of development.

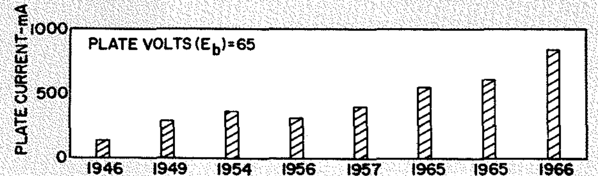


Fig. 3—Performance improvement of horizontal-output tubes.

Fig. 4—Grid lineup performed by experienced operator.

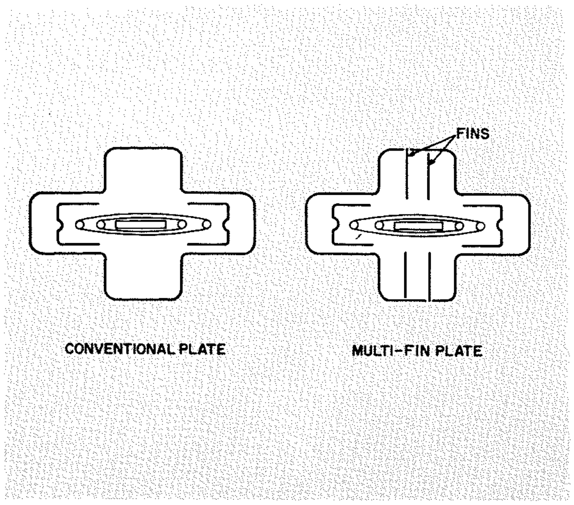


Fig. 5—Cross-sectional views illustrating conventional two-piece plate structure and the multi-fin plate structure.

uniform temperature distribution. Also, the dark, rough outer surface of alitized steel improves thermal radiation. Alitized steel is prepared by cladding steel with aluminum and heating it until a black aluminum-iron compound is formed.

The general trend toward higher output is shown in Fig. 2. The curves show the zero-bias plate current characteristics of typical horizontal output tubes in use from 1946 to 1966. To permit a uniform comparison between types, the curves show plate current at a fixed screen-grid voltage of 150 volts. It is evident that the zero-bias knee current has become progressively higher, with the most recent development for low B^+ color requiring nearly seven times the original output level.

A more realistic comparison is shown in Fig. 3. This chart takes into consideration the limiting factor of screen-grid dissipation, which is important in determining the available output of a horizontal-deflection tube. Plate current is shown at a fixed plate voltage (arbitrarily chosen as 65 volts, even though recent types have a lower knee voltage) and at rated screen-grid input. The trend toward improved performance is obvious.

Over a period of years, the anode power requirements for monochrome receivers have increased from 2.5 to 10 watts, whereas a typical 25-inch color receiver today requires approximately 40 watts. The trend toward wider angle deflection has placed an additional demand upon horizontal-tube output. The recent change, however, from 70° to 90° in color deflection was achieved without placing additional requirements upon deflection tubes. This refinement was possible because the increased angle was compensated for by a decrease in the diameter of the picture tube neck from 2 to 1.5 inches, thus increasing deflection sensitivity.

DESIGN CONSIDERATIONS

To meet the need of high output current and low knee-voltage characteristics, two alternatives are possible:

1) Use two tubes (or two units in one envelope) in a parallel circuit arrangement.
2) Develop a single tube with a high ratio of plate to screen-grid current and an ability to withstand high levels of dissipation.

The first approach doubles the plate current output without sacrificing desirable low-knee characteristics; however, it has the disadvantage that unequal units will frequently produce parasitic oscillations in the application. The second alternative presents a greater design challenge, but it is preferred.

The design parameters involved in obtaining this high plate-current characteristic are shown by the expression for cathode current of a power pentode as follows:

$$I_K = \frac{2.335 \times 10^{-6} A_k}{S^2 g_{1-k}} \left(\frac{Ec_2}{Mu} + Ec_1\right)^{3/2}$$

where A_k = area of the cathode; Sg_{1-k} = control-grid-to-cathode spacing; Ec_2 = screen-grid voltage; Ec_1 = control-grid voltage; and Mu = control-grid-to-screen-grid amplification factor (triode Mu).

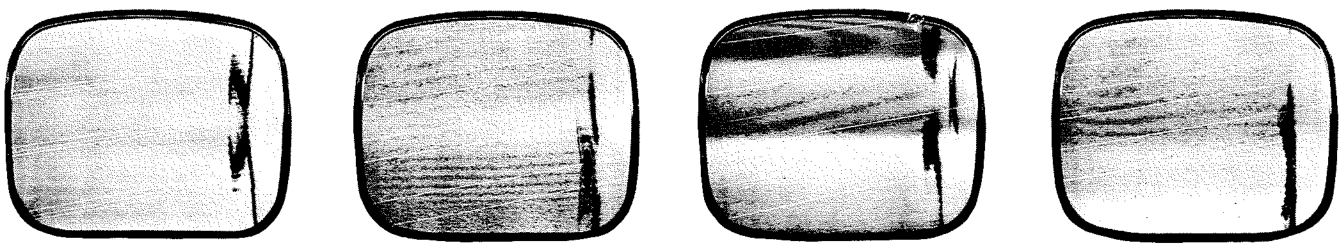
This expression shows that I_k can be increased either by increasing the perveance portion of this expression, $\frac{2.335 \times 10^{-6} A_k}{S^2 g_{1-k}}$, or by decreasing Mu. An increase in Ec_2 is limited by the screen-grid input rating.

RCA tube types 6DQ5, 6JE6A, and a developmental tube type achieve the higher current through higher perveance obtained by use of a greater cathode area and closer grid-to-cathode spacing. The limiting factor in reducing Mu is the need to maintain a good plate-current cutoff characteristic. The cutoff characteristic is highly significant in the application, because the tube must remain cut off during retrace, despite a pulse of several thousand volts on the plate. Recent tube types are rated to withstand 7500 volts on the plate.

PLATE-TO-SCREEN-GRID-CURRENT RATIO

A high ratio of plate-to-screen-grid current is of prime importance in achieving efficient horizontal-output-tube performance. This current ratio is of particular interest because the application requires that the tube be driven down to the knee region, where the plate voltage is low and the screen-grid current is starting to rise. The tube design must incorporate features that produce a high ratio of plate-to-screen-grid current. At the same time, the screen-grid input rating must be sufficiently high to avoid limiting the usable plate current. Low screen-grid current is obtained by two techniques: 1) careful lineup of the No. 1 and No. 2 grids, and 2) optimizing the electron-beam shape. Lined-up No. 1 and No. 2 grid wires cause shadowing of the No. 2 grid wires, which reduces the number of primary electrons intercepted by the screen grid as the electrons pass from cathode to plate. The grid dimensions

Fig. 6—Typical snivet interference on picture-tube raster.



are designed so that at zero bias the electron beams formed between grid lateral wires have minimum cross-section near the screen-grid plane. Thus, the number of electrons intercepted by the screen-grid wires is kept as low as possible, and fewer electrons are returned to the screen grid by the space charge barrier beyond it.

Grid lineup is a manual operation performed by experienced operators who frequently use a microscope for the purpose. Typical facilities for this operation are shown in Fig. 4.

The screen grid is made of carbonized molybdenum wire to increase the work function at the surface of the screen grid. Elaborate precautions are taken to keep the screen-grid temperature low to minimize screen-grid emission. The screen-grid stem leads provide efficient heat conduction away from the screen grid; on high-rated RCA types, two carbonized radiators provide a heat sink at the top of the mount structure.

MULTI-FIN PLATE STRUCTURE

A substantial improvement in overall performance was achieved by the development of a multi-fin plate structure. A conventional two-piece plate is compared in Fig. 5 with the new structure having added fins protruding perpendicularly from the plate wall.

In the conventional plate structure, primary electrons originating from the cathode bombard the plate at high velocity and cause secondary-electron emission. An application in which the plate potential is higher than the screen-grid potential would have no secondary-electron problem; however, in the deflection application the plate voltage, during scan, drops substantially below the screen-grid voltage, causing the secondary electrons to be attracted to the screen grid. These electrons contribute to the total positive screen-grid current and, as previously described, impair tube performance.

With the multi-fin structure, secondary electrons are trapped and do not reach the screen grid, despite the close proximity of the positive screen-grid potential. Plate-to-screen-grid current ratios of 18 in the critical knee region are attainable with such a design.

INTERFERENCE FROM SNIVETS

The secondary electrons described previously can travel all the way to the grid-No. 1 region and cause spurious oscillations repeated at the horizontal sweep rate. These oscillations result in interference (known as *snivets*) which appears as a vertical black bar, especially on UHF (Fig. 6). Snivets can be generated by 1) discontinuities in the plate characteristics coupled with a load line traversing those unstable areas, or 2) operation at zero-bias with the plate voltage substantially below the screen-grid voltage. Unfortunately, maximum circuit efficiency, minimum flyback-transformer cost, and best anode voltage regulation are achieved by operation in the very region most susceptible to snivets.

The load lines of typical horizontal deflection tubes are shown in Fig. 7. The heavy ringing of the load line, which is caused by the resonance of the flyback transformer and is normal, has a pronounced effect on the instantaneous plate voltage of the horizontal-output tube. In Fig. 7(a) the load line properly drives back to the knee of the tube characteristic curve for optimum circuit efficiency and snivet-free performance. In Fig. 7(b) the load line swings to considerably lower values of plate voltage, well below the knee of the tube. This condition may produce satisfactory deflection output, but it is strongly susceptible to snivets occurring at point *s*, or at several other points along the load line.

In addition to eliminating snivets by proper load line and improved knee characteristics, snivet interference can be reduced by applying a positive DC voltage (approximately 30 volts) to the beam plate (grid-No. 3) of the tube. This positive beam plate collects the stray secondary electrons before they return to grid-No. 1 and also lowers and rounds the knee-current characteristic. Recent deflection types all have a separate beam-plate connection to permit the use of a positive voltage.

Examples of good and bad knee characteristics are illustrated in Fig. 8. A very poor plate family is shown in Fig. 8(a). In this example poor grid symmetry results in a very high knee and unsatisfactory output in the application. In addition, the high knee and discontinuities along the diode-line of the plate family contribute to a severe snivet condition in the receiver. The diode-line is that part of the zero-bias plate-current characteristic below the knee.

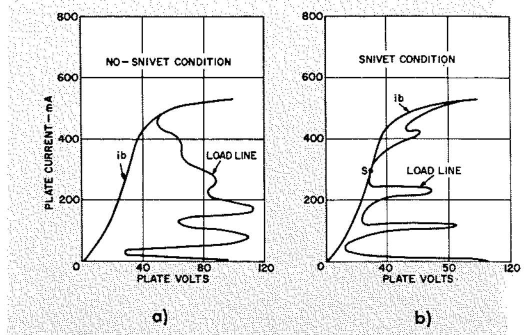


Fig. 7—Load line of horizontal-output tubes (a) snivet-free operation; (b) with snivet interference in picture.

In the plate family shown in Fig. 8(b), the knee is considerably lower; however, there is a pronounced overlap in the plate current at various levels of control-grid bias. This overlap illustrates an under-suppression problem due to improper beam-plate design. The consequence of this design, as described in the previous load-line discussion, is susceptibility to snivets.

Fig. 8(c) represents a horizontal-output tube having a desirable knee shape and a satisfactory diode-line characteristic. The multi-fin plate construction of this tube and the positive bias on the beam plate produce a good rounded knee located at a low value of plate voltage. This type of characteristic can yield optimum performance and snivet-free operation.

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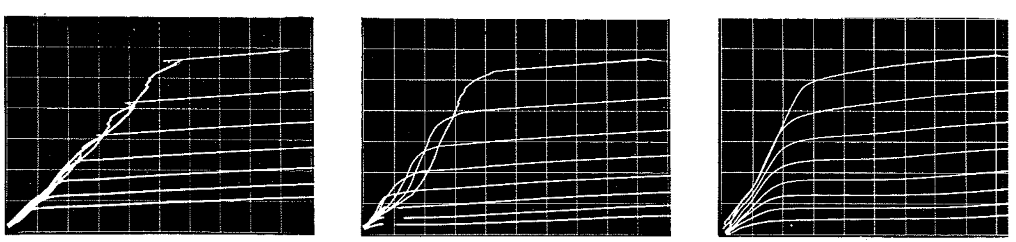
Fig. 8—Examples of knee characteristics of horizontal-output tubes: (a) poor characteristics due to poor grid lineup; (b) poor characteristic due to improper beam-plate design; (c) good characteristic.

a) b) c)



INTEGRATION OF TECHNICAL FACILITIES IN B&W AND COLOR TV PROGRAMMING

In a television plant, Technical Operations is frequently asked to provide facilities for handling programs involving the integration of combinations of film, video tape, live, and outside or field program sources. To avoid picture disturbances when switching to these various program sources, horizontal and vertical synchronization must be maintained between the sources at all times. This paper describes the various methods and procedures used to realize this synchronization both in color and black and white TV.

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It is normal procedure in the final stages of the installation of a TV studio plant to establish in-plant timing. This procedure times the TV synchronizing pulses in the plant to each camera and switcher so that switching between cameras and studios does not create picture disturbances in home receivers tuned to the station.

Color TV requires six synchronizing pulses which must be timed. These pulses include: 1) 3.58-MHz (Mc/s) phase, 2) vertical drive, 3) horizontal drive, 4) kinescope blanking, 5) synchronizing pulses, and 6) burst flag. These pulses are generated by the TV synchronizing generator and are all time related to each other. In distributing these synchronizing pulses throughout a plant, varying delays between pulses are encountered, since the path lengths, and thus the delays, are different. The color subcarrier 3.58-MHz signal is a sinusoidal wave that is easily corrected in delay by a 360° phase control. Vertical drive, a 60-Hz (c/s) signal, does not require cable-length correction because of its relatively long duty cycle. The timing of horizontal drive, horizontal blanking, horizontal sync, and burst flag pulses requires accurate delay correction to comply with FCC requirements. Very accurate relationships must be maintained between all pulses to realize the proper width of front and back porches (Fig. 1). Since studio and switching systems differ both electronically in delay and physically in distance from the input to output terminals, a pulse distribution system must provide in-plant timing so that the various video signals originating at different locations in the plant arrive, or are timed to be the same, at a particular reference location. Timing is difficult if the facilities are different. One studio may have a special-effects amplifier, whereas others may not. Furthermore, similar components, such as a special-effects amplifier made by different manufacturers, do not all have the same delay. In short, the pulse timing and video delay of various equipment components in a signal path must be compensated for in each studio.

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IN-PLANT TIMING

To time a TV plant, the longest pulse and video path lengths must first be determined to establish a reference time basis for the entire plant. Then, lump delays must be introduced in the pulse distribution system so that the path lengths in the smaller studios and switching systems match the pulse and video time reference initially established. In practice, in-plant timing for black-and-white TV is established using the kinescope blanking pulse. The timing of the blanking pulse and the position of the edges of this pulse must be very accurately positioned by the use of proper delays, since these edges are what primarily time the TV plant with the home receiver. In color TV the same procedure must be followed as in black-and-white TV, and it is also necessary to time the 3.58-MHz color subcarrier within the plant. A simple two-studio plant is shown in Fig. 2.

In-plant timing is at best a time-consuming, laborious task. It becomes more complicated when piggyback operation must be provided (i.e., when the output of Studio A, which integrates live and tape segments, must go through Studio B, which integrates the output of Studio A with live program and film commercials).

Piggyback operation must be designed into the plant. Means must be provided for switchable pulse delays. The approach of in-plant timing for piggyback operations is again to first establish maximum stacking of studios and thus pulse delays. A practical piggyback limit for a large network station is three studios. Beyond three-studio stacking, the switching of pulse delays, and possibly video delays, becomes inordinately complicated and expensive.

A simplified version of the two-studio TV plant with video tape and film facilities is shown in Fig. 3. If piggyback operation is anticipated, switchable pulse delays are required as shown in Fig. 4. The plant ordinarily can operate either with all delays in or all delays out.

A hypothetical TV plant timed to operate in a two-studio piggyback mode is shown in Fig. 5. It should be noted that the pulse signals fed to switcher A are not delayed. Delay is added to the Studio B video source so that the output of Studio A and all the video sources of Studio B arrive at the input to Studio B switcher at time T_2.

In-plant timing is not something done

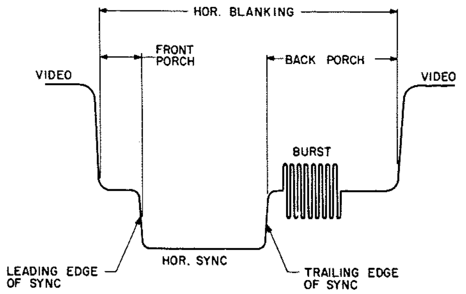

Fig. 1—The position of the horizontal sync pulse during horizontal blanking time is specified by the FCC. The leading edge of the pulse is used in receiver locking circuits; the trailing edge is frequently used to trigger clamp circuits in broadcast equipment.

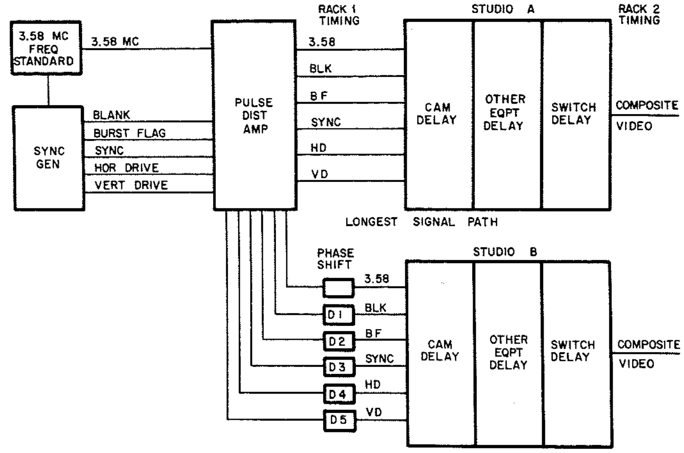

Fig. 2—Two-studio color plant timing. Studio A composite video output is the longest signal-path time of plant. Studio B composite video output must be made equal to Studio A in time by adding delays D1 through D5.

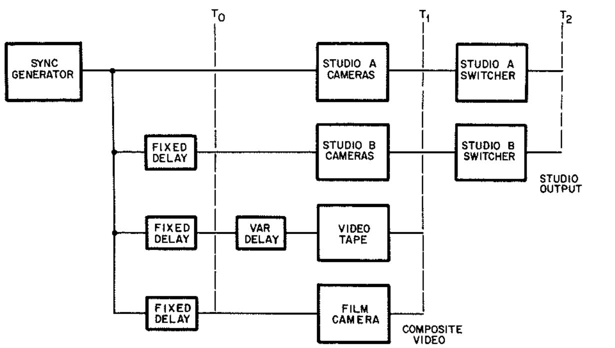

Fig. 3—Simplified version of two-studio TV plant with video tape and film facilities.

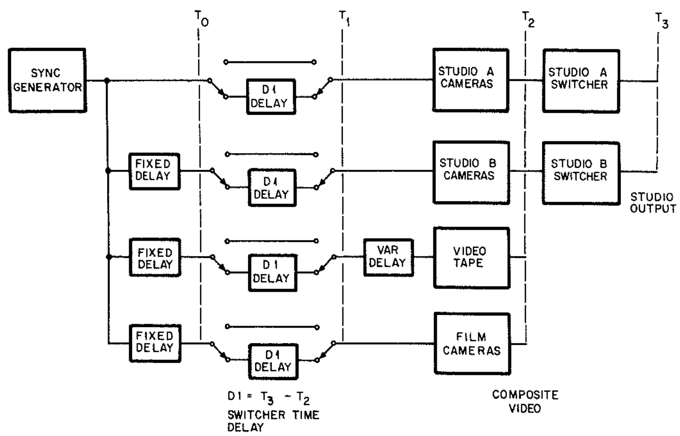

Fig. 4—Switchable pulse delays required for piggyback operation.

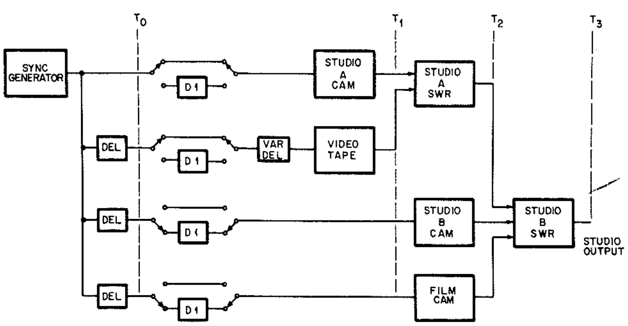

Fig. 5—TV plant timed to operate in two-studio piggyback mode.

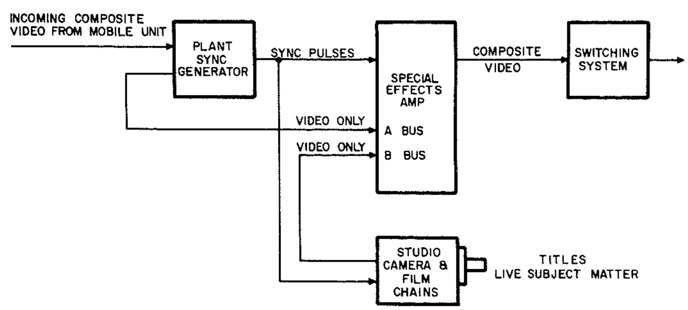

Fig. 6—Simplified diagram of genlock system.

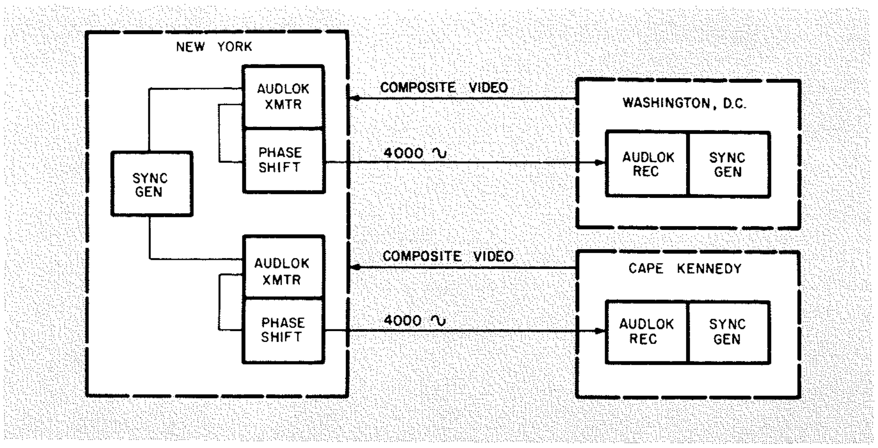

Fig. 7—Typical Audlok system.

only at the time of the installation of a TV plant. It is a continuing job. A change in switching systems, the addition of distribution amplifiers, new cameras, or film chains, or any other change that will affect pulse or video timing must be carefully studied before integration into a plant. The mixing of cameras of different manufacture or the integration of black-and-white and color cameras in a plant can create knotty problems. In general, the older the TV plant, the more difficult it becomes to expand facilities.

GENLOCK

The subject of remote timing comes up whenever the Program Department requires mixing, fading, or dissolving from the TV studio to a mobile field pickup. Then it is necessary to time-lock the TV plant sync generator to the mobile unit sync generator so that there is no picture disturbance when switching from studio to mobile unit or vice versa during broadcast. A technique known as *genlock* is used to operate two sync generators in series (or synchronism). In principle, the plant sync generator is locked to the mobile-unit sync generator. A single-line drawing of a genlock system is shown in Fig. 6. The TV plant is made available to the mobile unit, in a time sense, so that the TV picture can originate at either the studio or the mobile unit with no disturbance in picture during switching or dissolving to and from either program source. The system works equally well in black-and-white or in color TV. In color, of course, accurate phasing of the color subcarrier is required at the plant sync generator. This paticular function is not required in black-and-white operation.

AUDLOK

In a large TV plant it often happens that video tape recording, kinescope recording and TV broadcasting occur simultaneously. The TV plant may also be in a genlock mode for a mobile-unit pickup to be taped. The next program might, for instance, require the integration of a show with program control in New York with portions of the program from Washington, D.C. and Cape Kennedy. In such a situation, it is necessary to switch, dissolve, etc., to each pickup point without picture disturbances.

The problem is to time the program sources from Washington and Cape Kennedy to arrive at the same time in New York and be coincident with New York sync-pulse time. Stated in another way, with New York as a base of reference, Washington must be timed in advance of New York by the transit time of the signal between Washington and New York. Likewise, Cape Kennedy must be timed in advance of New York by the transit time of the signal between Cape Kennedy and New York.

It is not possible to use the ordinary genlock system of pulse synchronization in this problem. Genlock is a forward type of locking and is the condition whereby the control studio locks to the outside program source. Genlock might be used to synchronize New York to Washington, but it could not be used simultaneously to lock New York with Cape Kennedy.

A system known as *audlok* has been devised to solve this type of synchronizing problem. Audlok might be described as a backward type of genlock that synchronizes both Washington and Cape Kennedy to a New York time base. As the term implies, an audio frequency is used to time-lock synchronizing generators. In the system shown in Fig. 7, an audlok transmitter is located at Master Control in New York, and an audlok receiver, to be synchronized with New York, is located at Washington, D.C. Another audlok transmitter and associated receiver are located in New York and Cape Kennedy, respectively. Each audlok transmitter and associated receiver are interconnected by means of a rented telephone circuit. The horizontal line frequency of the New York synchronizing generator is processed by the audlok transmitter to produce a submultiple audio-frequency sine wave of about 4000 Hz, which can be easily and quickly phase-shifted by any desired amount. The audlok receiver in Washington processes the 4000-Hz tone from New York and multiplies it to lock the Washington sync generator. Then, by superimposition of New York and Washington pictures and by means of phase control, the Washington sync generator can be accurately timed, both horizontally and vertically, so that the Washington pulses are coincident with New York pulses at arrival in New York. With another audlok system, the sync generator at Cape Kennedy can be similarly adjusted so that the Cape Kennedy pulses are coincident with New York pulses. Thus, the timing problem between New York, Washington, and Cape Kennedy is solved.

FREQUENCY STANDARD LOCKING TECHNIQUE

The success of audlok has suggested several other schemes to eliminate the need for time-control circuits between pickup points and the control location. One such scheme would use very stable 3.58-MHz oscillators to control TV sync generators. The stability must be in the order of one part in 10^{11}. It is reasoned that if a stable 3.58-MHz oscillator is used to time a New York sync generator and another 3.58-MHz oscillator is used to time a sync generator in the Washington, D.C. studio, it should be possible for New York, on viewing a TV picture from Washington, to arrange by telephone for framing adjustment at one end or the other. Such a system is now in operation in black-and-white TV and requires only daily checking and adjustment of phase each morning. Tests in color TV indicate that at this stage the technique does not provide sufficient stability of color subcarrier reference to permit dissolves and special effects due to lag and short time-phase delays in the intercity circuits.

TRANSLATOR

Several systems of time-locking two or more television synchronizing generators have been described. In all these systems, timing is the common denominator of the synchronizing process. Time synchronization is possible only when the horizontal line frequency and picture frame rates are the same between two or more synchronizing gen-

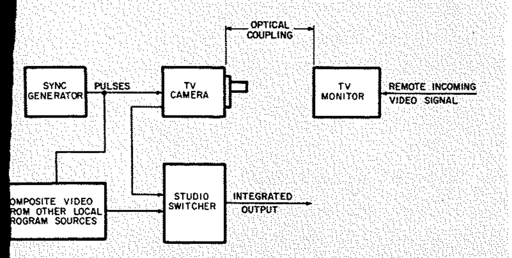

Fig. 8—Simplified diagram of translator.

erators and time is the only parameter that requires synchronization.

Many other disciplines have been confronted with the problem of time synchronization. In those instances where the timing differential has not been too large, some form of energy or information storage has solved the problem. Such a system of video information storage, where time synchronization cannot be used, has been developed for use in television. This system is known as *picture translation*. That is, a TV picture on one time base is translated to a TV picture on a different but compatible time base by video information storage. Translation is accomplished in a device called a translator (Fig. 8), which consists of a high-quality TV monitor and a TV camera. The camera is adjusted for full monitor scanning. The monitor displays the nonsynchronous program televised from some distant location. The translator camera, operated on the local sync generator pulses, views the monitor. The reading and writing rates are the same in this optically coupled system, but differ in time reference. However, because of the information storage in this system, i.e., storage characteristics of the kinescope phosphor and information storage of the camera tube, the difference in time reference or phase of the system no longer is a problem. In short, by means of video information storage, two time systems that differ only in phase can be coupled. Therefore, the translator camera can be integrated with any other program source originating in New York without any picture disturbance during switching time.

The translator must be a unity device to avoid picture degradation. It is also necessary to keep noise to a minimum to minimize picture degradation. Such a system has been used by several major broadcasters for some time and has proven very successful in black-and-white TV. It has been employed with some success in color TV using a tricolor kinescope and a color camera. Considerably more development is required to realize optimum translation in color.

STANDARDS CONVERTER

With the advent of video tape, the TV broadcaster has been faced with a more difficult timing problem: the integration of a video tape recording made on any one of the several European standards into a program to be broadcast on American TV standards. In this instance one must contend not only with a difference in horizontal-line frequency but also with a difference in frame rate. The problem has been solved to some extent by a device called a *standards converter*. As the title implies, the device is used to convert from one TV frequency standard to another. In concept it is similar to the translator previously described. It differs from the translator in that instead of reading and writing on the same TV standard, it reads on one standard and writes on another. Storage minimizes the problem of reading and writing at different horizontal frequencies. There is a problem, however, in reading and writing at frame rates which differ by large time increments. If you read or write at a 60-Hz rate, and write or read at a 50-Hz rate, the resultant difference of 10-Hz manifests itself as an annoying flicker. Circuits have been developed to minimize this 10-Hz flicker, but it has not as yet been completely eliminated. Multi-standard video-tape machines and standards converters have been in use for many years in black-and-white TV. As yet a color TV standards converter has not been made available to the broadcasting industry. Color translators and color standards converters can be made using the same approach used in black-and-white TV. But the complexity of demodulating a color signal to display the red, green, and blue components of the signal on three kinescopes, the registration required, and the picture degradation inherent in such a process have all but discouraged everyone from using this approach. Successful translation by electronic means has been accomplished in Europe for horizontal timing differences only (e.g., the translation of a 625-line, 50-Hz TV picture to a 405-line, 50-Hz TV picture). There is reason to believe that in the future a similar electronic approach will be developed to convert from European to American TV standards in color.

SUMMARY

The average TV receiver can accept without picture disturbance some discontinuity or timing error in synchronizing pulses, both vertically and horizontally, which may occur during the course of switching a program. When not tuned to a TV station the TV set will scan a raster and utilize the synchronizing signal from a station only to *lock in* the receiver vertical and horizontal oscillators. Many components of equipment used in a broadcast plant, however, are *driven* by the synchronizing generator, and any discontinuity or timing error of the pulses can result in a serious discontinuity in the TV picture. The video-tape machine is one such device that is very sensitive to any synchronizing pulse discontinuity or timing error. Experience indicates that synchronizing-pulse timing must be maintained within certain limits. The maximum timing error of vertical framing that can be tolerated during a program fade, dissolve, or switch, must not exceed the time of half a TV line or $\pm$ 32 usecs. The tolerance of the timing of kine blanking must be maintained to $\pm$ 0.05 usec to avoid visible picture disturbance during program switching. Finally, in color TV the phase of the 3.58-MHz color subcarrier must be adjusted to $\pm$ 3° to avoid color shift during a program switch. These tolerances must be maintained not only during a program originating from the local TV plant but between all other program sources (including local mobile unit, intercity pickup, or coast-to-coast pickup) that are synchronized by any of the methods described above for color or black-and-white TV.

In timing TV synchronizing signals that are not identical in line rate or frame rate, a more fundamental problem in time is encountered. Standards conversion as described is only a report on the state of the art; the technique is adequate but is not considered the ultimate solution. Some thorough investigations have been made of this problem. Information theory offers some possible solutions. A solution is to sample or quantize video information on one standard and reconstitute a picture on the desired standard. Results of such approaches have been encouraged and are useful in some disciplines, such as space and military applications. As yet, however, the results have not been considered acceptable for commercial TV broadcasting.

ACKNOWLEDGMENTS

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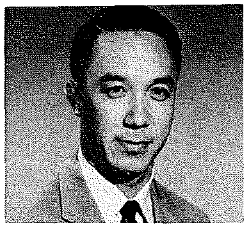
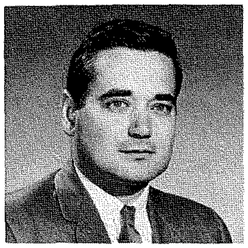


H. Amemiya

T. R. Mayhew



R. L. Pryor

H. AMEMIYA obtained his BSEE degree from the University of Tokyo in 1948 and a MS degree from Cornell University in 1952. He joined RCA in 1957 and has been engaged in a number of projects, including developmental work on a remote-control system for the gas industry, and the design of a magnetic-drum memory system for the RCA 110 Industrial Computer. He was the project engineer on the two-core-per-bit memory system about which this paper is written. He is currently engaged in advanced memory projects at DEP Applied Research.

T. R. MAYHEW received his BEE degree from the University of Florida in 1956 and his MSEE degree from the University of Pennsylvania in 1962. He joined the Industrial Advanced Development Section of RCA in 1956 and was assigned to the development of automatic inspection equipment for the paper and glass industries. In 1958 he joined the Industrial Computer Section where he was engaged in the design of a magnetic drum memory system for the RCA 110 Industrial Computer. Reassigned to the Computer Advanced Product Development Section, he participated in the development of a number of high-speed memory systems, including magnetic thin film and core memories. In 1963 he developed circuitry for digital threshold logic systems. More recently he participated in the design of memory drive circuits for the integrated-circuit RCA Spectra-70 Computer. He is presently engaged in high-speed digital circuit development in the Applied Research Department. Mr. Mayhew is the coauthor of two papers on memory systems.

R. L. PRYOR was graduated from the Missouri School of Mines and Metallurgy with a BS in Physics in 1955. He also attended California Institute of Technology and Iowa State University. Since joining RCA in 1955, he has worked on high-speed memory circuit development in the Computer Advanced Product Research Group, Applied Research. His engineering experience at RCA has included analog computer circuit design, developmental work on a glass inspection machine, a metal detector, and circuit design for a magnetic drum memory. For the past four years he has been engaged in system development and circuit design of various types of ferrite random-access memories. Mr. Pryor is the coholder of two patents in the ferrite memory area, and the coauthor of a paper on a high-speed microferrite memory.

HIGH-SPEED, TWO-CORE-PER-BIT, FERRITE MEMORY SYSTEM

Operation of large high-speed memories is complicated by the generation of noise and transient conditions that add considerably to memory cycle time. Thus, control of these transients is an important consideration in the design of the electronics for a high-speed memory. As a new approach to controlling transients, digit lines are treated as a system of mutually-coupled parallel transmission lines and are terminated for the existing modes of wave propagation. The memory system described in this article employs two cores per bit and is operated with partial switching, using transformer-coupled word lines to reduce the noise generated when a word is selected. Signal sensing is accomplished with a differential amplifier which employs a tunnel diode in the final stage as a threshold. Cables and strip lines are used throughout for interconnection, and some circuits are located on the memory stack to aid high-speed operation. The system provides an operating cycle time of less than 0.5 microsecond.

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Recent advances in computer technology have greatly increased the demand for very-high-speed memories. Scratch-pad memories of up to 128 words with cycle times faster than 500 nanoseconds are commonly found in computers on the market, but larger memories of the same speed range are not yet commercially available. The unavailability of larger memories is due to the complexity of the problems encountered in building a large memory, a much more complicated task than the building of a small memory. To better understand these problems, a 1024-word, 100-bit memory system was built. Storage cells consisted of a linear-organized array of ferrite cores in a two-core-per-bit arrangement. To simplify core-threading, only two conductors per core were used, one of which was plated, leaving only one to be threaded.

MEMORY CELL OPERATION

Linear selection (word-organized memory) and partial switching[1-7] are the two techniques commonly employed to achieve fast cycle times in high-speed ferrite memories and, consequently, are used in the system described here. Linear selection offers the advantage that read currents of large amplitude (limited only by the drivers) can be used to increase speed. This method contrasts with coincident current selection, in which read currents are dictated by the threshold characteristics of the ferrite cores used. Partial switching, in contrast to full switching, decreases memory cycle time by narrowing the drive pulses.

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Linear selection and partial switching techniques can be used in single-core-per-bit operation. However, as memory speed is increased by continually narrowing the width of the write and digit pulses and, subsequently, the width of the read pulse, a point is reached where two-core-per-bit operation becomes necessary. There are two reasons for this. First, the sense signal generated on reading a *0* becomes large as the rise time of the read pulse is decreased. Second, the difference in sense-signal amplitude when reading a *1* and reading a *0* becomes small, because the digit pulse in the presence of the write pulse switches irreversibly only a small fraction of the core. Two-core-per-bit operation provides a means of cancelling the reversible flux contribution to the total sense signal.[8-11]

The particular scheme used in the memory under discussion is shown in Fig. 1. Each core is threaded with one conductor in the word direction and the other in the digit direction. When writing, both core *A* and core *B* of the same bit pair receive a write pulse. In addition, either core *A* or core *B* receives a digit pulse, depending on the information being written in. When reading, a read pulse is applied to both core *A* and core *B* in the direction opposite to that of the write pulse. Sense signals, generated at both core *A* and core *B*, are added in a differential sense amplifier where that portion of the signal due to reversible flux change is cancelled. Noise that is capacitively coupled from the driven word line is also cancelled. Therefore, only the net signals (Fig. 1-c) reach the threshold circuit of the sense amplifier. This two-core-per-bit scheme has the following features:

1) Bipolar sense signals provide more reliable sensing than a unipolar sense signal.
2) Word line impedance is constant regardless of the information pattern, because each bit (i.e., each pair of cores) presents a constant impedance to word pulses even if a *1* or a *0* is stored.
3) Read and write pulses may have loose tolerances.
4) Balanced digit lines that are paired for one bit location offer a possibility of controlling wave propagation inside a memory stack.

The ferrite cores used in this memory have an outer diameter of 0.030 inch, an inner diameter of 0.010 inch, and a thickness of 0.010 inch. Operating conditions are shown in Table I.

TABLE I.
Operating Conditions of the Ferrite Cores

Drive Pulses	Amplitude (mA)	Rise Time (ns)	Fall Time (ns)	Width at 50% points (ns)
Write	220 ± 5%	40	40	80
Read	630 ± 5%	80	30	80
Digit	70 ± 3%	30	30	75

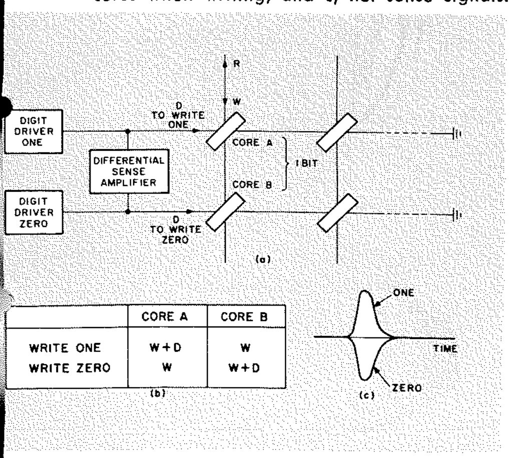
Sense signal from cores — ± 40 mV
Kickback voltage when reading — 0.25 volt/bit

WAVE PROPAGATION IN THE MEMORY STACK AND TERMINATIONS

A basic requirement of a memory system is that it must be able to store any information pattern desired at any word location. Since some words are located close to the digit drivers and sense amplifiers and others are located far away from them, digit lines must be able to carry digit pulses and sense signals without distortion. The two-core-per-bit scheme of Fig. 1 uses paired digit lines. Since either of the two lines of a pair is always driven by a digit pulse, regardless of whether a *1* or a *0* is being written in, digit pulse propagation can be considered a superposition of a differential-mode component and a common-mode component (Fig. 2) with current amplitude one-half that of the digit pulse on each line.

Digit lines are terminated to eliminate reflections, because undesired reflections reduce system reliability and prolong cycle time. However, digit lines propagating the differential-mode component require different impedance terminations than those propagating the common-mode component. The T-termination network shown in Fig. 3 will terminate lines with both components without reflections. The digit line transposition method equalizes coupling between any two adjacent line pairs and eliminates the bit-to-bit interference problem. Note in Fig. 3 that the digit drivers and sense amplifiers are connected to the midpoints of the digit lines. This connection minimizes the digit line delay measured from the driving and sensing point.

Fig. 1—Two-core-per-bit scheme: a) basic read-write scheme, b) magnetization applied to cores when writing, and c) net sense signals.



DESIGN OF MEMORY STACK

In the memory, one of the two conductors passing through each core is a conventional wire and the other is a plated conductor. Fig. 4 shows how memory cores are assembled into a strip. Individual cores are first metalized by vacuum deposition and then inserted into a groove cut in the middle of an insulator strip etched with connecting conductors. The strip is then electroplated to improve contact and to lower the overall resistance of the conductive path. Since the connecting conductors on an insulator strip connect two neighboring cores on the same side, the resulting conductive path has a zigzag pattern.

Each ferrite core strip contains 128 cores, and each memory plane holds 200 strips. Plated conductors were used as digit lines because they permit the pairing of two neighboring conductors to form a bit pair. Such pairing helps to maintain good balance between the two lines of a pair and aids in simplifying transposition. If, on the other hand, the plated conductors are used as word conductors, it is necessary to pair two nonadjacent digit lines because of the zig-zag pattern of the plated conductors. Another reason favoring the use of plated conductors as digit lines is that heating of cores due to repetitive switch-

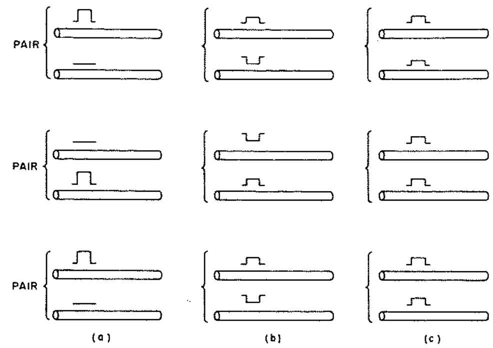


Fig. 2—Propagation of digit pulses: a) digit pulses on digit lines, b) differential mode component, and c) common mode component.

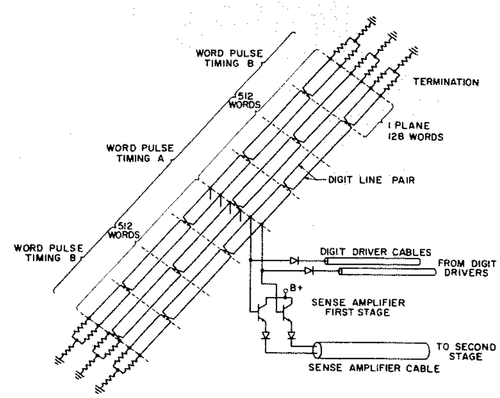


Fig. 3—View of unfolded digit lines showing T terminations, line transpositions, and connections to digit drivers and sense amplifiers.

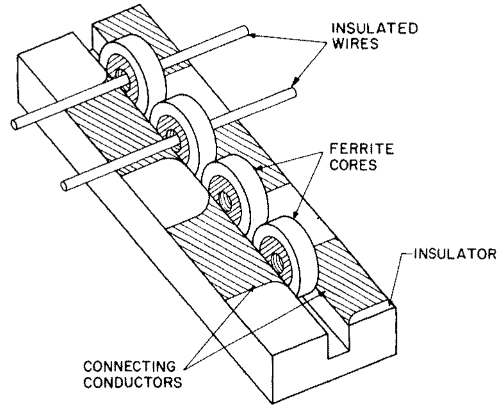


Fig. 4—Ferrite core strip.

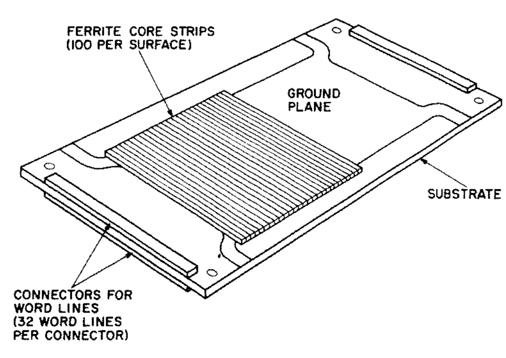


Fig. 5—Memory plane construction.

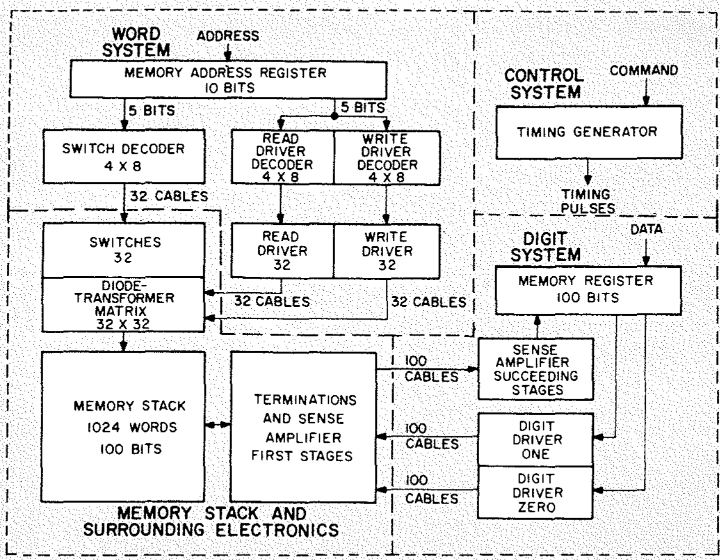


Fig. 6—Block diagram of the memory system.

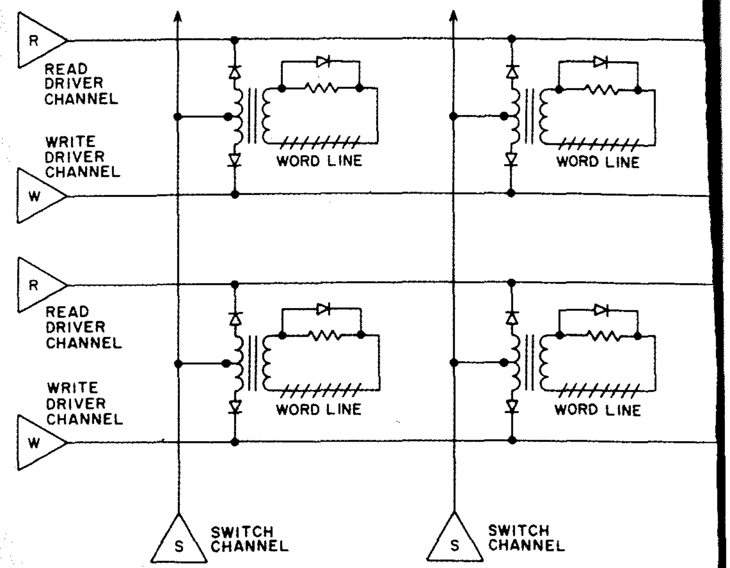


Fig. 7—Portion of the 32 x 32 bipolar matrix.

ing is greatly reduced when every core in a word is individually threaded by a plated digit conductor which doubles as a heat sink. This cooling effect by the plated conductors is very important when a word is addressed repeatedly at a high repetition rate.

A memory plane consists of a substrate and 200 ferrite core strips (Fig. 5), of which 100 are mounted on the top surface and the remaining 100 on the bottom surface. Ground planes are provided on the top and the bottom surfaces of the substrate, over which ferrite core strips are placed. The ground planes are connected to the supporting structure at the four corners of the memory stack assembly. Eight memory planes are assembled to form a complete memory stack.

ELECTRONICS FOR THE 1024-WORD MEMORY

The memory consists of four major portions (Fig. 6): 1) memory stack assembly, 2) control system, 3) word system and 4) digit system. The control system generates and supplies all timing pulses for the drive system and for data transfer. The word system, at the command of the control system, supplies the proper read and write current pulses to a selected word for reading and writing information out of or into the memory. The digit system is used in a dual fashion 1) to provide sensing of the information stored and 2) to write back into the memory, simultaneously with the write pulse, formerly stored or new information. Parts of both the word system and the digit system are packaged on the memory stack.

Control System

The control system generates the necessary timing pulses for each of the three cycle types: 1) read, 2) write, and 3) split. The first two are standard cycles for destructive-readout, random-access memories. Both require a two-step cycle (read-regenerate or clear-write). The unusual feature of the split cycle is that it combines the read-write operation and, thus, eliminates the regenerate and clear operations. The first command for this cycle generates only a read operation accompanied by a strobe of the sense amplifier. The retrieved information is available for processing but is not regenerated since the entire memory cycle has been temporarily suspended. When the second command is given, the memory register is reset again to receive the newly processed information, which is then stored in memory. Thus, the first half starts a conventional read cycle which stops itself in the middle, to continue upon later command as a write cycle after clearing the memory register. The time-saving features of this type of cycle are compatible with many of the common computer operations.

Word System

The word system is required to distribute a large read pulse with the generation of minimum noise, followed by a smaller write pulse of opposite polarity, to any one of the 1024 words addressed by the 10-bit address register. The bulk of this decoding is performed in a bipolar diode matrix (Fig. 7). The matrix is driven by 32 pairs of read and write drivers along one side and 32 switches along the other side. The 1024 intersections of this main matrix are transformer-coupled to the 1024 word lines of the memory stack. The DC level of the word line is restored by a diode-resistor network in the secondary of the transformer. Without this network, a DC quiescent current in the word line will build up at high repetition rates because the read current is larger than the write current.

The main problem encountered in designing a word drive system for a high-speed, high-bit-capacity memory is the minimizing of noise introduced into the stack. Since the electronics are a significant cost factor, it is desirable to use a bipolar diode matrix that performs the selective function with drivers and switches. Experience with the various types of bipolar matrices has revealed that severe switching transients are introduced when the switch selection is made. Moreover, it was found that the characteristics of a memory stack show much tighter capacitive coupling between the network of word lines and the network of digit lines than can be made to exist between either of these networks and a ground plane. The result is a tendency for the conventional word selection matrix to introduce a very sizeable noise pulse onto the network of digit lines.

The purpose of the pulse transformer in each word line is the capacitive decoupling of the word selection matrix and the memory stack. Since the interwinding capacitance of the transformer is a maximum of 7 picofarads, whereas the capacitance between a word line and all the digit lines connected together is about 60 picofarads, the resulting switching noise attenuation is about 10 to 1. This noise must be displaced in time from the sense signal by causing the timing pulse for the switch to start earlier than the timing pulse for the read driver. A switch is turned on for a specific length of time to let the read and the write currents go through; otherwise, no switch stays turned on. The switch noise is appreciably reduced by holding the switches off until after the memory address register has completely settled from the address transfer transient. Otherwise, a spurious selection of switches during the address transfer transient will inject additional noise into the stack. By turning off the switch as soon as the write pulse is terminated, the problem of slow switch turn-off can be easily eliminated.

Another closely associated problem is the injection of noise via the half-selected word lines controlled by the same switch during the read pulse. This condition results from the passage of the read pulse through the finite impedance of the switch circuit, and the mechanism of noise injection is very similar to the one described previously. This type of noise is a threat because it always coincides with the sense signal. It is obvious that this problem can be minimized by lowering the impedance of the switch circuit. A low switch impedance is also desirable from the standpoint of matrix operation, because it permits unimpeded flow of the read and the write pulses. The problem was solved by eliminating cables connecting the switch channels and the word selection matrix and, instead, packaging the output stages of the switch channels at the memory stack. Again, the isolation provided by coupling transformers alleviates the noise problem greatly.

As shown in Fig. 3, the read and the write pulses have two different timings, designated *A* and *B*, depending on the word address. This feature is necessary because the digit line delay of 20 nanoseconds from the driving and sensing point to the termination is not negligible compared with the drive current widths. In the present memory, the strobe and the digit pulses are fixed and the read and write pulses are shifted according to the word address. The 1024 words of the memory are divided into two groups of 512 words each, with one group being closer to the digit-driving and sensing points than the other. The former group uses word pulse timing *A*, the latter group word pulse timing *B*.

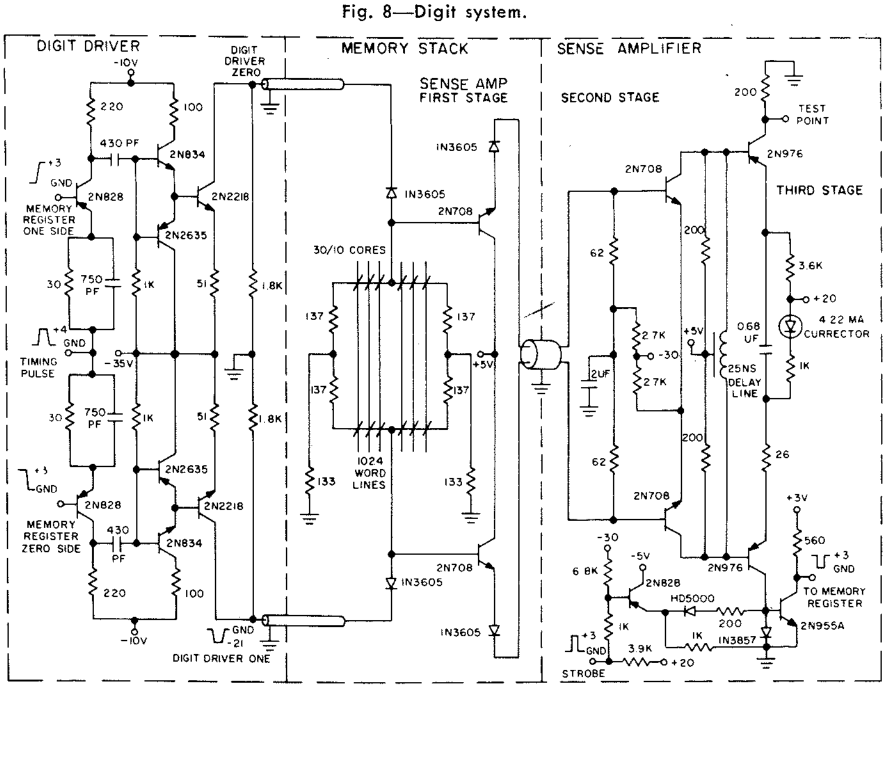
Digit System

The digit system (Fig. 8) consists of the circuits that detect and write or regenerate information in each of the 100 bits of a selected word. During the write time, the digit driver feeds a current pulse to one of the two digit lines to add to the write pulse in one of the two cores of a memory bit. The digit driver consists of two identical current drivers under the control of the timing generator and a flip-flop in the memory information register.

Digit lines are terminated at both ends to reduce the recovery time of the memory stack. As a result, only half the difference signal from the two cores of a bit is available at the sense amplifier input. The difference signal is bipolar, where one polarity represents a *1* and the other polarity represents a *0*. The sense amplifier amplifies the difference signal and is strobed during a portion of the read time. The polarity of the sense signal at strobe time is sensed and if a *1* is detected, the sense amplifier produces an output pulse which sets a flip-flop in the memory register. If a *0* is sensed, no change occurs at the sense amplifier output.

The first stage of the sense amplifier consists of two emitter followers connected to the center of the digit lines (Fig. 8). The high input impedance provided by this stage prevents the sense amplifier from loading the digit lines and from interfering with the termination of the lines. The emitter followers and series diodes are physically mounted near the center of the memory stack and are connected by a 125-ohm shielded twisted-pair cable to the plug-in board that contains the regeneration loop circuits.

Fig. 8—Digit system.



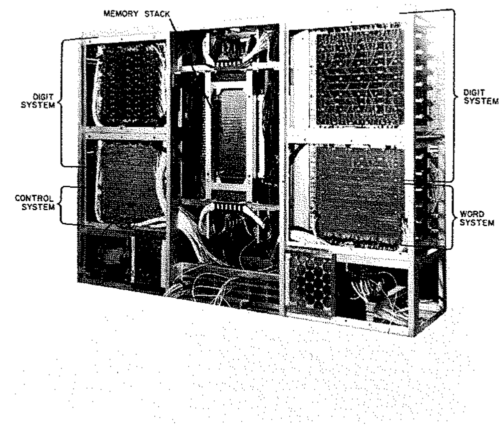


Fig. 9—Memory system.

The second stage of the sense amplifier is a differential amplifier with the transistor collectors connected together through a delay line. This stage amplifies the difference between the input signals and sums the inverted amplified difference signal and the delayed amplified difference signal. This action produces output voltage waveforms at the collectors that do not have a DC level shift with repetition rate variations. The delay of the delay line is long enough that a usable amount of the inverted amplified sense signal is passed before the output is reduced by the delayed amplified sense signal. The delay in this system is 25 nanoseconds, approximately one-half the base width of a sense signal.

The third stage of the sense amplifier is an AC-coupled differential amplifier. One output is used as a test point for observing amplified sense signals, and the other output drives the next stage. The last stage is the strobe and pulse-stretching circuit, which contains a bistably biased 5-milliampere germanium tunnel diode which drives an output transistor. The tunnel diode is normally biased in the low-voltage state and is unable to switch to the high-voltage state during the digit transient because of the current-limiting action of the third stage. During a portion of the read time the sense amplifier is strobed by removing the inhibit current, thereby biasing the tunnel diode in the low-voltage state near the knee. A difference signal of 5 millivolts at the input of the sense amplifier in the polarity of the *1* signal is sufficient to trigger the tunnel diode to the high-voltage state. The tunnel diode turns on the output transistor, which produces a pulse used to set a flip-flop in the memory register. The tunnel diode remains in the high-voltage state until the inhibit current is applied by the strobe circuit. The inhibit current resets the tunnel diode and terminates the output pulse.

PACKAGING

The circuits of the memory, except for those parts that had to be near the memory stack for special reasons, are packaged in four nests surrounding the memory stack (Fig. 9). Each nest has 10 removable mother boards, each of which can accommodate up to 56 small plug-in modules. Individual modules contain such parts as logic blocks, portions of drivers, and portions of the sense amplifiers. When a nest is completely assembled, all of the circuits within it are interconnected by 70-ohm, printed-strip wiring on both sides of the mother boards and grandmother boards. (Mother boards plug into the perpendicular grandmother boards.) Nests are interconnected by coaxial cables. Some of the memory circuits that did not lend themselves to modular packaging, because of power dissipation or size considerations (such as driver output stages), were packaged on specially built mother boards by removing some or all of the provisions for plug-in modules. All logic level interconnections are made via 70-ohm cables, and read and write driver outputs are transmitted to the bipolar diode matrix at the stack via 70-ohm cables. To obtain a lower impedance, the output stages of the switch channels are located at the memory stack. These stages are connected to the rest of the switch channels via 50-ohm cables. The digit driver outputs are transmitted to the stack via 100-ohm cables, and the sense amplifier first stages are connected to the rest of the sense amplifier by twisted-pair balanced cables having a common ground sheath and a differential impedance of 125 ohms.

TEST RESULTS

The 1024-word, two-core-per-bit memory was built with a complete word system and a full digit system of 100 bits.[12] Also, a special memory exerciser was built for thorough testing of the memory system. Fig. 10 shows switch voltage, read and write currents, and digit current waveforms. The switch waveform was observed at the center tap of a word line transformer primary winding (Fig. 7). The undulations on

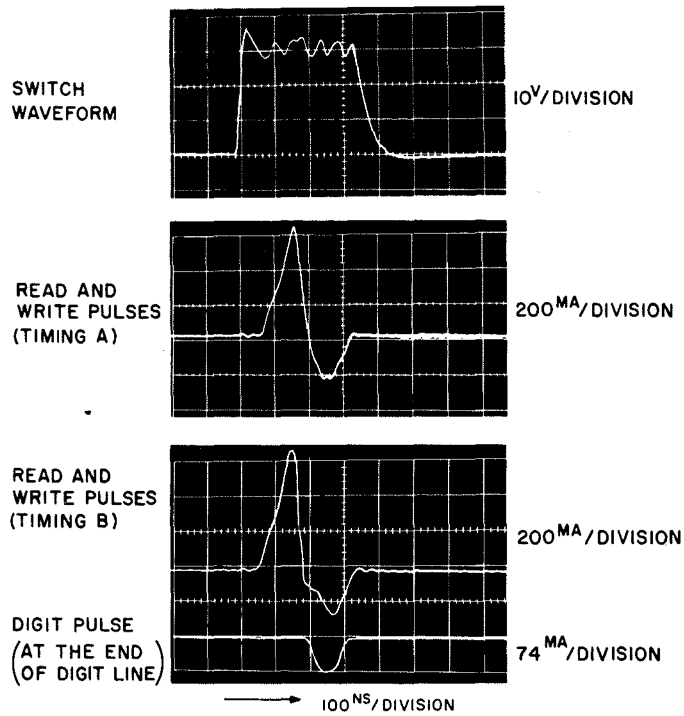


Fig. 10—Switch voltage, read and write pulses ("Timing A" and "Timing B"), and digit pulse.

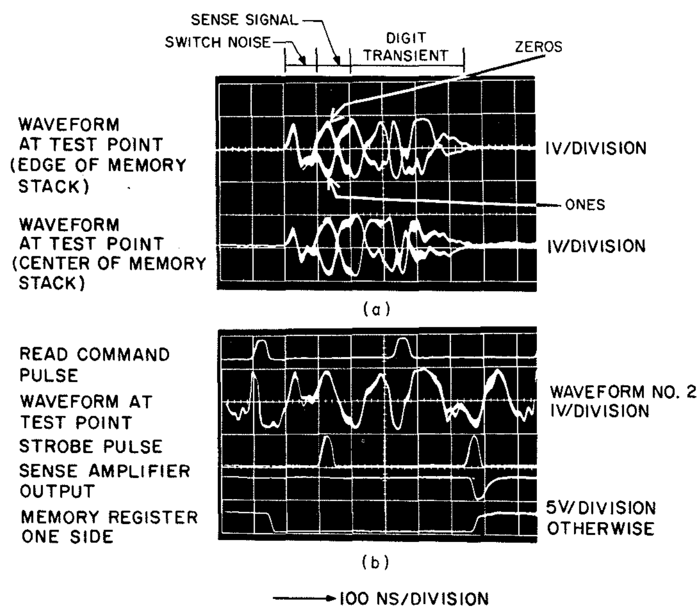


Fig. 11—Waveforms at sense amplifier test points: a) regeneration of *1*'s and *0*'s, and b) regeneration of *1*'s and *0*'s at 450-nsec cycle time.

the plateau were caused by the flow of read and write currents through the switch circuit. These undulations would have been much larger if the switches had not been mounted on the memory stack. Read and write pulses of both timing A and timing B are shown in the figure. Note that the read and the write pulses of timing A are close together, while those of timing B are slightly separated. The digit pulse was observed at the end of a digit line.

Fig. 11 shows related waveforms at a sense amplifier test point. Fig. 11(a) shows two bits, one at the edge of the memory stack and the other at the center, regenerating *1's* and *0's* alternately over the entire memory of 1024 words. Note that the sense signals are delayed to avoid the switch noise. The switch noise, although comparable in amplitude to the sense signal, is minimized through the use of coupling transformers. The negative-going sense signal represents a *1* and the positive-going sense signal a *0*. The digit transient takes about 350 nanoseconds to decay, measured from the start of the digit pulse. This time includes approximately 300 nanoseconds attributed to the base width of the digit pulse and the stock recovery time, plus 50 nanoseconds attributed to the sense amplifier. This relatively slow recovery of the stack, even with the elaborate T termination, appears to result from the imperfection of digit lines as transmission lines.

Fig. 11(b) shows operation at a repetion rate of about 450 nanoseconds cycle time. It depicts the regeneration of *1's* and *0's* on alternate words over the entire memory. Information is available at the memory register within about 230 nanoseconds from the beginning of the read command pulse. Note that switch noise and digit transient recovery are made concurrently without affecting the sense signals. The waveforms shown represent only a small portion of the tests performed on the memory with the aid of the memory exerciser.

Proper digit line termination was actually more important than expected. In the beginning, digit lines were terminated for the differential mode only, using one resistor per line connected directly to the ground. This method was used because it was felt that the T termination, which uses three resistors per digit line pair, was too complex to be practical. However, as the result of sustained common-mode propagations caused by digit drive, the best cycle time achieved was about 700 nanoseconds. The reduction of cycle time to 450 nanoseconds was made possible by changing to the T termination.

The use of ground planes in the memory stack did not cause ringing problems, in spite of a general belief that it would. The ground planes are connected to the memory plane supporting structure on the four corners (Fig. 5). However, the planes are not used as return paths for digit currents. Such paths are provided by a low-impedance common return outside the memory stack. It is believed that ground planes are useful in bringing common-mode propagation under control. (The differential mode, by its nature, does not require ground planes.) In the particular memory stack design used, the ground plane also provides the necessary isolation between the digit lines on the top surface of a memory plane substrate and those on the bottom surface.

The independence of bits in this memory is excellent. As observed at the sense amplifier test point, there were no changes in the sense signal waveforms regardless of signals propagating on other digit lines. The digit transient waveform showed some interaction, but the recovery time was unchanged.

CONCLUSIONS

Development of a word-organized, two-core-per-bit ferrite memory with a short cycle time requires careful consideration of the various transient conditions in the memory system. These transients produce noise that can interfere with reliable detection of the readout signals. In the memory described here, the use of transformers to couple read and write pulses to individual word lines has proved very successful in alleviating the noise problem associated with the word selection matrix. Transformers are used for capacitive decoupling of transients in the word selection matrix from the memory stack. There is a possibility of reducing this noise further by reducing the transformer interwinding capacitance.

Noise was further reduced by placing the voltage switch in the word selection matrix near the memory stack. This provided a low-impedance path to ground for the large word currents and reduced the noise injected during the read pulse via the half-selected word lines controlled by a common matrix switch.

In the digit system, consideration was given to the control of wave propagation along the digit lines to minimize digit recovery time. The basic requirements for control are:

1) Use of two neighboring digit lines as a pair for one-bit location.
2) Equalization of coupling between the digit lines.
3) Use of differential sense amplifiers.
4) Termination of digit lines on both ends for all existing wave propagations, with particular emphasis on the differential mode termination.

The use of the ferrite core strips with plated conductors provided a path for heat conduction, and core heating effects were negligible at high repetition rates. The uniformity of core signals was good, and attenuation through the stack was insignificant.

Two-core-per-bit operation provided bipolar sense signals, and tight control of sense amplifier gain was not required. A delay line in the differential amplifier minimized the problems of DC imbalance and level shifting when sensing the small sense signals in an environment of large unipolar digit pulses. The use of a tunnel diode strobe circuit provided low-level thresholding and high-speed operation.

ACKNOWLEDGEMENTS

The authors wish to acknowledge the assistance given on this project by the members of the Computer Advanced Product Research Group in designing the two-level logic modules, developing the new packaging techniques, and designing and constructing the memory exerciser. Special credit is due to Mr. H. C. Nichols for his invaluable contribution in the fabrication of the memory system. The Memory Products Department, RCA Electronic Components and Devices, constructed the memory stack.

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RADIATION TEST TECHNIQUES

RCA has designed and fabricated prototype models of a Preflight and Operational Status Test Set (PTS) for checking out avionic equipment in U.S. Army light aircraft. This PTS had to meet the following operational requirements: 1) use radiating techniques to reduce or eliminate complications of test connections to the aircraft equipments, and 2) permit testing of all electronic systems without removing the equipment from the aircraft. This paper describes the results of this program, including the extensive field evaluation period under actual and simulated operating conditions.

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Fig. 1—Typical test setup.

The Preflight and Operational Status Test Set (PTS) acts as a ground receiver/transmitter station and radiates appropriate test signals to, and receives response signals from, the aircraft electronics equipment under test. The PTS checks communications, navigation, and instrument landing system (ILS) equipments in the frequency range from 100 kHz (kc/s) to 400 MHz (Mc/s), and identification friend or foe equipment (IFF) at transmit/receiver frequencies of 1030/1091 MHz.

FIELD EVALUATION OF PTS RADIATING TECHNIQUES

A typical test setup for checking the electronic complement of an aircraft by radiation techniques is shown in Fig. 1. The effectiveness of this technique, i.e., the accuracy with which PTS can make measurements, is influenced by the variables in different aircraft configurations and test conditions. An important feature of PTS is that it can be programmed to eliminate or minimize the effects of variables observed during the initial field trials with the developmental models. Programming for these variables can reduce the test tolerance that must be taken into consideration when testing a specific unit under test (UUT). However, programming requirements can be established only by a comprehensive field test program. The purposes of the test program were to evaluate the PTS system and antenna modifications resulting from earlier tests and to determine the radiating test requirements for improvement of the design and operational use of PTS.

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Two basic types of tests were performed: one to determine the optimum position of PTS with respect to an aircraft, and another to collect data on the effect of the variables when PTS was operated within specified location tolerances from the optimum position. The basic measurement was the PTS signal output required to induce a 1-microvolt output at the aircraft antenna.

The field test program was performed using a representative sample of Army helicopters and fixed-wing aircraft under typical operating and environmental conditions. Table I lists the aircraft and avionic systems for which test data were obtained. All data were considered for the evaluation but only the more significant measurements were included here.

The significant result of the field test program is the establishment of system test tolerances when the PTS is programmed for each of the variables. The evaluation of the PTS performance is projected further to include analyses of error probabilities (in terms of transceiver power/sensitivity) and transceiver intelligibility (in terms of signal plus noise to noise ratio) which are combined to operate the *go/marginal/no-go* test result indicator. Included also with these analyses are trade-offs made possible by consideration of the avionic equipment mission profile.

PROCEDURE AND TEST DATA

System test tolerance data were obtained with the PTS positioned 50 feet from the aircraft at an orientation of 315° (i.e., at 45° in the forward port quadrant) with its antenna perpendicular to the aircraft antennas. Readings were taken at different positions within prescribed position tolerances (± 5 feet on radial distance and ± 10° on angular positions) and averaged. The PTS was set up specifically for each avionic equipment, and the measurement was the PTS signal output required to induce 1-microvolt at the aircraft antenna. Significant variables

TABLE I—Aircraft and Avionic Systems for which Test Data Were Obtained

Aircraft \ Avionics	ARN-30	ARN-12	ARN-6	ARN-59	ARN-32	ARC-44	ARC-55	ARC-73	ARC-60	ARC-59	ARC-12 T-366	R-511	APX-44	R-746	R1041	ARA-31
OH-13G						X			X							
U-1A	X	X	X			X	X	X		X		X	X			
O-1A						X			X			X				
U-6A	X			X	X	X	X				X			X		X
OV-1C	X			X		X	X	X					X	X	X	
CV-2B	X			X		X	X				X					
CH-34C	X			X	X	X	X	X		X						X
OV-1B	X			X		X	X	X					X	X	X	
U-8F	X					X	X			X	X			X		

TABLE II—System Test Tolerance as a Function of Programming Variables

System Test Tolerance \ Programming Variables	UUT	Aircraft Type	Test Frequency	Steel-Mat Ground Conditions	Concrete Ground Conditions	Precise Positioning of PTS
±17 dB	X					
±13 dB	X	X				
± 9 dB	X	X	X			
± 8 dB	X	X	X	X		
±4.6 dB	X	X	X	X	X	
± 3 dB	X	X	X	X	X	X

DAVID GOLDBERG received his BSEE in 1949 and a BS degree in Radio Engineering in 1950 from Indiana Technical College. During 11 years with RCA, he has participated in the design and development of transistorized circuits for audio, servo, analog, and digital applications. He has been responsible for hardware design and fabrication on the following programs: APCHE, Program 706, Multisystem Test Equipment, and Preflight Test Set and Mohawk Adapter.

for aircraft/avionic configuration were:

1) Antenna configuration versus avionic equipment.
2) Operating frequency for avionic equipment.

The significant variables for ground conditions were:

1) Wet and dry ground in a cleared area (control condition).
2) Wet and dry reinforced concrete.
3) Wet and dry steel mats.

Average readings for the above-ground conditions are shown in Fig. 2. The spread of these data indicates that when the PTS is set up to test a given avionic equipment, the test tolerance is ± 17 dB. As the various conditions are taken into account, and programmed for, the spread rapidly reduces.

Examination of the spread in signal output for similar UUT's using different antennas reveals that antenna configurations have a significant effect. For example, addition of the APS-94 radar antenna in the OV-1B Mohawk reduces the strength of the received PTS signal output by 10 dB when testing the ARC-44 at 49.9 MHz, as compared with the signal required when the ARC-44 operates in the OV-1C Mohawk. The configuration of a helicopter, with its blades, causes a change in reception compared to the configuration of an airplane.

Similarly, UUT reception and transmission will vary subject to the antenna used. An ARC-44 when used with a whip antenna will operate over longer distances, and receive better than when it is used with a long wire antenna.

Distance from, and angular position with respect to, the aircraft affects the transmit/receive tolerance, e.g.:

1) Radial distance from the aircraft (50 ± 5 feet): ± 2 dB.
2) Radial position to the aircraft (315° ± 10°): ± 1 dB.

In addition, a large variation in the received signal is caused by changing frequency when testing the same aircraft antenna under the same test conditions. Part of this change is due to the radiation path loss. Much of the remainder is the effect of cancellation and addition of the signal caused by aircraft reflections and antenna characteristics. Similarly, minor effects are caused by wings, struts, and other aircraft structures.

By programming the test set to compensate for variances caused by aircraft and operating frequency, the test data given in Fig. 2 can be normalized and plotted as shown in Fig. 3A.

Another major contributor to the system test tolerance is the effect of ground conditions. The data plotted in Fig. 3A show that the distribution of the test conditions is almost completely random except the steel-mat condition. Data for the steel-mat condition are generally lower than the other test results. Because of the highly conductive ground plane of the steel mat, less PTS signal output is required to induce 1 microvolt at the antenna terminals. Also, note that the steel-mat effect is negligible at the lower frequencies, i.e., below 20 MHz.

Normalizing the data to program for ground conditions results in Fig. 3B plot.

The effects of the major variables on system test tolerance become a function of programming requirements. The testing tolerance can be reduced from ± 17 dB (Fig. 2) to ± 3 dB, as depicted in Table II.

Fig. 2—Field data plot of average PTS signal output for avionic equipment in aircraft. (All frequencies shown are in MHz.)

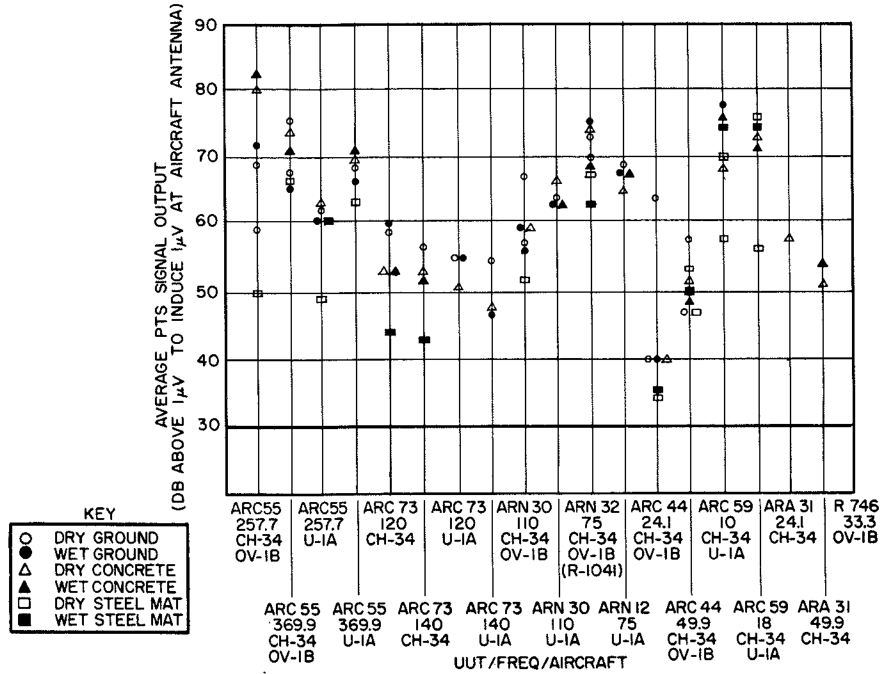
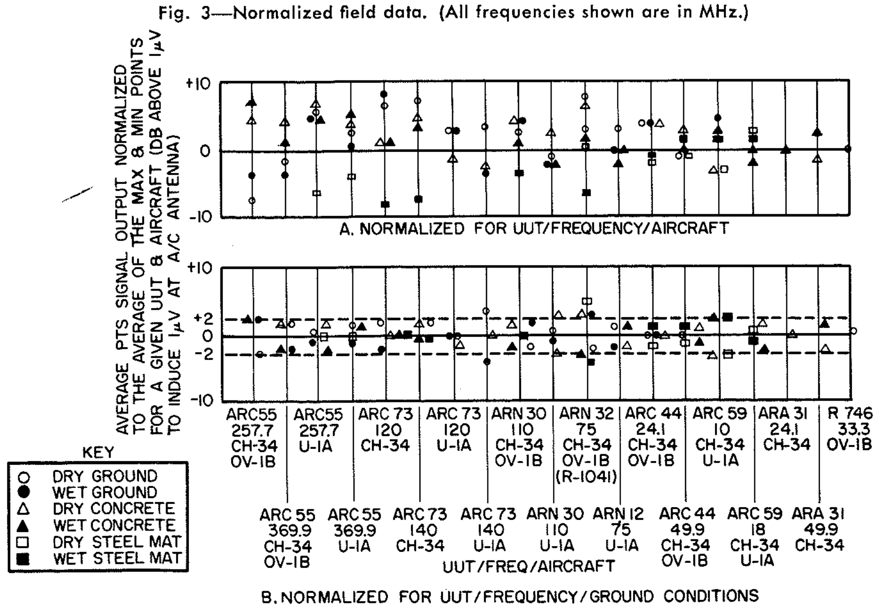


Fig. 3—Normalized field data. (All frequencies shown are in MHz.)



TESTING ACCURACY

The signal plus noise to noise ratio, $\frac{(S+N)}{N}$, is an accurate measurement of receiver performance. The intent of field checkout is not to measure the $\frac{(S+N)}{N}$ ratio but to determine whether the UUT is performing correctly, i.e., whether the $\frac{(S+N)}{N}$ ratio is above or below a required value. The accuracy of this measurement is important; an accuracy to $\pm$ 1 dB is readily attainable.

CHECKOUT ERROR PROBABILITIES

At the time of issue, or after bench tests, a UUT meets the minimum power sensitivity level specified in the technical manual and generally has a range capability in excess of the normal mission requirement. This fact suggests a method for establishing test limit criteria based on the probability that among all the UUT's being tested, a unit would be accepted when it is faulty, or vice versa.

An error analysis was made by assuming the system test tolerance of $\pm$ 3 dB or $\pm$ 1.5 dB, a *good* decision limit being 6 dB below the nominal value of the UUT, and a *bad* decision limit being 3 dB farther away. These limits result in a *marginally good* area of 3 dB. The equations, plus curves, of the analysis are given in the Appendix. As shown in Table III, the probability of judging a bad unit good or a good unit bad is no worse than 0.28 percent.

CONCLUSIONS

The PTS program demonstrated that the use of radiating techniques can eliminate connections to aircraft avionics. The prototype models of the PTS that were designed and built, and the field evaluations that were conducted under various conditions verified that design goals can be met. Operational checkout by PTS of a UUT provides a probability of error of less than 0.8 percent, thus raising pilot confidence in equipment prior to departure on any mission.

Although the PTS has not been adopted for Army use, significant conclusions based on data obtained during the field evaluation are summarized here:

1) Radiation testing is feasible on the flight line under various operating environments, i.e., steel mat, reinforced concrete apron, bare ground, in or near hangars, in the vicinity of other aircraft, and under wet or dry conditions.
2) The overall system test accuracy that can be achieved can vary from $\pm$ 3 dB to $\pm$ 17 dB, depending on the degree of programming used. The variables that can be programmed include type of UUT, type of aircraft, frequency attenuation, type of ground cover, (i.e., steel mat, concrete, bare ground), and ground condition (i.e., wet or dry).
3) Variations in test tolerances caused by vertically or horizontally polarized antennas are negligible.
4) Using radiating techniques, the UUT can be checked out operationally with less than an 0.8-percent probability of error.

Special antennas are not required in the frequency range covered. Precise analyses indicated that idiosyncrasies of near field testing (distances $<5\lambda$) might make test results meaningless. As a result, the data taken in testing the ADF equipment (ARN-6, ARN-54, ARN-59, R-511) with a loop antenna were reviewed carefully. The results indicate a consistency that gives confidence in the equipment configuration.

APPENDIX
Probability Density Distribution

When avionic equipments are bench-tested to check operating performance according to the power/sensitivity ratings prescribed by the applicable technical manual, they must meet or exceed a minimal operating level. This fact leads one to certain basic assumptions for establishing probability density distributions for the equipment operating level:

1) The average operating level of all equipments of the same type at the time of installation in the aircraft exceeds the minimum acceptable level.
2) The operating levels of all equipments of the same type can be expected to be statistically distributed according to a Gaussian probability distribution function.

When the equipment power/sensitivity is tested during checkout, an input stimulus is applied and the output is measured. Gaussian errors in the input stimulus and the output measurement lead to a Gaussian error in the determination of *true* equipment operating level. The Gaussian probability density distributions for 1) the true equipment operating level, and 2) the error in determining the operating level, suggest a bivariate problem in the determination of the percentage of erroneous checkouts as a function of the test (or decision) limit.

The probability density distribution, *f(T)*, for the true equipment operating levels is shown in Fig. 4. Value *S* is the specification limit for operating levels that are unacceptable (*no-go*). Value *D* is the decision limit for operating levels that are acceptable (*go*). The region between *S* and *D* represents marginally

Fig. 4—Probability density distribution for true equipment operating levels.

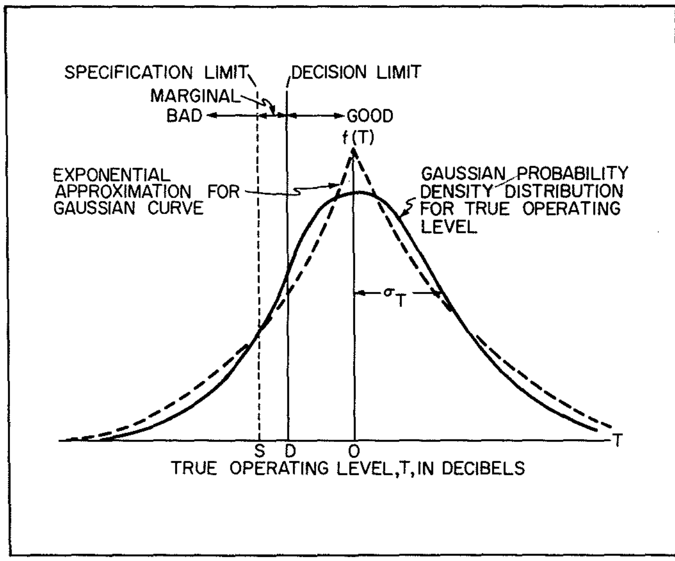
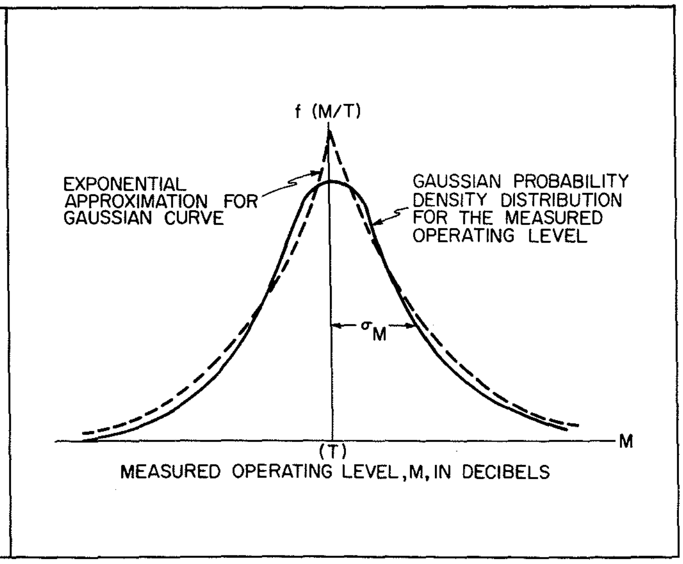


Fig. 5—Probability density distribution for error in determining operating levels.



good operating levels. The probability density distribution, $f(M/T)$, for the error in determining the operating level is shown in Fig. 5.

In Figs. 4 and 5, the Gaussian curves have been approximated by back-to-back exponential curves to get a closed-form answer to the bivariate problem. The equations for the approximations are:

$$
\begin{aligned}
f(T) &= \frac{1.15}{2\sigma_T} \exp\left[1.15\left(\frac{T}{\sigma^m}\right)\right] \quad \text{for } -\propto < T \leq 0 \\
&= \frac{1.15}{2\sigma_T} \exp\left[-1.15\left(\frac{T}{\sigma^m}\right)\right] \quad \text{for } 0 < T < \propto
\end{aligned} \tag{1}
$$

where σ_T = standard deviation of the true operating level distribution.

$$
\begin{aligned}
f(M/T) &= \frac{1.15}{2\sigma_M} \exp\left[1.15\left(\frac{M-T}{\sigma_M}\right)\right] \quad \text{for } -\propto < M \leq T \\
&= \frac{1.15}{2\sigma_M} \exp\left[-1.15\left(\frac{M-T}{\sigma_M}\right)\right] \quad \text{for } T < M < \propto
\end{aligned} \tag{2}
$$

where σ_M = standard deviation of the error in determining the operating level.

Error Probabilities

Three types of errors in determining operating levels were considered:

1) The probability of calling a bad UUT good, i.e., the probability of an undetected defect, P_{ud}.
2) The probability of calling a bad UUT marginally good, i.e., probability of a marginal undetected defect, P_{mud}.
3) The probability of calling a good UUT bad, i.e., the probability of a false alarm, P_{fa}.

An undetected defect may result in sending an aircraft on a mission with defective equipment that should have been repaired. This type of error is the most serious. A marginal undetected defect may have similar results, but the fact that the measurement showed the equipment to be marginal rather than bad is a safeguard. The effect of a false alarm may mean needless expenditure of time, money, and spare parts while a mission is delayed for unnecessary maintenance. Calculations for various error probabilities are given in Fig. 6.

Numerical Example

Equations 3, 4, and 5 are solved in terms of parameters based on values specified in the technical manual for the AN/ARC-55 Communications Transmitter/Receiver. For this example, three sets of values for S, D, σ_T, and σ_M are postulated in Table III to bracket anticipated field conditions. In this case, Fig. 4 depicts a distribution of UUT's where the nominal value is either 9 dB or 7.5 dB greater than the technical manual stated value. Fig. 5 depicts a system test tolerance, of measured value, of ± 3 dB (see Table II). Table III gives for each of these sets of postulated values the corresponding calculated probabilities of undetected defect, marginal undetected defect, and false alarm.

TABLE III—Calculated Error Probabilities for Sensitivity Measurements on AN/ARC-55

Parametric Values				Error Probabilities		
S	D	σ_T	σ_M	P_{ud}	P_{mud}	P_{fa}
−9 dB	−6 db	6 dB	1 dB	0.00020	0.00616	0.00050
−7.5 dB	−6 dB	6 dB	1 dB	0.00151	0.00699	0.00280
−7.5 dB	−4.5 dB	6 dB	1 dB	0.000266	0.00810	0.00066

From the error probabilities shown in Table III, when testing for sensitivity, PTS would judge:

1) 0.02 to 0.151 percent as good when they are really bad.
2) 0.616 to 0.81 percent as marginally good when they are really bad.
3) 0.05 to 0.28 percent as bad when they are really good.

Similar calculations of error probabilities could be made for the other types of UUT's that can be checked out by PTS.

Fig. 6—Error probability calculations for: a) undetected defect, b) marginal undetected defect, and c) false alarm defect.

(a) Probability of an undetected defect:

$$
P_{ud} = \int_{-\propto}^{S} f(T)\left[\int_{D}^{\propto} f(M/T)\,dM\right]dT = \int_{-\propto}^{S} \frac{1.15}{2\sigma_T}\exp\left[1.15\left(\frac{T}{\sigma_T}\right)\right]\left\{\int_{D}^{\propto}\frac{1.15}{2\sigma_M}\exp\left[-1.15\left(\frac{M-T}{\sigma_M}\right)\right]dM\right\}dT \tag{3}
$$

$$
= \left[\frac{\sigma_M}{4(\sigma_T+\sigma_M)}\right]\exp\left[1.15\left(\frac{S\sigma_T - D\sigma_T + S\sigma_M}{\sigma_M\sigma_T}\right)\right]
$$

(b) Probability of a marginal undetected defect:

$$
P_{mud} = \int_{-\propto}^{S} f(T)\left[\int_{S}^{D} f(M/T)\,dM\right]dT = \int_{-\propto}^{S} \frac{1.15}{2\sigma_T}\exp\left[1.15\left(\frac{T}{\sigma_T}\right)\right]\left\{\int_{S}^{D}\frac{1.15}{2\sigma_M}\exp\left[-1.15\left(\frac{M-T}{\sigma_M}\right)\right]\,\mathrm{dM}\right\}\mathrm{dT} \tag{4}
$$

$$
= \left[\frac{\sigma_M}{4(\sigma_T+\sigma_M)}\right]\exp\left[1.15\left(\frac{S}{\sigma_T}\right)\right] - \exp\left[1.15\left(\frac{S\sigma_T - D\sigma_T + S\sigma_M}{\sigma_T\,\sigma_M}\right)\right]
$$

(c) Probability of a false alarm defect:

$$
P_{fa} = \int_{D}^{\propto} f(T)\left[\int_{-\propto}^{S} f(M/T)\,dM\right]dT = \int_{D}^{0}\frac{1.15}{2\sigma_T}\exp\left[1.15\left(\frac{T}{\sigma_T}\right)\right]\left\{\int_{-\propto}^{S}\frac{1.15}{2\sigma_M}\exp\left[1.15\left(\frac{M-T}{\sigma_M}\right)\right]dM\right\}dT \tag{5}
$$

$$
+ \int_{0}^{\propto}\frac{1.15}{2\sigma_T}\exp\left[-1.15\left(\frac{T}{\sigma_T}\right)\right]\left\{\int_{-\propto}^{S}\frac{1.15}{2\sigma_M}\exp\left[1.15\left(\frac{M-T}{\sigma_M}\right)\right]dM\right\}dT
$$

$$
= \left[\frac{\sigma_M}{4(\sigma_T-\sigma_M)}\right]\exp\left[1.15\left(\frac{S\sigma_T - D\sigma_T + D\sigma_M}{\sigma_T\,\sigma_M}\right)\right] - \left[\frac{\sigma_T\,\sigma_M}{2\,(\sigma_T^2-\sigma_M^2)}\right]\exp\left[1.15\left(\frac{S}{\sigma_M}\right)\right]
$$

A SCAN CONVERTER FOR APOLLO TELEVISION

Most television camera systems for spacecraft have been designed to obtain scientific data. The Apollo television pictures will have a different purpose; as they are received from the moon, they will be retransmitted to the American public over home TV channels. Because of stringent bandwidth requirements, Apollo TV pictures will be at a different scan rate than commercial TV pictures. Therefore, these pictures must be changed from one set of standards to another by a ground-station scan converter. This paper reviews underlying considerations of TV scan conversion along with some past approaches. It also presents NASA's approach to the problem and describes the equipment built by RCA to implement this approach.

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THE objective of the Apollo program is to land a man on the moon. Included in this mission is a TV camera system that will enable the American public to watch the epochal proceedings on their home TV sets. Although TV camera systems for spacecraft are satisfactory for obtaining scientific data, special requirements must be met to supply TV pictures for commercial broadcasting. In addition to providing scientific data, the Apollo TV system is designed to produce a broadcast picture of acceptable quality which can be converted to broadcast standards by a ground-station scan converter.

APOLLO TV SYSTEM

Because of the great distance from moon to earth and the high cost of providing transmitter power from the moon, the Apollo TV system relies on different scanning standards than broadcast television. A bandwidth of only 500 kHz (kc/s) is employed, compared to 4.5MHz (Mc/s) of broadcast television. This reduction of nine to one in transmitter power, due to the bandwidth reduction, is extremely important because of the high cost of power consumption in a space vehicle. Of course, if the bandwith is reduced, then the scan (frame and line) rates have to be reduced. For the Apollo TV system, a noninterlaced frame rate of 10 frames per second was chosen. This rate produces some erratic motion when objects move rapidly, but the resultant picture is still quite usable. By using a frame rate which is an integral submultiple of the 30 frames-per-second rate of commercial TV, scan conversion systems are easier to design. The use of 320 active lines yields a TV picture with 220-line resolution both horizontally and vertically compared to a median value of approximately 330 lines of resolution for broadcast television. An additional requirement of the Apollo TV system is that it must transmit high-definition *still* photographs for scientific purposes. To accomplish this within the 500kHz bandwidth requirement, these photographs are transmitted at a rate of one frame per 1.6 seconds, allowing 1280 lines per frame.

SCAN CONVERSION TECHNIQUES

The TV signal received from the moon must be converted from the Apollo scan standards to broadcast scan standards for proper display on a commercial receiver. Over the past several years, a number of scan conversion systems have been investigated. The earliest method consisted of a broadcast-rate vidicon camera pointed at an Apollo-rate monitor. The vidicon target acted as the storage medium (all scan conversion schemes require storage capabilities). Three broadcast frames were read out for each Apollo picture laid down; however, this system had a serious defect. Vidicon target material with low lag properties has a very destructive readout in that very little signal can be obtained on an additional readout unless more light has been imaged onto the target. Vidicon target material with less destructive readout properties has high lag. It was impossible to get three frames of video reasonably alike (which is needed to avoid the flicker that would result if the frames were different), without also getting a large amount of stickiness and motion smearing. No usable compromise of these two target parameters could be obtained.

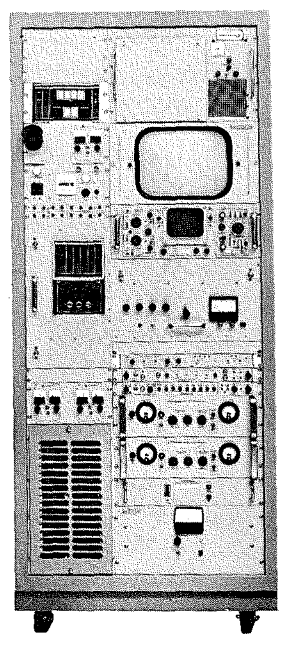
Another approach utilized storage tubes. Because of the prepare-write-read cycle requirements of the storage tubes, three tubes were required with one tube being in each of the above modes at any one time. There are two problems with this arrangement: 1) the beams in the three storage tubes are very difficult to register, and 2) shading and transfer characteristic differences produce a 3⅓-cycle-per-second flicker.

After studying the problems associated with the above approaches, NASA decided upon the approach described below. This approach is basically sound and promises to be acceptable for converting other slow-scan TV systems to broadcast standards. A photograph of the complete scan-converter equipment is shown in Fig. 1.

BASIC SCAN CONVERTER SYSTEM

A block diagram of the system with signal waveform points is shown in Fig. 2; video waveforms are shown in Fig. 3. The Apollo TV signal is applied to the Apollo monitor, and the picture formed on the monitor is focused onto a broadcast vidicon camera. In effect, the vidicon target stores the Apollo signal. For every third broadcast frame, which is once during each Apollo frame, the camera reads out one frame of video signal at broadcast rates. For the next two frames, the scanning beam of the camera is gated *off* and no video is produced. The video from the camera is an interrupted video: one frame *on*, two frames *off*. In addition, the two video fields comprising the video frame have different video levels because the readout of the first field reads out a portion of the second field. In broadcast TV,

Fig. 1—Photo of the scan converter equipment.



Final manuscript received February 7, 1966.

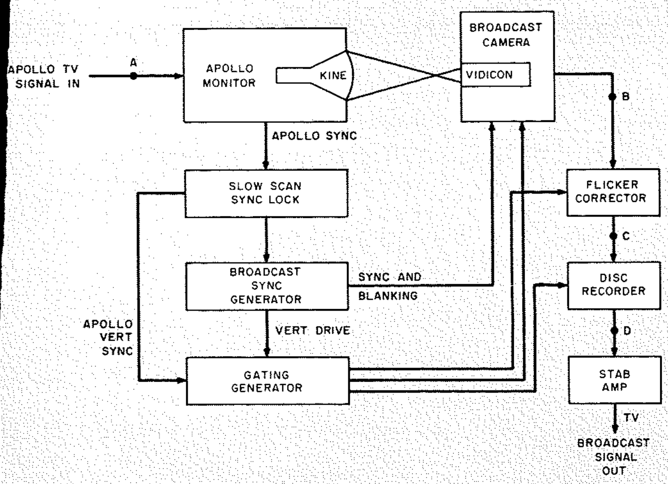


Fig. 2—System block diagram with signal waveform points.

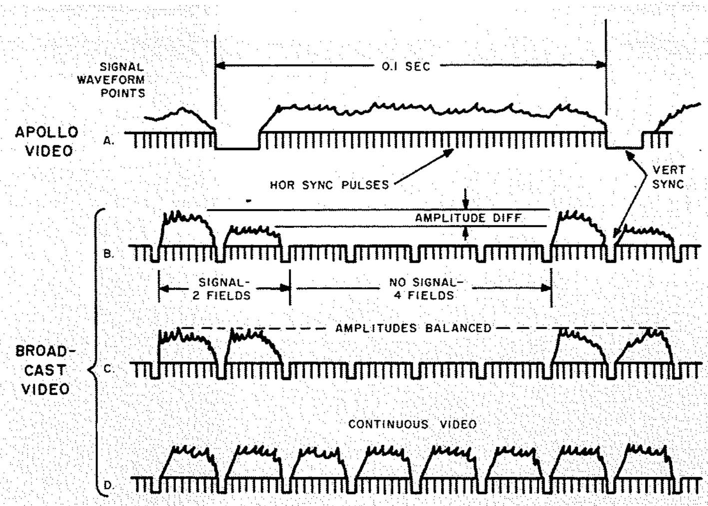


Fig. 3—Video waveforms.

each frame is composed of two interlaced fields in order to increase the number of pictures from 30 to 60 per second and thus eliminate flicker without increasing system bandwidth requirements. This technique is unnecessary in the Apollo system since the system must be scan converted to broadcast standards before transmission over the commercial TV networks. To equalize the video levels in the two fields comprising one broadcast frame, the video signal is fed to the flicker corrector, which is an electronically controlled, variable-gain amplifier. The interrupted video is fed next to a magnetic disc recorder, where the interrupted video is converted to continuous video. Each active frame is recorded on the disc, read out three times to fill in the off frames, and then erased to make room for the next active frame. From the magnetic disc recorder the video is fed to a conventional stabilizing amplifier, where the sync pulses are cleaned up. At this point the video is ready for transmission over the commercial broadcast networks.

The slow-scan sync lock keeps the vertical sweeps synchronized, and a sync generator controls the vidicon camera. A gating generator supplies pulses to: 1) turn the vidicon beam *on* for desired broadcast frame readout, 2) control the timing of the gain change in the flicker corrector to equalize the video levels in the two fields comprising the one active frame readout, and 3) control the disc recorder so that it will record only when the camera is putting out useful video.

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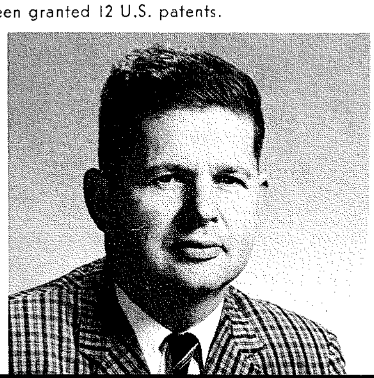

STORAGE TIMING

Storage timing is a critical phase of the conversion process. The video signal is stored at two different places in the scan converter: on the photoconductor target of the vidicon, and on the magnetic disc recorder. The first storage process converts one frame at Apollo scan rates to one frame at broadcast scan rates; the second storage process repeats each of the *on* broadcast frames three times to produce continuous broadcast video.

The relative timing between the vertical deflections on the Apollo monitor and on the vidicon camera is shown in Fig. 4. The beam on the monitor scans downward, *painting* a picture onto the vidicon. This picture is stored as a two-dimensional charge pattern on the photoconductor. When the Apollo beam is one-sixtieth of a second (corresponding exactly to one broadcast field) from the bottom of the picture, the vidicon beam is gated *on*, and the vidicon reads off the first field of a frame. The vidicon beam catches up with the monitor beam during vertical retrace, when both beams are blanked *off*. Since the vidicon beam moves faster, it can read off the second field of the same frame while the monitor beam follows behind painting the next frame onto the vidicon. In this manner, the beams never interfere with each other and the vidicon can read out both fields. If the vertical deflections were not locked together by the sync lock, a bright horizontal stripe would appear as the vidicon beam passed the monitor beam.

The timing on the disc recorder is straightforward. The disc rotates at 1800 r/min, locked to the broadcast video. One complete frame can be stored on one revolution of the disc. When a frame of video arrives at the recorder, the previously recorded frame is erased, and the new frame is recorded on the disc using a record head. At the same time, a read head, diametrically opposite from the record head, is reading out the signal continuously; hence, the read head does not need gating signals. However, as described above, the record head and the vidicon control grid require gating signals from the gating generator, and these are locked to the incoming Apollo vertical sync.

SINGLE FIELD OPERATION

The operating mode described above depends upon the ability of the flicker corrector to equalize the two active interlaced fields to reduce flicker. There is a second operating mode which exhibits no flicker. In this mode, only the first of the two active fields is recorded and this same field is played back five additional times, using both heads for playback by switching, alternating from one to the other. However, by using only one field of the two available, the vertical resolution has been cut in half. At present, both operating modes are being investigated; the trade off between the flicker in the one mode and loss of vertical resolution in the other mode is yet to be finalized.

MONITOR

The monitor, built by RCA specifically for the scan converter, employs a 5CEP11 flying-spot tube. An anastigmatic coil reduces line structure without affecting horizontal resolution. As previously mentioned, the Apollo TV system has two scanning rates: 10 frames per second with 320 lines per frame, and 1 frame per 1.6 seconds with 1280 lines per frame. The deflection circuits were specially designed to maintain size, centering, and linearity when switched be-

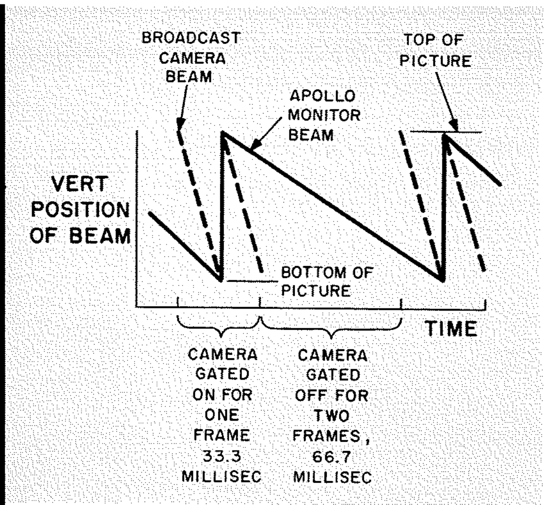

Fig. 4—Vertical deflection and camera beam timing.

tween the two scanning rates: a 4:1 horizontal rate change, and a 16:1 vertical rate change.

The 10-frame-per-second Apollo TV system uses either a conventional pulse-type sync or a tone-burst sync at 409.6kHz. The high-definition *still* system uses the tone burst for its sync. The purpose of the tone burst is to reduce the transmitter power required. Since the tone burst occupies the same frequency band and gray-level domain as the video, sync separation is more difficult. The sync separator for the pulse-type sync uses a conventional amplitude-type sync stripper; the separator for the tone-burst sync employs a crystal oscillator that is locked to the incoming 409.6kHz bursts. This frequency is counted down to generate the Apollo horizontal timing pulses. Proper phasing is obtained by comparing the counted-down pulses to an amplitude-detected tone burst and feeding error signals back to the count-down chain. The amplitude-detected tone burst is jittery and not immune to noise, but by building inertia into the pulse comparison circuitry, the amplitude-detected tone burst becomes a reliable reference. The amplitude-detected tone burst is also used to gate the tone burst to the crystal oscillator.

VIDICON CAMERA

The vidicon camera is an RCA TK-22 film model, modified to accept an external vidicon gating pulse. It is also modified by the addition of gamma correction circuitry that is variable so an operator can place the *break points* in the transfer characteristic wherever desired. This feature is necessary because the scan converter should be as linear as possible so as not to compress or expand either the black end or the white end of the transfer characteristic. The transfer characteristics of kinescopes and vidicons tend to complement each other, but the match is far from perfect.

When the vidicon reads out the first field, it reads out a little of the second field also, due to the finite size of the scanning aperture. To minimize this condition, a larger-than-normal vidicon (type 8480) is employed. This vidicon has a diameter of $1\frac{1}{2}$ inches and a limiting resolution greater than 1000 TV lines in any part of the picture. A high resolution implies a small-diameter scanning aperture.

FLICKER CORRECTOR

If the video level in the second field differs from the level in the first field by more than approximately 5 percent of the peak highlights, an objectionable flicker will result. The flicker corrector varies the gain of the second field with respect to the first field so that the video levels of the two fields will be the same. This gain variation requires a timing pulse from the gating generator that differs from the gating pulse fed to the camera and to the disc recorder. The variable gain is obtained by controlling the drain resistance of a field-effect transistor in a resistive voltage-divider arrangement. The drain resistance is varied by appropriate settings of the gate voltage. By restricting the drain voltage (DC + video) to plus or minus several tenths of a volt, the field-effect transistor is made to act as a linear resistor.

DISC RECORDER

The disc recorder is similar to those used for instant replay in televising sporting events, but the design is modified considerably; the logic timing for the record head was revised for this application. The disc, about 12 inches in diameter, rotates at 1800 r/min and is servo-controlled for long-term stability of $\pm$ 0.2 μs; short-term stability is much better because of the inertia in the metal disc. The record head is on one side of the disc near the edge, and the read head is approximately 180° away. The heads can be adjusted radially to a new track if the disc shows signs of wear. The disc recorder was designed for this application by MVR Inc.

PULSE AND TIMING UNITS

The sync generator, sync lock, and gating generator are solid-state units employing mostly digital logic to derive the desired timing pulses. The sync generator is a standard unit, but the sync lock was designed especially for this application by Telemet Co. The gating generator was made by RCA especially for this system. There have been several modifications in the scan converter system and its components since NASA built the breadboard feasibility model. To satisfy the requirement for high-definition *still* photographs, the sync lock was extensively modified to control the sync generator in this mode; the sync lock does not produce a gating pulse in this mode. The addition of the flicker corrector to the system and modifications to the disc recorder by MVR Inc. resulted in a requirement for three separate gating pulses. For these reasons, a separate gating generator is now required, where, originally, the sync lock supplied the necessary gating pulse.

PERFORMANCE

The 10-frames-per-second readout by the Apollo live pickup vidicon yields slightly more lag and motion jerkiness than normal broadcast TV. The scan converter cannot improve this; it can only convert what it gets. Since all photoconductors have some lag, the scan converter actually increases the lag a little on moving objects, but it does not affect the jerky motion. The magnitude of these effects is fairly subjective and hard to describe quantitatively. Slow motion, however, is completely satisfactory. The photograph in Fig. 5 is representative of the overall performance of the system, from pickup by an Apollo-rate camera, through the scan converter, to display on a broadcast monitor. There has been very little overall picture degradation. The scan converter has good gray-scale fidelity, a signal-to-noise ratio better than 35 dB, and a resolution response down less than 6 dB at 4 MHz.

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The author is grateful for the encouragement given by J. Lowrance, M. Mesner, and M. Sullivan in the design of the equipment and the writing of this paper, and for the invaluable assistance of A. Hasler, C. Heitzenroder, E. Hippe, F. Hubit, P. Murray and L. Saxton in the construction of the equipment. The author also wants to thank H. TePoel, P. Lipoma, and B. Seay of NASA for their assistance in this project.

Fig. 5—Photo taken by the Apollo TV camera channeled through the scan converter.

INTELLECTUALISM AND THE AMERICAN ENGINEER

This paper describes a limited survey made to determine the intellectual attitudes of a group of engineers. Intellectualism is defined, the sample group is described, and the results of the survey are analyzed.

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The term *intellectual* has no precise meaning, as is evidenced by dictionary definitions. Merriam-Webster's New International Dictionary, Third Edition, Unabridged, gives three separate meanings: 1) Endowed with intellect, 2) having the power of understanding, and 3) having the capacity for the higher forms of knowledge or thought. Fowler's *Modern English Usage* says: "Intellectual—one in whom the part played by the mind, as distinguished from the emotions and perceptions, is greater than in the average man." Its meaning is so varied that to some it represents qualities that are to be distrusted (e.g. egghead, long hair), whereas to others it is the highest achievement. To the Philistine, the intellectual obstructs the serious business of living; to others, he is the soul and conscience of the community. This very imprecision of the term requires that it be defined to suit the purposes of the ensuing discussion.

S. M. Lipset[1] defines an intellectual as anyone with a university education; or he who creates, distributes, and applies culture, including science, art, and religion. Such a wide definition leaves out a significant group who are not engaged in any of these pursuits but who are generally *mind-oriented* rather than *thing-oriented.*

Neither all academic men nor all professional men may be called intellectual without qualification. Intellectualism, unlike intelligence, implies a capacity for detachment from immediate experience. Intelligence requires one to limit his attachment to the pragmatic tasks of the moment. Richard Hofstadter has said[2] that intellectuals are concerned with the critical, creative, and contemplative functions of the mind while merely intelligent people seek to group, manipulate, reorder, and adjust. The intellectual feels that he must go beyond the concrete task at hand and be concerned with meanings and values, so that other future concrete tasks may be more readily accomplished. Intellectuals are concerned with "core values of society." They are concerned with moral and ethical aspects of the simplest concrete tasks. Intellectuals are not satisfied with the status quo, but seek constantly to examine and analyze to find a deeper meaning. It is for this reason that intellectuals are disturbing to the ordinary citizen.[3]

In our times we have seen the emergence of a highly educated, extremely intelligent group of people. They can solve the deepest problems, they are able to create, they can seize nature and shake down her secrets, and, if need be, they can destroy the world. As a group, RCA engineers are highly educated, extremely intelligent people who do, in fact, solve some of nature's deepest problems. But are they concerned with more profound truths that have been mentioned, such as the moral and ethical aspects of their professional activities? Do they make the value judgements or the moral judgements necessary in our increasingly complex world? The most difficult problem to solve is to determine whether they have this attitude of intellectualism necessary to make such judgements. How one measures this attitude is a question to which an answer is attempted in the following material.

For the purposes of this paper, the initial approach to defining intellectualism is to call it an *attitude* of the individual in which he is most concerned with the ideas behind things. This concern with ideas makes understanding possible. Consequently, intellectualism is concerned with the *non-thing* interest of man because this will provide him with the data for understanding. To define intellectualism in another fashion: man is a unitary whole or system consisting of mind and matter. This unitary whole or system comes in contact with everything outside the unitary whole—his environment. This environment consists of other systems made up of mind and matter like himself and other systems consisting of simply matter. An understanding of these systems requires intellectual knowledge, as opposed to sense knowledge which merely enables him to know these systems.

A group of questions was constructed to attempt to determine the intellectual attitudes of a sample group of engineers. Admittedly the questions are *loaded* if by loaded one means that they were constructed so that certain kinds of answers could be drawn from them. The author does not believe it possible to construct a question that is not loaded. The key to the questions is not what kind of answers are received (because certain questions lead inevitably to certain answers), but how many persons would give one kind of answer opposed to another kind of answer. The attitude of the persons questioned on matters that relate broadly to the intellectual view of the world can be determined in this way.

DESCRIPTION OF THE SAMPLE

It is estimated that 800,000 to 900,000 individuals are classified as engineers in the various breakdowns of the U.S. labor force. This figure includes the more than 100,000 members of the Institute of Electrical and Electronic Engineers. This organization represents a merger of the American Institute of Electrical Engineers and the Institute of Radio Engineers, and its members are associated with all phases of the

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electrical and electronic industries. These individuals are, therefore, associated with industries and activities that have been identified with one of the most rapidly expanding and developing technologies. They are in the forefront of new uses of engineering sciences in missile, space, advanced communications, and atomic energy applications. To restrict the data to manageable levels, an engineering group at one geographical location of a major manufacturer was selected for this study. This engineering group, consisting of approximately 1200 engineers with a variety of educational backgrounds and various levels of experience, was concerned with applied research, analysis, design, and development associated with highly sophisticated electronic equipment, including data-processing and space communications equipment. This group is compelled to be abreast of the state of the art of the latest technologies and, in many instances, is pushed beyond the state of the art to provide better equipment to satisfy functional needs. Approximately 13% of these engineers were randomly selected and supplied with a questionaire. Ninety-three replies to this questionaire have been received. This sample, 7.8% of the total group of 1200, was considered statistically valid for the initial data compilation for this paper.

ANALYSIS OF THE SURVEY

The group replying to the questions was analyzed to establish some typical features. It was found that the average number of years in the engineering profession was 13. Of this group, 78% identified themselves as electrical/electronics engineers and 18% as mechanical engineers; the remaining 4% included physicists and chemists. Consequently, the tested group can be typified as a group of electronics engineers with 13 years of professional experience.

As has been noted, a series of questions was formulated and presented to the engineers in an endeavor to gain some insight into their intellectual attitudes. These questions were selected to shed light on the following:

1. Predisposition to study engineering (questions 1 and 2)
2. Undergraduate curriculum (questions 3 and 4)
3. Graduate studies (questions 5 and 6)
4. Reading habits (questions 7 and 8)
5. Level of particular knowledge (questions 9 and 10)
6. Attitude toward intellectual life and attainment (questions 11 and 12)

A specific purpose was built into each question. (See Appendix.) The construction of the questions depended on the subjective approach of the questioner and the problem of their interpretation. Others might have worded these questions differently, used different questions, or, perhaps, interpreted the results in a different light. The following analysis is the author's evaluation of this survey.

Questions 1 and 2

1. Did you decide to study engineering because you enjoy taking things apart to see how they work and then attempt to put them together again (e.g., automobile engine, radio, alarm clock)?

RESPONSE: Yes—43; No—48.

2. When you were starting your engineering undergraduate study, which one of the following did you consider the most accomplished: a) Henry Ford; b) James Clerk Maxwell; c) Thomas Edison; d) Lee DeForest?

RESPONSE: Edison—60; Ford—13; DeForest—8; Maxwell—7.

Question 1 indicates that almost half of the respondents were predisposed to study engineering because of their interest in taking *things* apart and putting them together again. Since the group was almost evenly divided, one must go to question 2, where an overwhelming number selected Thomas Edison as the most accomplished. There is no point in attempting to detract from Edison's obvious accomplishments, but it must be noted that Edison's talents were those of an inventor—an applier of principles rather than a developer of principles. Based on the results of questions 1 and 2, it can be stated within the constraints of the sample size and the problematical language of the questions that the majority of engineers tested entered the profession predisposed toward *things* and the application of *things*. No attempt is made to judge whether this is the best kind of predisposition.

Questions 3 and 4

3. What percentage of your undergraduate credits were taken in nontechnical subjects, such as history, literature, languages, and philosophy? a)5%; b) 10%; c) 15%; d) 20%.

RESPONSE: 10%—30; 20%—28; 15%—17; 5%—15.

4. If your undergraduate college permitted it, would you have not pursued such courses as English, languages, or history?

RESPONSE: Yes—40; No—51.

Question 3 bears out the expected fact that in most engineering curricula, the percentage of courses in the Humanities would average about 15. Question 4, however, is much more probing but more difficult on which to base judgements. The question was framed in a negative fashion to emphasize the positive reply required from people who were not satisfied with Humanities courses. Slightly less than half of the respondents indicated they had a decided feeling that the Humanities courses had limited value and pursued them only because they were required. This reply follows logically from the findings in questions 1 and 2, from which it was concluded that the predisposition of the prospective engineering student was *thing* oriented.

Questions 5 and 6

5. What is the level of your graduate studies? a) None; b) 10 semester credits; c) 20 semester credits; d) 30 semester credits; e) Master's degree; f) Doctorate.

RESPONSE: Master's—30; 10 credits—21; None—19; 30 credits—10; 20 credits—8; Doctorate—3.

6. Was your main purpose in taking graduate work any of the following? a) Importance to your job; b) Amplify your background; c) Deepen your technical knowledge; d) Everyone says it's important.

RESPONSE: Deepen technical knowledge—43; Amplify background—23; Importance to job—9; Everyone says it's important—3.

Question 5 reveals that approximately one-third of the respondents were at the Master's degree level and above. Question 6 indicates their motivation was, in most instances, simply to deepen technical knowledge. It is interesting to note that at least a substantial portion did not look upon graduate work as job oriented, as is evidenced by the 23 replies stating that the motivation was to amplify background.

Questions 7 and 8

7. What was the estimated number of nontechnical books you read in 1964? a) More than 25; b) More than 15; c) More than 5; d) None.

RESPONSE: More than 5—52; None—18; More than 15—12; More than 25—8.

8. Have you read any of the following periodicals? a) None of the following; b) Kenyon Review; c) The Reporter, d) Ramparts; e) Saturday Review of Literature.

RESPONSE: None—48; *Saturday Review*—37; *The Reporter*—15; *Kenyon Review*—0; *Ramparts*—0.

Questions 7 and 8 were very revealing as to reading habits. Since 70 of the respondents stated that they had read no more than 15 books per year, and since these books may have included popular novels as well as detective stories, one is almost driven to conclude that, as a group, reading was not a common activity. Admittedly, one could do a great deal of reading and yet not read any books; but for present purposes, the position is taken that the reading of books is a valid basis for judgement, since it is uncommon for one who reads as a regular activity *not* to read books. Question 8 probes deeper into the area of reading habits. Four periodicals were arbitrarily selected as indices. Other periodicals could have been chosen, and this will be done in a later study. The *Saturday Review* and *The Reporter* can be purchased at most news stands in terminals and hotels, and yet only about half of the respondents said they had read them. Note that the question does not ask if they were regularly read. Slightly more than half of these people stated they had never read *any* of the listed periodicals. One would think the library exposure required of a graduate student would give him some knowledge of a periodical such as *Kenyon Review*, and the controversial nature of *Ramparts* makes its articles sufficiently newsworthy to be reported in large-circulation newspapers. (The *New York Times* regularly reports on important articles in current issues of *Ramparts*.) These two questions indicate a rather negative reading activity, and since reading is so fundamental to intellectual pursuits, they provide an insight into the intellectual habits of the tested group.

Questions 9 and 10

9. What would be the extent of what you could write about T. S. Eliot? a) Nothing; b) 10 words; c) 50 words; d) 500 words.

RESPONSE: 10 words—35; Nothing—27; 50 words—21; 500 words—7.

10. What would be the extent of what you could write about Existentialism? a) Nothing; b) 10 words; c) 50 words; d) 500 words.

RESPONSE: Nothing—44; 10 words—18; 50 words—18; 500 words—11.

Questions 9 and 10 could have both been answered in the 500-word category by any individual who reads the daily newspaper with any degree of care. The questions were very timely when asked. Since T. S. Eliot is a major contemporary poet, and since Existentialism is considered one of the significant philosophical systems of our time, it would appear that one with intellectual interests should be able to write more than 500 words on each of the subjects.

Questions 11 and 12

11. Make a list of the following activities, the most acceptable first, the least acceptable last. a) Accountant; b) Choreographer; c) Poet; d) Market researcher.

RESPONSE: Market researcher—41; Accountant—21; Poet—15; Choreographer—7.

12. If you were permitted to select one of the following as your hobby, which would you choose? a) Card playing; b) Woodworking shop; c) Sculpture; d) Reading.

RESPONSE: Woodworking—40; Reading—30; Sculpture—15; Card playing—6.

Question 11 supports the idea that the respondents considered activities concerned with *things* the most acceptable. The choreographer, who ranked lowest, is the least practical of the activities identified. Question 12 follows the same pattern, with the preponderant interest in woodworking; but it is difficult to explain one-third of the respondents selecting reading. Perhaps the clue to the answer is in the wording of the question—it does not ask what they do, but what they would do.

CONCLUSIONS

It is dangerous to draw conclusions from any data, and it is extremely dangerous in the controversial area that has been described. It appears that one can make several judgements concerning the limited sample tested and the limitations imposed by the questions. These are:

1) Despite a relatively high level of academic training, intellectual interests as defined earlier do not seem to be ~~a~~ common interest to the test sample.
2) Reading as defined by the questions asked seems to be very limited for the group tested.
3) The level of intellectual knowledge and attitude toward intellectual pursuits seems to be below that expected, considering the educational background of the test sample.

It must be noted that these judgements are limited severely by the data available, the limitations imposed by the questions asked, the small size of the sample, and the personalities of individuals who may be suspicious of questionaires and the questioner.

The intent of this paper is simply to provide a point of departure for further discussion, investigation, and, hopefully, interest in intellectual endeavors.

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APPENDIX
Intent of Questions

1. Ascertain whether the individual, in deciding to study engineering, was motivated by the *mechanical* aspects of devices or by the principles that made them possible. A *no* answer would possibly indicate that he was motivated by the concepts involved in the devices or simply that he had never thought about the matter.

2. This lists four people varying in interests from Maxwell, who developed basic magnetic theory, to Ford, who was not a scientist but was very successful in the mechanical sphere.[4] Answers would be weighed as follows, with 1 associated with the person with the least intellectual attitude and 4 with the most intellectual attitude; a=1; b=4; c=2; d=3.

3. Obtain insight into the type of undergraduate curriculum.

4. A *yes* answer would indicate the individual felt that such courses were of little value to his whole educative process.

5. Establish the level of graduate achievement.

6. Identify the purpose of graduate school. Answer (b) would indicate an attitude of understanding intellectualism; answer (c) might or might not be vocationally oriented. Answer (a) is clearly vocationally oriented, whereas (d) shows lack of promise.

7. Indicate the number of nontechnical books read. One can answer (a) and still have a reading program restricted to detective stories or popular novels; however, even this offers a wider view of the world.

8. Provide additional insight into reading habits. *SRL* is well known but is concerned with intellectual matters on a popular basis. *Reporter* is an important, politically oriented magazine. *Ramparts* and *Kenyon Review* are significant, intellectual, avant-garde types of magazines.

9. Assess familiarity with an important literary man who had much to say about our contemporary world.[5] At the time of the survey Eliot had just died and one could answer to level (c) by simply reading a good newspaper; but to do so required that the person pay attention to such matters in the newspapers.

10. Same type of question as No. 9. At the time of the survey, Sartre, a leading Existentialist, had rejected the Nobel Prize. Several columns were devoted to his work in most newspapers that used UPI or other wire services.

11. Obtain insight into the attitude of the respondents concerning two diametrically opposed types. The poet and choreographer are uncommon callings; they are *long hairs*, and the very nature of their interest is intellectually oriented, according to our definition. The accountant and market researchers are common occupations with little intellectual overtones.

12. Establish which hobby one *would* choose, not the one he presently follows. Woodworking and sculpture would reveal manual interests, with intellectual overtones in sculpturing. Reading, broadly considered, is intellectually oriented.

REFERENCES

1. S. M. Lipset, *Political Man*, Doubleday, New York, 1960
2. R. Hofstadter, *Anti-Intellectualism in American Life*, Knopf, New York, 1963
3. L. Coser, *Men of Ideas*, The Free Press, New York, 1965
4. R. M. Wik, "Henry Ford's Interest in Science," *Proceedings of the Tenth International Congress of History of Science*, 1962
5. T. S. Eliot, "The Waste Land" (any volume of collected works)

SCIENTIFIC COMPUTER APPLICATIONS PROGRAM CATALOG

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WITH the increasing use of digital computers in solving research, development, and production problems, there is a need for better communication among RCA scientists and engineers on available computer programs for scientific applications. A corporate-wide catalog is being established to furnish better knowledge of existing programs. This should result in the actual solution of more problems, minimize wasteful duplication of effort, and promote the interchange of information between RCA engineers and scientists with related interests or problems. The Scientific Computer Applications Program Catalog (SCAPC) will help in transferring information within each activity as well as between different company locations.

Development and operation of the catalog is decentralized, coordination being achieved through designated representatives at a number of RCA locations where there are centralized responsibilities for programming for technical applications. The initial representatives are as follows:

Location Symbol	
AEDH	R. Goerss, Scientific and Technical Computations Group, AED, Hightstown
ASDB	Joel L. Richmond, Data Processing, Engineering, ASD, Burlington
BCDM	F. M. Brock, Microwave Engineering, BCD, Camden
BCDT	LaFarr Stuart, Systems Application Development, EDP, Cherry Hill
DEPA	George Zorbalas, TV Tape Dept., BCD, Camden
EDPS	R. D. Smith, Applied Research, DEP, Camden
LABS	R. W. Klopfenstein, Applied Mathematics, RCA Laboratories, Princeton
M&SR	R. Faust, Scientific Information Processing, M&SR, Moorestown
WCDV	A. E. Cressey, Information Control Systems, WCD, Van Nuys

These representatives will have the responsibility for maintaining local reference copies of the catalog, for

Final manuscript received February 24, 1966.

TABLE I—SCAPC Programs and Contributors

Program No.	*Contributors*	*Program Title, Descriptors, and Languages*
AEDH-0001	S. A. Komianos R. Hilton	"Nodal I" (Circuit Analysis; Complex Matrix Inversion; Bode) 7090-7094 FORTRAN II
AEDH-0002	S. Komianos R. W. Meldrum	"TIROS Spin Increments in Elliptical Orbits" (Magnetic Attitude Control; Torquing; Spin; Plotting) 601 FORTRAN II
AEDH-0003	S. Komianos	"SADACT, Satellite Array Degradation from Actual Current Telemetry" (Solar Array; Least Squares Curve Fitting; Plotting) 601 FORTRAN II
AEDH-0004	S. Komianos R. Hilton L. Miller	"SATEB, Satellite Energy Balance" (Array Output; Electrical Losses; System Simulation) FORTRAN II
AEDH-0005	S. Komianos	"CAT, Catalogue of Array Output vs. Temperatures and Orientation" (Array Output; Solar Cells; Plotting) 601 FORTRAN II
AEDH-0006	S. Komianos D. Almy	"Satellite Contact Time" (Orbital Analysis; Ground Stations; Radiator Coverage) 601 FORTRAN II
AEDH-0007	S. Komianos	"Delay Distortion Effects on a Series of Pulses" (Distortion; Phase Distortion; Filter; Response; Plotting) 601 FORTRAN II
AEDH-0008	Louis J. Ciabattoni	"Thermal I" (N Body; Thermal Analyzer) FORTRAN II
AEDH-0009	Louis J. Ciabattoni	"A Computer Program to Calculate Radiative Configuration Factors" (CONFAC; Shape Factors; Configuration Factors) FORTRAN II
AEDH-0010	Louis J. Ciabattoni	"A Computer Program to Calculate the Isolation on a System of Planar, Convex, Polygonal Surfaces (Specular Reflections)" (Solar; Solar Absorption) FORTRAN II
AEDH-0011	Louis J. Ciabattoni	"A Computer Program to Calculate the Isolation on a System of Planar, Convex, Polygonal Surfaces (Diffuse Reflections)" (Diffuse) FORTRAN II, FAP, (GAEC Matrix Package)
AEDH-0012	Louis J. Ciabattoni	"Thermal XV (Earth)" (ALBEDO) FORTRAN II
AEDH-0013	Louis J. Ciabattoni	"Thermal XV (Moon)" (ALBEDO) FORTRAN II
AEDH-0014	Louis J. Ciabattoni Ronald A. Velosky	"Thermal XXII" (Absorptivity; Emissivity; Reflectivity) FORTRAN II
AEDH-0015	Louis J. Ciabattoni	"N Body Heat Conduction" (Heat Conduction) FORTRAN II
AEDH-0016	Louis J. Ciabattoni Ronald A. Velosky	"AUTOMAT (AUTOmatic Matrix Arithmetic Translator)" (Matrix Package) 501 Basic
AEDH-0017	Louis J. Ciabattoni Ronald A. Velosky	"Inertia and Balance Control" (Moment of Inertia; Dynamic Stability) 501 Interpreter
ASDB-0001	R. L. Mattison R. S. Fisher	"Special Matrix Subroutines to Invert and/or Solve a Set of Simultaneous Equations" (Real Matrix; Complex Matrix; Gauss-Jordan) FORTRAN IV
ASDB-0002	D. E. Farnsworth R. A. J. Gildea	"Runga-Kutta-Gill Subroutine" (First Order Ordinary Differential Equations) FORTRAN II
ASDB-0003	H. Platt	"Conjugate Gradient Subroutine" (Conjugate Gradient) FORTRAN II
BCDM-0001	F. M. Brock T. G. Marshall J. Parker	"Program Library Tape FPB (BCD #1); Computer Programs for Filter and Network Analysis & Design" (Filter; Network; Analysis and Design) Bell L1/301 ML
BCDM-0002	F. M. Brock	"Evaluation of Magnitude, Phase, and Envelope Delay of Network Functions" (Polynomials; Network Functions—Magnitude, Phase, Delay) 301 FORTRAN II
BCDM-0003	F. M. Brock	"Delay Functions" (All-pass Functions; Delay Equalizers) 301 FORTRAN II

TABLE 1—SCAPC Programs and Contributors (cont'd)

Program No.	*Contributors*	*Program Title, Descriptors, and Languages*
LABS-0001	A. Pelios	"EIGENS—A Program for Computing Eigenvalues and Eigenvectors of a Symmetric Matrix" (Single Precision; Eigenvalues; Eigenvectors; Single Precision Symmetric Matrices) 601 FORTRAN II
LABS-0002	A. Pelios	"EIGEND—A Program for Computing Eigenvalues and Eigenvectors of a Symmetric Matrix" (Double precision; Eigenvalues; Eigenvectors; Double Precision Symmetric Matrices) 601 FORTRAN II
LABS-0003	R. L. Crane	"PHASE" (Function Evaluation) FORTRAN II
LABS-0004	R. W. Klopfenstein	"Automatic Saddle Point and Branch Point Location in Root Tracing of Complex Functions of a Real and a Complex Variable" (Zeroes; Non-Linear Functions; Ordinary Differential Equations) 601 FORTRAN II
LABS-0005	R. L. Crane	"All-Pass Delay Equalizer Synthesis" (Circuit; Phase; Equalization) 601 FORTRAN II
M&SR-0001	R. Faust R. Wattis	"RCA 4101/4102 Rollback Tape Disassembly—DF043" FORTRAN II, FAP
WCDV-0001	J. G. Mackinney	"RCA 4104 A-1 Assembly Program" (Symbolic Language; Assembler) 4104 Assembly
WCDV-0002	J. G. Mackinney	"RCA 4104 Debug Monitor" (Debugging; Monitor; Trace; Storage Dump) 4104 Assembly
WCDV-0003	Jacqueline Spangler	"Electronic Switching Programming System" (Store and Forward; Message Switching) 4104 A-1 Assembly
WCDV-0004	R. L. Hooper	"RCA 110A SLAP 2 Software System" (Assembler; Loader; Operating System; Edit Program; Test; Maintenance and Utility; Mathematical Subroutines) 110A SLAP 2

R C A   No. BCDM - 0001 - ( )

SCIENTIFIC COMPUTER APPLICATIONS PROGRAM CATALOG (SCAPC)

Program Data Sheet

Program Title: Program Library Tape FPB (BCD #1)

Computer Programs for Filter & Network Analysis & Design

Descriptors: Filter; Network; Analysis and Design

Contributors: F. M. Brock, T. G. Marshall, J. Parker

Languages: Bell L1/301 ML

Machines: RCA 301-40K, Prog'md fltg. pt. (305 Processor)

ABSTRACT

This Master Program Library Tape contains 8 computer programs useful for filter and network analysis and design. The programs were originally in Bell L1 language but have been translated into valid 301 ML using the 301-40K Bell Interpretive Translator program prepared by J. Parker, RCA Lancaster. Tape FPB contains an operating control system so any program can be called into service by 2 control cards placed ahead of the program input data card deck. This operating system is arranged for normal 381 tape file maintenance.

Programs on Master Tape FPB:

FPB01 - "Ladder Network Sensitivity Analysis" - Steady state analysis of ladder networks for transmission & input/output characteristics; also for branch sensitivity.

FPB02 - "Ladder/Lattice Network & Polynomial Analysis" - Steady state analysis of ladder & lattice networks; also analyses polynomial functions (in root form) for magnitude & phase.

FPB03 - "Minimum Inductance Transformation" - Transforms normalized even order Cauer LPF to BPF with a minimum number of coils.

FPB04 - "Image Parameter BPF Image Atten. Calc." - Calc. image atten. of specified image parameter BPF sections.

FPB05 - "Image Parameter BPF 1/2 Section Calc." - Calc. element values for specified image parameter BPF 1/2 sections.

FPB06 - "Image parameter BPF Full Section Calc." - Combines 1/2 section element values as obtained from FPB05.

FPB07 - "Singly Terminated LPF Realization" - Computes element values for singly terminated LPF realizing Butterworth, Chebychev responses up to 10th order.

FPB08 - "Maximum value of a Function" - Iterates to find the maximum magnitude value of a polynomial function (expressed in root form).

EM-6404

Documentation available: Card deck?__ Program listing? x Report No.____

Reference:

Other: Limited write-ups on the individual programs are available.

Program Status: Individual pgms. in use since 1961. FLT in use since Nov. 1965.

Comments:

Date submitted Jan. 31, 1966   Form #2049 - 12/65

Fig. 1—Sample Program Data Sheet used for inputs to SCAPC. Data Sheet forms can be obtained from any of the local reference centers listed in this paper.

EDITOR'S NOTE

Engineers who use or plan to use computers in research, development, or design work are urged to send their programming ideas as directed herein. Contributors to the corporate catalog described in this paper will be given due recognition on a regular basis in the RCA ENGINEER.

encouraging the submission of input material to the catalog, and for screening these submissions. Primary input to the catalog will be a one-page data sheet, prepared in a standardized format. The program *contributor* should be someone who has used the program and is therefore familiar with it. He may or may not be the original author; indeed, useful catalog entries will come from adaptations of programs originally written by others both inside and outside RCA. Equally as useful as the availability of the programs themselves, will be the knowledge that the contributor is familiar with a particular type of problem and might be contacted by his counterpart from another location who has a related problem.

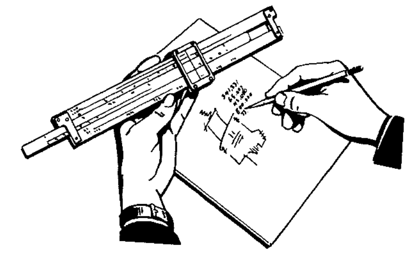
Additions to SCAPC will be reported regularly in the RCA ENGINEER. The first entries are given in Table I by contributor and title, with brief additional information and an identifying number designating the originating source and the serial order of issuance. A sample data sheet is shown in Fig. 1.

Initiation of the catalog and assistance in getting it under way has been provided by the following ad hoc committee members: R. Gildea, ASD; R. Goerss, AED; T. Hilinski, DEP-AR; C. Katz, EDP; R. Klopfenstein, RCA Laboratories; and J. Kurshan, RCA Laboratories, Chairman.

DR. JEROME KURSHAN received his AB (with honors in Math and Physics) from Columbia University in 1939 and his PhD in Physics from Cornell University in 1943. He was an Assistant in Physics at Columbia University in 1939 and at Cornell University from 1939 to 1943. Dr. Kurshan joined the RCA Laboratories in 1943, where he has conducted research on electron tubes and semiconductor devices. He has served as Mgr., Graduate Recruiting; Mgr., Technical Recruiting and Training; and Mgr., Employment and Training. He was appointed to his present position, Mgr., Research Services Laboratory, in March 1959. Dr. Kurshan is a senior member of the IEEE, and a member of the American Physical Society, Phi Beta Kappa, Sigma Xi, Phi Kappa Phi, and Pi Mu Epsilon.

Engineering and Research NOTES

BRIEF TECHNICAL PAPERS OF CURRENT INTEREST



High-Power VHF Overlay Transistor for Single-Sideband Applications

R. Rosenzweig

Z. F. Chang

R. ROSENZWEIG and
Z. F. CHANG, *Electronic Components and Devices Somerville, N.J.*

Final manuscript received January 31, 1966.

To date, most high-frequency power transistors have been designed for class C operation because forward-biasing into Class B or AB drives the device into a region where second breakdown occurs. For single-sideband operation, however, the requirement of linear amplification precludes the use of class C biasing, which is inherently nonlinear. As a result, vacuum tubes rather than transistors have traditionally been used in single-sideband applications. Recently, however, demands for lightweight, portable single-sideband equipment have led to the development of a new overlay power transistor for this type of application.

The two major advantages of single-sideband transmission are conservation of power and reduced channel width. Signal distortion resulting from nonlinear amplification minimizes these advantages, since it causes interference in other channels. Therefore, to permit operation as a class AB amplifier, a transistor for single-sideband applications must incorporate the following design features: 1) improved second-breakdown resistance, and 2) ability to maintain a forward-bias point that does not shift with temperature. In the new RCA Dev. No. TA2656 high-power VHF overlay transistor, protection from second breakdown is obtained by individual resistive ballasting of a large number of small emitter sites, and bias-point stability is achieved by use of a temperature-compensating diode located near the transistor pellet.

Second breakdown is characterized by localized thermal runaway that produces a negative-resistance region in the collector voltage-current characteristics and causes eventual destruction of the device from localized alloyed regions. In the initial stage of investigation, the factors affecting second breakdown in silicon interdigitated triple-diffused transistors were analyzed. Observation of the device surface prior to second breakdown showed that *hot spots* (as revealed by temperature-sensitive phosphors) were randomly distributed over the device area and were not associated with physical defects. The random distribution of the localized alloy regions after second breakdown is shown in Fig. 1. These investigations indicated that second breakdown is an inherent condition of a transistor rather than a defect phenomenon, and that the design of a single-sideband transistor should incorporate protection against second breakdown. Protection in the TA2656 is provided by a subdivided emitter in an overlay structure in which the individual emitter sites are resistively ballasted.

The TA2656 transistor has an overlay emitter configuration in which 180 parallel emitter sites are interconnected by metal fingers. A current-limiting diffused resistor is in series with each emitter site between the metalizing and the emitter-base junction. These resistors are diffused into the structure and can be varied in value from 1 to 25 ohms per site. This structure combines the high-frequency advantages of the overlay design with the advantage of resistively ballasted sites for second breakdown protection. The device has a high ratio of emitter periphery to collector area and emitter area, and thus combines good high-current performance with small capacitance.

A group of transistors were made with three different values of emitter resistance. Second-breakdown voltage for the three resistors at a collector current of one ampere was as follows:

Total Resistance (ohms)	*Second-Breakdown (DC volts)*
0.005	50
0.013	65
0.080	108

These data demonstrate that the addition of resistors results in considerably increased second-breakdown protection.

The TA2656 typically delivers a peak envelope power of 50 watts at a frequency of 50 MHz (Mc/s) with a gain of 10 dB in class C

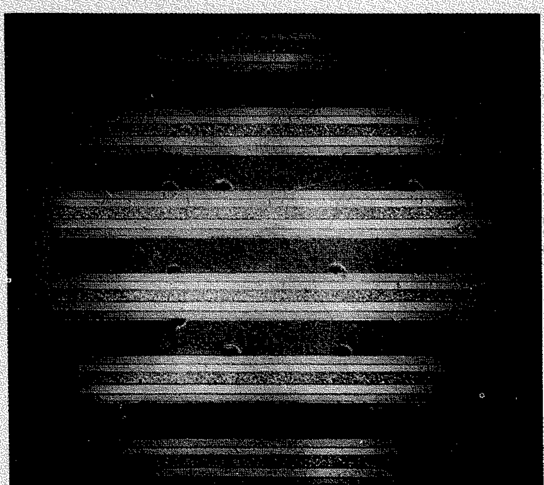


Fig. 1—Surface of interdigitated silicon transistor after second breakdown, showing random distribution of localized alloy regions.

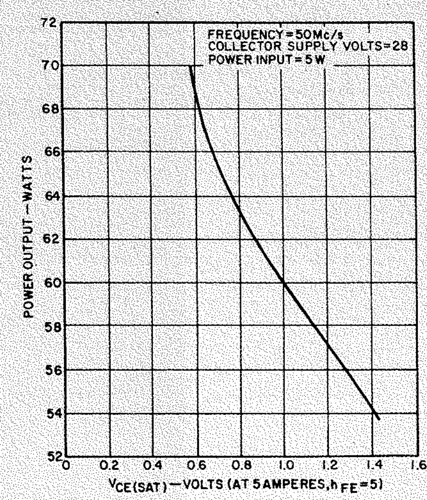


Fig. 2—Reduction in output power caused by increased emitter resistance, as measured by $V_{CE(SAT)}$.

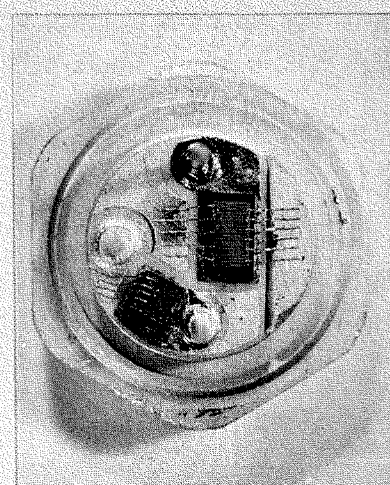


Fig. 3—Temperature-compensating diode adjacent to transistor pellet on TO-60 stud package.

service. The curve in Fig. 2 shows that peak power decreases as a function of emitter resistance, as evidenced by an increased $V_{CE(SAT)}$. At lower power levels (10 to 20 watts), gain is not affected by the addition of resistance.

The transistor pellet shown in Fig. 3 is in a TO-60 grounded-emitter package, adjacent to a temperature-compensating diode which provides forward-bias stability. The pin normally used for the emitter connection provides a terminal for the diode.

Temperature variations caused by device dissipation or changes in ambient temperature produce a shift in the quiescent operating point that degrades amplifier linearity and that can result in thermal runaway. The use of a forward-biased diode in the same package provides a base-voltage source which varies with junction temperature in the same manner as the emitter-base voltage of the transistor. The current through the diode must be large compared to the current into the base of the transistor to approximate a *stiff* voltage source.

The TA2656 is designed specifically for single-sideband linear applications in the 30- to 80-MHz range. This transistor is designed for 28-volt operation with a maximum thermal resistance of 2°C/watt. Typical single-sideband performance for third-order intermodulation distortion of —30 dB is given below for operation at 30 and 80 MHz:

Parameter	*30 MHz*	*80 MHz*
Peak envelope input power, P_{in} (watts)	2	3
Peak envelope output power, P_o (watts)	40	30
Third-order distortion (dB)	—30	—30
Power gain (dB)	13	10
Collector efficiency (%)	50	45

Third- and fifth-order distortion of the TA2656 are shown in Fig. 4 as functions of output power at 80 MHz with a 28-volt supply and a threshold bias of 0.5 volt. Peak envelope power of approximately 50 watts is obtained at an intermodulation distortion of —23 dB. Intermodulation distortion at 25 watts as a function of forward bias is shown in Fig. 5. Third-order distortion is reduced by 10 dB when the device is biased to the threshold of conduction. Further biasing into class AB reduces the distortion by an additional 5 dB and results in a stable single-sideband amplifier with a gain of 10 dB at a peak envelope power of 25 watts and a frequency of 80 MHz with intermodulation distortion of —33 dB.

This new overlay technique makes it feasible to incorporate the distributed emitter ballast resistance that is essential for thermal stability. Single-sideband performance of state-of-the-art devices has been comparable to that of tube counterparts. The light weight and high efficiency of this new developmental transistor have made possible peak envelope powers up to 100 watts in transistorized portable and vehicular single-sideband equipments.

The work described in this note was supported by the USAF Avionics Laboratory, Research and Technology Div., Air Force Systems Command.

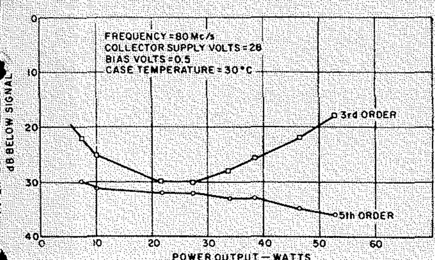


Fig. 4—Third- and fifth-order intermodulation distortion of RCA Dev. No. TA2656 as a function of output power at 80 MHz.

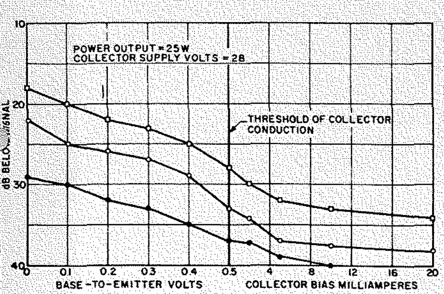


Fig. 5—Intermodulation distortion as a function of forward bias at peak envelope power of 25 watts.

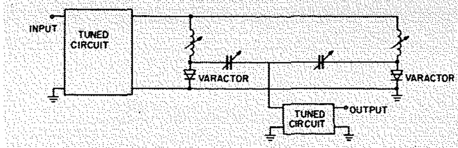


Fig. 1—Simplified schematic of frequency multiplier.

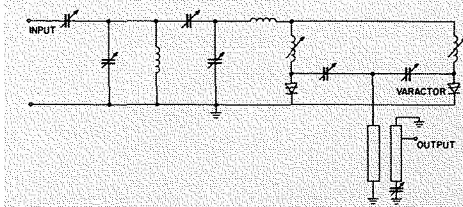


Fig. 2—Input-matching network of frequency multiplier.

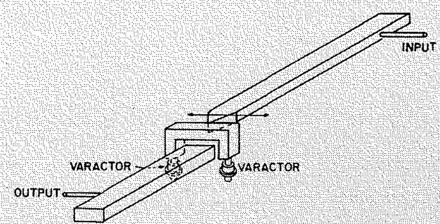
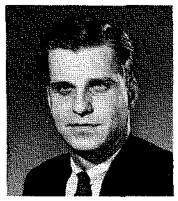


Fig. 3—Simplified diagram of stripline multiplier.

Parallel-Varactor-Diode Frequency Multiplier



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Final manuscript received February 21, 1966.

Various embodiments of parallel-varactor-diode frequency multipliers adapted for wideband, high-power operation are illustrated in Figs. 1, 2, and 3.

In Fig. 1, an input wave at a fundamental frequency is applied to an input circuit tuned to the fundamental frequency. The output of the tuned circuit is applied directly in parallel and without an intermediary coupling transformer across the parallel circuits. Since the tuned input circuit is coupled directly to the parallel circuits, the frequency multiplier of Fig. 1 can be operated at higher frequencies than would be possible with a coupling transformer inserted between the first tuned circuit and the parallel circuits.

Each parallel circuit (Fig. 1) includes an inductor and a varactor in series. An output wave at a harmonic frequency of the input wave appears at the points in the parallel circuits between the inductors and the varactors. This harmonic frequency is applied through the variable coupling capacitors to the tuned output circuit. By adjusting the inductors, the amount of current fed to the varactors at fundamental frequency can be equalized. The capacitors isolate the output terminals from one another. Adjustment of the capacitors and inductors can control the amplitude and phase of the applied fundamental frequency and the produced harmonically related output frequency. Thus, economical, standard varactors (as distinct from closely matched ones) can be made to appear matched and are therefore useful in this frequency multiplier.

The harmonic generator of Fig. 2 includes a bandpass filter between the input terminals and the parallel circuits, including the varactors. This circuit can provide 25 watts of RF power at 300 MHz (Mc/s) with a bandwidth of 14% at the secondary of the output transformer.

A harmonic generator adapted for use in the microwave region is shown in Fig. 3. The input conductor slides along and makes contact with the back of the upper part of the inverted U-shaped conductor. The legs of the inverted U-shaped conductor which lie between the sliding input conductor surface and the ends of the U correspond to the inductors shown in Figs. 1 and 2. The capacitance between the legs of the inverted U conductor and the output conductor corresponds to the capacitors of Figs. 1 and 2. The open terminals of the varactors are suitably grounded by connection to a container (not shown). The sliding conductors together with the capacitances between the conductors and the apparatus container of Fig. 3 are tuned respectively to the input and output frequencies.

The Laser as a Gas Detector

Dr. H. J. Gerritsen
RCA Laboratories, Princeton, N. J.

Final manuscript received February 15, 1966.

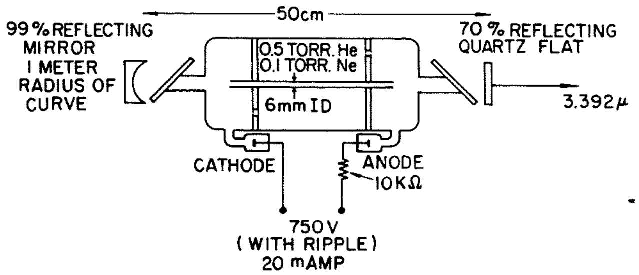


Fig. 1—Laser gas detector.

The monochromaticity and directionality of the laser make it well suited for the detection of traces of certain gases in the atmosphere. This property can be useful in detecting the presence of noxious or explosive gases, and in detecting leaks in pressurized gas sytems.

The gas detector is based on the use of an infrared laser whose beam traverses a space containing the gas to be detected, and then impinges upon an infrared detector. All gases that possess an electric dipole moment have strong fundamental absorption bands in the infrared, usually in the region of 2.5 to 10 microns. The laser wavelength is chosen to coincide with an absorption band in the gas, so that the strength of the absorption is a measure for the gas concentration.

The most important advantage of the laser over conventional infrared, spectroscopic methods is its monochromaticity. Directionality and, in some cases, efficiency are additional attributes.

One may wonder whether the width of the gaseous absorption lines is so small that it cannot easily be matched by the width obtainable with traditional spectroscopic methods. Pressure broadening due to the atmosphere results in a widening of the absorption line to a width of the order of 10^{-4} micron. A large and expensive infrared monochromator is required to reach such a high resolution; consequently, less precise instruments are normally used which transmit and receive a much wider frequency band. Since the gas to be detected does not absorb over most of this range, the sensitivity in such a method is far below that which is theoretically possible; however, the laser method can achieve the theoretical sensitivity.

Based on previous experience with infrared absorption of methane,(1,2) a simple apparatus was assembled for the detection of this gas in air (Fig. 1). This device consisted of a 20-inch-long, amplitude-modulated, helium-neon gas laser (which emits 4 milliwatts of infrared radiation at 3.3923 microns for a power input of 12 watts) and a photoconductive lead sulphide cell followed by an amplifier. The power reaching the detector drops to one-half when a 1% methane concentration is present in a 3-inch-long path in air. Such a methane concentration is generally accepted as the safety limit for prolonged breathing. Since methane-air mixtures containing 5% or more methane can cause dangerous firedamp explosions in coal mines, the installation of laser methane detectors might prevent such calamities in the future.

In addition to its potential use as a methane detector in mines, the device could also be used to locate leaks in natural-gas pipelines. Natural gas, which consists of more than 50% methane, is used increasingly as a source of energy and is distributed in the U.S. through a rapidly expanding 20-billion-dollar piping network.

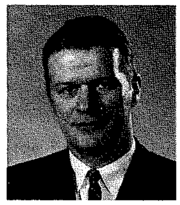
Other gases whose detection would be desirable are the poisonous carbon monoxide and air pollutants, such as sulphur and nitrogen oxides. In view of the hundreds of gas lasers known today, there is a good chance of finding a coincidence between one of these lasers and one of the rotational vibrational absorption lines in the gas to be detected.

Gas lasers are reliable, have long life, and are very monochromatic, but they are also usually rather large, moderately expensive, and generally inefficient. Although infrared injection lasers may overcome these deficiencies, at present they have certain disadvantages, such as poorer monochromaticity and shorter life expectancy than gas lasers. If continued research can solve these problems, an increasing use of lasers in determining gas concentrations can be expected.

REFERENCES

1. H. J. Gerritsen and S. A. Ahmed, *Physics Letters* 13, 41 (1964).
2. H. J. Gerritsen and M. E. Heller, *Applied Optics Suppl.* 2, 73 (1965).

Electron "Scrubbing" of High-Vacuum Systems



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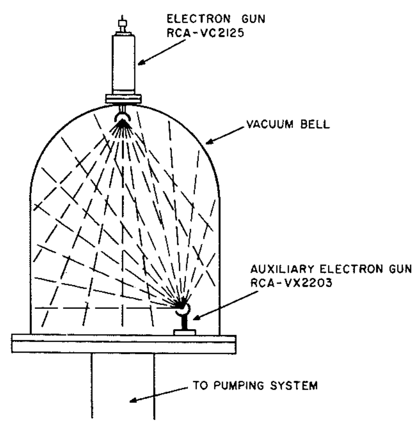
Electron *scrubbing* is a technique for *cleaning* high-vacuum systems in which high-velocity electron bombardment and the accompanying radiation desorb gaseous materials and keep them mobile until they are pumped out of the system. Until recently, baking (normally to 450°C) was generally used to liberate or desorb primarily surface (or *adsorbed*) gases by imparting average kinetic energy of the order of 0.05 electron-volt per molecule. Electron scrubbing, however, energizes and liberates gases under the surface (*absorbed* gases) by use of an electron voltage of 1000 to 40,000 volts and an electron flux density of the order of one microampere per square centimeter. Continuous scrubbing with electrons prevents gas molecules from accumulating on the surface being cleaned, and makes it possible to achieve surface cleanliness equivalent to a vacuum of 10^{-10} torr when the measured vacuum is 10^{-8} torr.

The technique of high-voltage electron bombardment for degassing purposes was first used in the manufacture of RCA superpower tubes. The use of high-velocity electron bombardment and the accompanying radiation to desorb gaseous materials and keep them mobile until they are pumped out of the tube insures ultrahigh vacuum throughout the life of the tube. This technique is still employed as a production tool in the processing of such tubes to prevent breakdown when the tubes are operated at megawatt power levels.

THE PHYSICAL MECHANISM OF OUTGASSING

Gases are adsorbed at the surface of a vacuum system by both physical and chemical reactions. The physical adsorption is the result of Van der Waal's forces; the chemical adsorption is caused by hydrogen-bond forces. Although the nature of these forces is known, their magnitudes are not, and it is difficult to conceive an accurate means of measuring them. The average kinetic energy of about 0.05 electron-volt imparted to the gas molecules in normal

Fig. 1—Diagram showing location of master and auxiliary guns in high-vacuum-system chamber for maximum coverage of surface area.



baking practice (450°C) is sufficient to energize and liberate only those gases which have activation energies below 15 kilocalories per mole. Gases that have higher energies are best liberated by electron scrubbing with high electron voltage and electron flux density of the order of one microampere per square centimeter, or 6 x 10^{12} electrons per square centimeter per second.

For example, if a condensed gas such as hydrogen or carbon monoxide forms a tightly packed, unimolecular layer on a surface, the number of molecules per layer present in an area of one square centimeter is approximately 10^{14} to 10^{15}. The collision cross section of electrons to the adsorbed molecules is then given as follows:

a) $10^{-19}/cm^2$ for adsorbed hydrogen and $10^{-16}/cm^2$ for gaseous hydrogen,
b) $10^{-18}/cm^2$ for adsorbed oxygen,
c) from 10^{-18} to $10^{-21}/cm^2$ for adsorbed carbon monoxide.

The electrons supplied for the scrubbing process, however, are not the only source of high energy available for cleaning. Soft X-rays, secondary electrons, Compton electrons, and long-wave photons are also generated. These effects tend to conserve energy so that the incident electrons impart energy to the surface by various means. When the total supplied energy excites the adsorbed molecules sufficiently, they are desorbed from the surface.

Since high-voltage electrons penetrate many molecular layers, electron scrubbing is also effective in areas below the surface. In a high-vacuum system, some gases are on the surface, whereas others are under the surface and about to migrate to the surface to be desorbed. All materials continuously outgas as the result of gases migrating and diffusing randomly through the material. Because the rate of this diffusion depends primarily on temperature rather than on surface or vacuum conditions, it is impossible to outgas a surface at a rate faster than the rate at which these diffusing gases are arriving to the surface. It is possible, however, to *flush* the surface adjacent to the vacuum environment so that the apparent rate of outgassing is reduced for a time.

For example, if a gas such as monatomic hydrogen is diffusing through aluminum at a rate of 0.005 centimeter per hour, high-voltage electrons may be used to penetrate to a maximum depth of 0.013 centimeter; the aluminum is then outgassed to this depth, and the surface is apparently free of hydrogen for a period of two to three hours. Electron scrubbing also results in some breakdown and chemical dissociation in which complex molecules are reduced to more mobile gases, which are then pumped out of the system.

TECHNIQUES OF ELECTRON SCRUBBING

The electron-beam gun used for scrubbing should be positioned in the vacuum chamber for coverage of the maximum surface area. For very large chambers, additional auxiliary guns can be located on the chamber wall, as shown in Fig. 1, to eliminate blind areas as much as possible. Such auxiliary guns consist of only a filament mount, which is connected in parallel with the filament of the *master* gun to the filament power supply outside the chamber. When the filaments are placed at a high negative potential, the chamber wall (at ground potential) acts as the anode. The use of auxiliary guns permits a simplified electrical arrangement and eliminates the need for additional openings in the chamber wall for extra scrubbing guns.

Before the electron scrubbing gun is turned on, the chamber should be vacuum-pumped as well as possible by convenient rough and intermediate pumping. For pressures into the ultra-high-vacuum range, an electronic pump of the sputter-ion type can be used. The electron scrubbing gun may be turned on at any time during the operation of the electronic pump. Observations with a mass spectrometer indicate that gases of all types are released from the walls of the vacuum chamber during cleaning. When bombardment does not remove additional gases and the pressure reaches equilibrium, the gun should be turned off.

If an ion pump is used, it is disconnected when the gun is turned off, and a cryogenic pump is started. The cryogenic pump removes the gases released from the wall of the chamber. Pressures as low as 10^{-12} torr have been achieved by use of this technique. It must be emphasized that in electron scrubbing the vacuum-pumping equipment associated with the clean chamber must be capable of sustaining the degree of cleanliness achieved; a cryogenic pump is suitable for this purpose.

The techniques described in this note are used in the RCA Dev. No.-J1931 Extreme-High-Vacuum System. This system employs the RCA VC2125 Electron Gun and the RCA Dev. No.-J1902 Cryogenic Pump.

Ferrite Bead Application to AM Transistorized Receiver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The use of a ferrite bead on the emitter lead of a convertor transistor of an AM superheterodyne receiver improves performance. The areas of improvement are:

a) Elimination or reduction of parasitic oscillations
b) Improvement of oscillator injection voltage slope across the band
c) Improvement of sensitivity slope across the band
d) Improvement of signal-to-noise characteristics
e) Reduction in the reception of undesired stations in the bands above the broadcast band.

The improvements are produced by the degenerative effect on frequencies above the AM-band frequencies, when a ferrite bead is placed in the emitter lead of the convertor transistor. The frequencies at which degeneration occurs and the magnitude of degeneration depends on the type, size, and shape of the ferrite bead used. The electrical equivalent circuit of a ferrite bead is shown in Fig. 1. R and L are frequency dependent. Inductance (L) decreases with frequency; resistance (R) increases with frequency.

The ferrite bead is placed in the emitter circuit of a convertor as shown in Fig. 2. When the circuit shown in Fig. 3 is used, high-frequency parasitic oscillations are sometimes encountered. These parasitics may have sufficient magnitude to cause breakup of the convertor waveform, thus rendering the receiver inoperative. The placement of the ferrite bead in the emitter lead of the convertor, as shown in Fig. 2, eliminates this phenomenon. The bead has little effect on the desired oscillator frequencies.

Improvement in the slope of the oscillator injection and sensitivity slope across the band is due to the varying degree of degeneration across the band achieved by the use of the ferrite bead. Purity of the oscillator waveform and narrowing of the frequency band contribute to the reduction in signal-to-noise characteristics. A transistorized AM receiver is more sensitive to undesired stations (spurious responses) when it is operated from a line rather than from batteries. When the ferrite bead is positioned as shown in Fig. 2, the magnitude of the undesired stations is reduced to a level where they are not detrimental to the performance of the receiver.

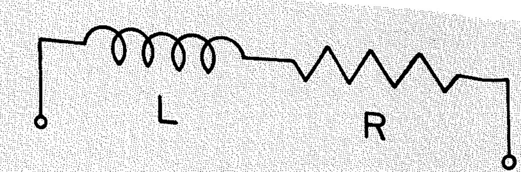


Fig. 1—Electrical equivalent circuit of ferrite bead.

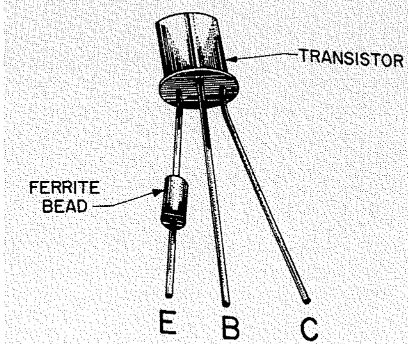


Fig. 2—Physical placement of bead.

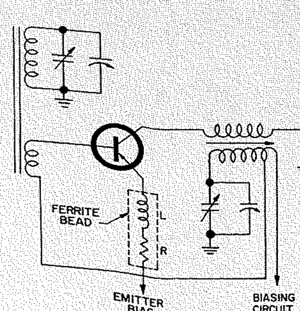
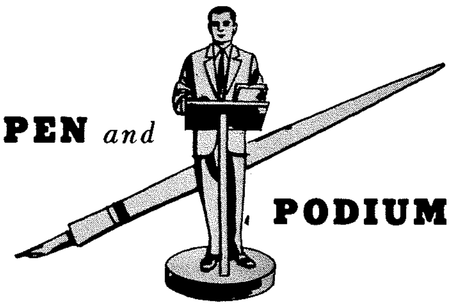


Fig. 3—Simplified converter with bead.



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SPACE ENVIRONMENT

(ASTRONOMY): Where We Stand—A Quick Astronomical Survey—A. L. Rosenberg (WCD, Van Nuys) Vet's Admin. Hospital, Toastmasters Club, Sepulveda, Calif., Nov. 22, 1965; also presented at Glendale Fed. Svgs. Assoc., Committee Room, Polaris Astronomical Assoc., Canoga Park, Calif., Feb. 24, 1966

SPACE NAVIGATION
(& guidance & tracking)

(TRAJECTORIES): Optimal Free-Space Fixed-Thrust Trajectories Using Impulsive Trajectories as Starting Iteratives—M. Handelsman (SEER, Mrstn) 3rd Aerospace Science Mtg., American Inst. of Aeronautics & Astronautics, N. Y., Jan. 24-26, 1966; to be published in *AIAA Journal*

SPACECRAFT
(& space missions)

WORLD WEATHER WATCH, Global Communications for the—H. Gurk, P. Stebe, M. Rehm (AED, Pr) 1966 Winter Convention on Aerospace & Electronic Systems, Los Angeles, Calif., Feb. 4, 1966; *Conf. Record*

WORLD WEATHER WATCH, The Role of Satellites in the—L. Krawitz (AED, Pr) Practical Space Applications Natl. Mtg.; AAS, San Diego, Calif., Feb. 23, 1966

SPACECRAFT INSTRUMENTATION

(THERMOELECTRICS): Application of Silicon-Germanium Thermoelectrics to Isotope Space-Power Systems—G. S. Lozier (ECD, Hr) Symp. on Radioisotope Applications in Aerospace, Wright-Patterson Air Force Base, Dayton, Ohio, Feb. 15-17, 1966

SUPERCONDUCTIVITY
(& cryoelectrics)

HYSTERESIS (Intrinsic) in Type II Superconductors Plated with a Normal Metal—J. I. Gittleman, B. Rosenblum (Labs, Pr) American Physical Society, N. Y., Jan. 26-29, 1966

HYSTERESIS in the Mixed State of Superconductors—A. Rothwarf (Labs, Pr) Rutgers Univ., New Brunswick, N. J., Feb. 16, 1966

MICROWAVE SURFACE IMPEDANCE of Type II Superconductors, Interpretation of—B. Rosenblum, J. I. Gittleman, A. Rothwarf (Labs, Pr) American Physical Soc., N. Y., Jan. 26-29, 1966

TRANSITION from Type I to Type II Superconductivity; Observation of Type III Behavior—G. Fischer, R. Klein, J. P. McEvoy (Labs, Pr) Institute of Physics, Manchester, England, Jan. 4-7, 1966

TELEVISION BROADCASTING
(mass-media)

TELECASTING BY SPACECRAFT After 1975, Operational—R. Marsten (AED, Pr) Practical Space Applications National Mtg., San Diego, Calif., AAS, Feb. 23, 1966

TELEVISION EQUIPMENT
(non-mass-media)

IMAGE SENSOR, A Thin-Film Solid-State—P. K. Weimer, G. Sadasiv, L. Meray-Horvath, H. Meyer, F. V. Shallcross, H. Borkan (Labs, Pr) Solid-State Circuits Conf., Phila., Pa., Feb. 10, 1966; *Conf. Digest*

TELEVISION RECEIVERS
(mass-media)

TELEVISION INTERFERENCE in Color and Black & White Receivers—E. C. Steinman (Svc. Co., C. H.) Missouri Valley Eng. Conf., Kansas City, Missouri, April 3, 1966

TRANSMISSION LINES

THREE-RESONATOR FILTERS (General) in Waveguide—R. M. Kurzrok (CSD, N. Y.) *IEEE Trans. on Microwave Theory & Techniques*, Vol. MTT 14, pp. 46-47, Jan. 1966

TUBES, ELECTRON

COLOR PICTURE TUBES, Problems Encountered in the Development of—W. J. Harrington (ECD, Lanc) Research Eng. Scientific Soc. of America, Owens-Illinois Chapter, Toledo, Ohio, Jan. 12, 1966

TRAVELING-WAVE TUBES, Observations on High-Efficiency—M. J. Schindler (ECD, Hr) *Electronic Communicator*, Jan. 1966

TUBE COMPONENTS
(& fabrication)

GETTERS—E. P. Bertin (ECD, Hr) *Encyclopedia of Chemistry*, 2nd Ed., Jan. 1966

VACUUM
(techniques & equipment)

GETTERS—E. P. Bertin (ECD, Hr) *Encyclopedia of Chemistry*, 2nd Ed., Jan. 1966

HIGH-VACUUM CHEMISTRY, Thin Films and—L. Pensak (Labs, Pr) American Vacuum Soc. Section Mtg., Phila., Pa., Jan. 20, 1966

AUTHOR INDEX

Subject listed opposite each author's name indicates where complete citation to his paper may be found in the subject index. Where an author has more than one paper, his name is repeated for each.

Patents Granted

TO RCA ENGINEERS

As Reported By RCA Domestic Patents, Princeton

RCA LABORATORIES

Cryoelectric Inductive Switching Circuits—R. A. Gange (Labs, Pr) *U.S. Pat. 3,238,378*, March 1, 1966

Dual Element Superconductive Memory—G. A. Alphonse (Labs, Pr) *U.S. Pat. 3,238,512*, March 1, 1966

Timing Apparatus—K. J. Magnusson (Labs, Pr) *U.S. Pat. 3,236,107*, Feb. 22, 1966

Screen Electrode for Color Cathode Ray Tube—R. D. Thompson (Labs, Pr) *U.S. Pat. 3,237,038*, Feb. 22, 1966

Corona Generating Circuits—M. M. Sowiak (Labs, Pr) *U.S. Pat. 3,237,068*, Feb. 22, 1966

Bistable Electron Device Comprising Axially Spaced Dynodes—J. W. Schwartz (Labs, Pr) *U.S. Pat. 3,229,213*, Jan. 11, 1966 (assigned to U. S. Gov't.)

Noise Reduction of Traveling-Wave Tubes by Circuit Refrigeration—S. Bloom (Labs, Pr) *U.S. Pat. 3,211,947*, Oct. 12, 1965 (assigned to U. S. Gov't.)

Monoclinic Zinc Phosphide Preparation—I. J. Hegyi (Labs, Pr) *U.S. Pat. 3,230,044*, Jan. 18, 1966

Logic Circuits Employing Negative Resistance Elements—M. H. Lewin (Labs, Pr) *U.S. Pat. 3,230,384*, Jan. 18, 1966

Unidirectional Signal Propagation Circuit Including Negative Resistance Elements—M. H. Lewin (Labs, Pr) *U.S. Pat. 3,230,385*, Jan. 18, 1966

Cryoelectric Switching Trees—R. W. Ahrons (Labs, Pr) *U.S. Pat. 3,230,391*, Jan. 18, 1966

Sequential Switching Circuit which Employs the Turn-Off Delay of a Saturated Transistor for Interim Storage—J. J. Amodei (Labs, Pr) *U.S. Pat. 3,230,393*, Jan. 18, 1966

Art of Making Electron-Optical Reticles—H. B. Law (Labs, Pr) *U.S. Pat. 3,231,380*, Jan. 25, 1966

Cascaded Phosphor Layer for Color Display Including One of Discrete Coherent Particles—D. H. Pritchard (Labs, Pr) *U.S. Pat. 3,231,775*, Jan. 25, 1966

Method and Apparatus for Producing Phase Modulation of Light with a Semiconductor—B. Rosenblum (Labs, Pr) *U.S. Pat. 3,233,108*, Feb. 1, 1966

Integrated Insulated-Gate Field-Effect Transistor Circuit on a Single Substrate Employing Substrate-Electrode Bias—F. P. Heiman (Labs, Pr) *U.S. Pat. 3,233,123*, Feb. 1, 1966

Acoustic Apparatus and Method for Analyzing Speech—H. Belar (Labs, Pr) *U.S. Pat. 3,234,332*, Feb. 8, 1966

Air Ionizer—A. Streib (Labs, Pr) *U.S. Pat. 3,234, 432*, Feb. 8, 1966

Thin Film Cryotron—G. A. Alphonse (Labs, Pr) *U.S. Pat. 3,234,439*, Feb. 8, 1966

ELECTRONIC COMPONENTS AND DEVICES

Electron-Guns—M. K. Brown (ECD, Lanc) *U.S. Pat. 3,238,409*, March 1, 1966

Method of Manufacturing a Thermoelectric Device—C. W. Horsting (ECD, Hr) *U.S. Pat. 3,235,957*, Feb. 22, 1966

Optical Heterodyne Demodulator—F. Sterzer (ECD, Pr) *U.S. Pat. 3,237,011*, Feb. 22, 1966

Memory Construction—D. M. Chamberlain, N. H. Savoie (ECD, Needham) *U.S. Pat. 3,237,173*, Feb. 22, 1966

FM Stereo Multiplex Receiver Having Limiting Means in the Pilot Channel—L. Plus, R. A. Santilli (ECD, Som) *U.S. Pat. 3,235, 663*, Feb. 15, 1966

Method of Fabricating a Cathode Ray Tube—K. A. Barnes (ECD, Marion) *U.S. Pat. 3,230,600*, Jan. 25, 1966

Single Ended Transistor Amplifier Including a Biasing Network with Capacitor Voltage Stabilization—C. E. Wheatley Jr. (ECD, Som) *U.S. Pat. 3,233,184*, Feb. 1, 1966

Storage Circuits—S. B. Dinman (ECD, Natick) *U.S. Pat. 3,234,401*, Feb. 8, 1966

Heater-Cathode Mount—J. S. Posner, P. P. Spann (ECD, Hr) *U.S. Pat. 3,234,425*, Feb. 8, 1966

RESEARCH AND ENGINEERING STAFF

Signal Translating Circuit—C. J. Hirsch (R&E, Pr) *U.S. Pat. 3,235,806*, Feb. 15, 1966

DEFENSE ELECTRONIC PRODUCTS

Techniques Associated with Inductive Sensing of Tunnel Diode Memory Cells—M. M. Kaufman (DEP-AppRes) *U.S. Pat. 3,222,756*, Dec. 14, 1965 (assigned to U.S. Gov't.)

Radio Direction Finder—W. C. Grubb, Jr. (DEP-MSR, Mrstn) *U.S. Pat. 3,217,327*, Nov. 9, 1965 (assigned to U.S. Gov't.)

Automatic Maintenance Circuit—I. Brown (DEP-MSR, Mrstn) *U.S. Pat. 3,219,913*, Nov. 23, 1965 (assigned to U.S. Gov't.)

Socket for Micro-Modules—W. E. Comfort (DEP-CSD, Cam) *U.S. Pat. 3,225,324*, Dec. 21, 1965 (assigned to U.S. Gov't.)

Tape Control Apparatus—R. W. Chambers (DEP-MSR, Mrstn) *U.S. Pat. 3,231,163*, Jan. 25, 1966

Methods for Preparing Etch Resists Using an Electrostatic Image Developer Composition—L. J. Sciambi (DEP-CSD, Cam) *U.S. Pat. 3,231,374*, Jan. 25, 1966

Electrostatic Printing—L. J. Sciambi, V. L. Marquart (DEP-CSD, Cam) *U.S. Pat. 3,231,375*, Jan. 25, 1966

Stereophonic Subcarrier Signal Generator—A. H. Bott (DEP-CSD, Cam) *U.S. Pat. 3,231,673*, Jan. 25, 1966

Asynchronous Binary Counter Circuits—A. S. Merriam (DEP-AppRes, Cam) *U.S. Pat. 3,238,461*, March 1, 1966

Logic Circuit—W. C. Mavity (DEP-WCD, Van Nuys) *U.S. Pat. 3,237,024*, Feb. 22, 1966

Variable Capacitor Analog to a Digital Converter—I. S. Shair, R. H. Colbourne (DEP-ASD, Burl) *U.S. Pat. 3,237,188*, Feb. 22, 1966

RCA VICTOR HOME INSTRUMENTS

Phase Detector—G. E. Theriault (HI, Indpls) *U.S. Pat. 3,233,122*, Feb. 1, 1966

Direct Coupled Circuit Utilizing Field-Effect Transistors—G. E. Theriault (HI, Indpls) *U.S. Pat. 3,233,186*, Feb. 1, 1966

Article Holding Device—J. M. Ammerman (HI, Indpls) *U.S. Pat. 3,233,856*, Feb. 8, 1966

ELECTRONIC DATA PROCESSING

Tape Handling Apparatus—S. Klein (EDP, Cam) *U.S. Pat. 3,236,429*, Feb. 22, 1966

Trigger Circuits—R. H. Jenkins (EDP, Cam) *U.S. Pat. 3,237,021*, Feb. 22, 1966

Binary Recording System—R. H. Jenkins (EDP, Cam) *U.S. Pat. 3,237,176*, Feb. 22, 1966

Code Translator—R. H. Yen, Y. Wu (EDP, Cam) *U.S. Pat. 3,237,185*, Feb. 22, 1966

Document Handling Apparatus—Yow-Jiun Hu (EDP, Cam) *U.S. Pat. 3,231,266*, Jan. 25, 1966

Roller Mechanism—D. J. Wray (EDP, Fla) *U.S. Pat. 3,233,807*, Feb. 8, 1966

Data Processing System—L. L. Rakoczi, E. Gloates (EDP, Cam) *U.S. Pat. 3,234,518*, Feb. 8, 1966

Data Processing System—L. L. Rakoczi, Fuh-Lin Wang (EDP, Cam) *U.S. Pat. 3,234,520*, Feb. 8, 1966

Data Processing System—J. A. Weisbecker (EDP, Cam) *U.S. Pat. 3,234,521*, Feb. 8, 1966

Semi-Permanent Memory—N. C. Lincoln, P. H. Reynolds (EDP, Cam) *U.S. Pat. 3,234,528*, Feb. 8, 1966

Semi-Permanent Memory—C. Y. Hsueh, D. E. Williams (EDP, Cam) *U.S. Pat. 3,234,529*, Feb. 8, 1966

Bidirectional Amplifiers—G. H. Wells (EDP, Cam) *U.S. Pat. 3,238,310*, March 1, 1966

Meetings

MAY 1-6, 1966: **129th Mtg. of the Electrochemical Soc.**, Sheraton-Cleveland Hotel, Cleveland, Ohio. **Prog. Info.:** The Electrochemical Soc., Inc., 30 E. 42nd St., N.Y., N.Y. 10017. Attn. Dr. Ernest B. Yeager.

MAY 2-4, 1966: **12th Nat'l ISA Aerospace Instrumentation Symp.**, Marriott Motor Hotel, Phila., Pa. **Prog. Info.:** R. L. Reich, Div. Director, Bolsey Assoc., Inc., 15 Crescent St., Glenbrook, Conn. 06906.

MAY 2-4, 1966: **1966 AIAA Communications Satellite Systems Conf.**, AIAA, IEEE; Wash., D.C. **Prog. Info.:** N. Feldman, Electronics Dept., The Rand Corp., 1700 Main St., Santa Monica, Calif.

MAY 4-6, 1966: **1966 Electronic Components Conf.**, IEEE, G-IGA, ASME-SAE; San Francisco, Calif. **Prog. Info.:** IEEE Headquarters, Box A, Lenox Hill Station, N.Y., N.Y.

MAY 5-7, 1966: **7th Symp. on Human Factors in Electronics**, IEEE, G-HFE, Minneapolis, Minnesota. **Prog. Info.:** IEEE Headquarters, Box A, Lenox Hill Station, N.Y., N.Y.

MAY 10-12, 1966: **Nat'l. Telemetering Conf.**, IEEE, G-AES, G-ComTech, AIAA-ISA, Prudential Center, Boston, Mass. **Prog. Info.:** Dr. A. J. Kelley, NASA-ERC, 575 Tech. Sq., Cambridge, Mass.

MAY 11-13, 1966: **12th Nat'l ISA Symp.**, ISA, Shamrock Hilton Hotel, Houston, Texas. **Prog. Info.:** G. I. Doering, Tech. Prog. Chairman, Industrial Nucleonics Corp., 650 Ackerman Rd., Columbus, Ohio.

MAY 16-18, 1966: **NAECON (Nat'l Aerospace Elec. Conf.)**, IEEE, G-AES, AIAA, Dayton Sect., Dayton, Ohio. **Prog Info.:** IEEE Dayton Office, 1414 E. 3rd St., Dayton 3, Ohio.

MAY 16-18, 1966: **9th Nat'l ISA Power Instrumentation Symposium**, Statler Hilton Hotel, Detroit, Michigan. **Prog. Info.:** R. C. Austin, Gen. Chairman, The Detroit Edison Co., 2000 Second Avenue, Detroit, Mich. 48226.

MAY 16-18, 1966: **1966 Internat'l Symp. on Microwave Theory & Techniques**, IEEE, G-MTT, Palo Alto, Calif. **Prog. Info.:** Rudolph E. Henning, c/o Sperry Microwave Co., P.O. Box 1828, Clearwater, Fla.

MAY 16-19, 1966: **4th Nat'l ISA Symp. on Biomedical Sciences Instrumentation**, ISA, Disneyland Hotel, Anaheim, Calif. **Prog. Info.:** Dr. Thomas B. Weber, Prog. Co-Chairman, Beckman Instruments, Inc., 2500 Harbor Blvd., Fullerton, Calif.

JUNE 1-4, 1966: **Accoustical Society of America Mtg.**, Boston, Mass. **Prog. Info.:** L. Batchelder, Raytheon Co., 20 Seyon St., Waltham, Mass.

JUNE 6-8, 1966: **Design & Construction of Large Steerable Aerials for Satellite Communications, Radio Astronomy & Radar Mtg.**, IEEE, IEE, et al., IEE, Savoy Pl., London, England. **Prog. Info.:** IEEE Headquarters, Box A, Lenox Hill Station, N.Y., N.Y.

JUNE 6-9, 1966: **1966 San Diego Symp. for Biomedical Engrg.**, San Diego, Calif. **Prog. Info.:** L. T. Gregg Pub. Chairman, 2063 Cardinal Dr., San Diego, Calif. 92123.

JUNE 6-10, 1966: **1966 Third Internat'l Research Symp. on Electric Contact Phenomena**, Orono, Maine. **Prog. Info.:** IEEE Headquarters, Box A, Lenox Hill Station, N.Y., N.Y.

JUNE 13-14, 1966: **Chicago Spring Conf. on Broadcast and Television Receivers**, G-BTR Chicago Sect., IEEE, O'Hare Inn, Des Plaines, Ill. **Prog. Info.:** IEEE Headquarters, Box A, Lenox Hill Station, N.Y., N.Y.

JUNE 13-17, 1966: **Soc. for Applied Spectroscopy Mtg.**, Chicago, Ill. **Prog. Info.:** J. E. Burroughs, Borg-Warner Corp., Roy C. Ingersoll Res. Ctr., Wolf and Algonquin Rds., Des Plaines, Ill.

JUNE 15-17, 1966: **1966 IEEE Intl. Communications Conf., IEEE**, G-ComTech, et al., Sheraton Hotel, Phila., Pa. **Prog. Info.:** IEEE Headquarters, Box A, Lenox Hill Station, N.Y., N.Y.

JUNE 20-25, 1966: **3rd Congress of the Int'l Federation of Automatic Control** (IFAC), London, England. **Prog. Info.:** Prof. Gerald Weiss, Polytechnic Institute of Brooklyn, 333 Jay St., Brooklyn, N.Y. 11201.

PROFESSIONAL MEETINGS

DATES *and* DEADLINES

Be sure deadlines are met—consult your Technical Publications Administrator or your Editorial Representative for the lead time necessary to obtain RCA approvals (and government approvals, if applicable). Remember, abstracts and manuscripts must be so approved BEFORE sending them to the meeting committee.

JUNE 21-23, 1966: **Conf. on Precision Electromagnetic Measurements**, IEEE, G-IM NBS, NBS Standards Lab., Boulder, Colo. **Prog. Info.:** Dr. Kiyo Tomiyasu, Genl. Electric Co., Schenectady, N.Y.

JULY 4-8, 1966: **Rarefied Gas Dynamics Mtg.**, Amer. Phys. Soc., Oxford, England. **Prog. Info.:** C. L. Brundin, Dept. of Eng. Science, Univ. of Oxford, Parks Rd., Oxford, England.

JULY 11-13, 1966: **Electromagnetic Compatibility Symp.**, IEEE, G-EMC, San Francisco Hilton Hotel, San Francisco, Calif. **Prog. Info.:** A. Fong, Hewlett Packard Co., 1501 Page Mill Rd., Palo Alto.

JULY 11-15, 1966: **1966 Aerospace Systems Conf.**, IEEE, G-AES, Olympic Hotel, Seattle, Wash. **Prog. Info.:** T. J. Martin, 3811 E. Howell St., Seattle, Wash. 98122.

JULY 12-14, 1966: **William Hunt Eisenman Conf. on Failure Analysis**—Amer. Soc. for Metals at the Waldorf-Astoria Hotel, N.Y.C. **Prog. Info.:** Ronald J. Seman, News Bureau Mgr., Amer. Soc. for Metals, Metals Park, Ohio.

JULY 18-22, 1966: **Nuclear and Space Radiation Effects Conf.**, IEEE, G-NS, Stanford Univ., Stanford, Calif. **Prog. Info.:** Dr. Victor vanLint, Gen'l Atomic P.O. Box 608, San Diego, Calif.

JULY 25-27, 1966: **Rochester Conf. on Data Acquisition & Processing in Biology & Medicine**, IEEE, G-EMB, Univ. of Rochester, Rochester, N.Y. **Prog. Info.:** IEEE Headquarters, Box A, Lenox Hll Station, N.Y., N.Y.

AUGUST 1-5, 1966: **3rd ISA Research Conf. on Instrumentation Science** (Attendance by application only), William Smith College, Geneva, N.Y. **Prog. Info.:** K. B. Schnelle, ISA Headquarters, 530 William Penn Place, Pittsburgh, Pa. 15219.

AUGUST 17-19, 1966: **7th Annual Joint Automatic Control Conf. (JACC)**, ISA, AIAA, Univ. of Washington, Seattle, Wash. **Prog. Info.:** A. E. Bryson, Jr., Prog. Chairman, AIAA Headquarters, 1290 Sixth Ave., N.Y., N.Y. 10019.

AUGUST 30-SEPT. 1, 1966: **21st Nat'l Conf., Assoc. for Computing Machinery**, IEEE, ACM, Ambassador Hotel, Los Angeles, Calif. **Prog. Info.:** B. R. Parker, Tech. Prog. Chairman, 21st Nat'l Conf., P.O. Box 4233, Panorama City, Calif.

Calls for Papers

AUG. 23-~~26~~, 1966: **WESCON (Western Electronic Show & Convention)**, IEEE, WEMA, Sports Arena, Los Angeles. **Deadline:** Abstracts, approx. **5/1/66.** FOR INFO.: IEEE LA. Office, 3600 Wilshire Blvd., Los Angeles, Calif.

SEPT. 5-9, 1966: **Intl. Organization for Pure and Applied Biophysics Mtg.**, Amer. Phys. Soc., Vienna, Austria. **Deadline:** Abstracts, **5/15/66.** TO: E. Weidenahus, Viennese Medical Academy, Alserstr 4, Vienna 9, Austria.

SEPT. 8-13, 1966: **Physics of Semiconductors Mtg.**, Physical Soc. of Japan, Tokyo, Japan. **For Deadline Info.:** G. M. Hatoyama, PSJ, Hongo, P.O. Box 28, Tokyo, Japan.

SEPT. 12-14, 1966: **Eastern Conv. on Aerospace & Electronic Systems**, IEEE, G-AED, Wash. Hilton Hotel, Wash., D.C. **For Deadline Info.:** IEEE Headquarters, Box A, Lenox Hill Station, N.Y., N.Y.

SEPT. 21-23, 1966: **Physics of Semiconducting Compounds**, Inst. of Physics and Physical Soc., Univ. of Wales. **For Deadline Info.:** R. H. Jones, Dept. of Physics, Univ. College of Swansea, Singleton Park, Swansea, England.

SEPT. 22-24, 1966: **16th IEEE Broadcast Symp.**, IEEE, G-B, Mayflower Hotel, Wash., D.C. **For Deadline Info.:** IEEE Headquarters, Box A, Lenox Hill Station, N.Y.

SEPT. 23-24, 1966: **14th Annual Cedar Rapids Communication Symp.**, Cedar Rapids, Iowa. **For Deadline Info.:** IEEE Headquarters, Box A, Lenox Hill Station, N.Y.

SEPT. 25-28, 1966: **1st Natl. Conf. on Non-Conventional Energy Conversion Applications**, ASME, AIAA, AIChE, IEEE, International Hotel, Los Angeles, Calif. **For Deadline Info.:** R. E. Henderson, The Allison Co., Indianapolis, Ind.

OCT. 3-5 1966: **Natl. Electronics Conf.**, IEEE, et al., McCormick Place, Chicago, Ill. **Deadline:** Abstracts, approx. **5/1/66.** FOR INFO.: Natl. Elec. Conf., 228 N. LaSalle St., Chicago 1, Ill.

OCT. 3-5, 1966: **Aerospace & Electronic Systems Conv.**, IEEE, G-AES, Sheraton Park Hotel, Wash., D.C. **For Deadline Info.:** IEEE Headquarters, Box A, Lenox Hill Station, N.Y.

OCT. 5-7, 1966: **Allerton Conf. on Circuit & System Theory**, IEEE, G-CT, Univ. of Ill., Conf. Center Univ. of Illinois, Monticello, Ill. **For Deadline Info.:** Prof. W. R. Perkins, Dept. of EE, Univ. of Ill., Urbana, Ill.

OCT. 13-14, 1966: **4th Canadian Symp. on Communications**, IEEE, Region 7, Queen Elizabeth Hotel, Montreal, Canada. **For Deadline Info.:** Prof. G. W. Farnell, McGill Univ., 805 Sherbrooke St. W., Montreal, Canada.

OCT. 19-21, 1966: **13th Nuclear Science Symp.**, IEEE, G-NS, Statler Hilton, Boston, Mass. **For Deadline Info.:** IEEE Headquarters, Box A, Lenox Hill Station, N.Y., N.Y.

OCT. 20-22, 1966: **Electron Devices Mtg.**, IEEE, G-ED, Sheraton Park Hotel, Wash., D.C. **Deadline:** Abstracts approx. **8/1/66.** FOR INFO.: IEEE Headquarters, Box A, Lenox Hill Station, N.Y., N.Y.

OCT. 24-27, 1966: **21st Annual ISA Conf. & Exhibit**, Hotel New Yorker & Statler Hilton Hotel; Coliseum; N.Y.C. **Deadline:** Abstracts, approx. **6/15/66** TO: Conference Prog. Coordinator, c/o ISA Headquarters, 530 William Penn Place, Pittsburgh, Pa. 15219.

OCT. 26-28, 1966: **East Coast Conf. on Aerospace & Navig. Elec. (ECCANE)**, IEEE, G-AES, Balt. Sect., Baltimore, Md. **Deadline:** Abstracts, approx. **6/4/66.** FOR INFO.: IEEE Headquarters, Box A, Lenox Hill Station, N.Y., N.Y.

NOV. 2-4, 1966: **N.E. Research & Eng. Mtg. (NEREM)**, IEEE, Region 1, Boston, Mass. **Deadline:** Abstracts approx. **6/7/66.** FOR INFO: IEEE Headquarters, Box A, Lenox Hill Station, N.Y., N.Y.

NOV. 7-9, 1966: **Symp. on Automatic Support Systems for Advanced Maintainability**, St. Louis Sect. Colony Motor Inn, Clayton, Mo. **Deadline:** Abstracts, **5/1/66** to IEEE Hdqtrs., Box A, Lenox Hill Station, N.Y., N.Y.

NOV. 8-10, 1966: **Fall Joint Computer Conf.**, IEEE, AFIPS, (IEE-ACM), Brooks Hall, Civic Center, San Francisco, Calif. **For Deadline Info.:** AFIPS Headquarters, 211 E. 43rd St., N.Y., N.Y.

NOV. 14-16, 1966: **19th Annual Conf. on Eng. in Medicine & Biology**, ISA, IEEE, Sheraton-Palace Hotel, San Francisco, Calif. **Deadline:** Abstracts, **6/1/66.** TO: Dr. Victor Bolie, Genl. Chairman, Autonetics, 3370 Miraloma Ave., Anaheim, Calif.

NOV. 15-17, 1966: **Electric Welding Conf.**, IEEE, G-IGA, Rackham Bldg., Engrg. Society of Detroit, Detroit, Michigan. **For Deadline Info.:** M. Zucker, Myron Zucker Eng. Co., 708 W. Long Lake Rd., Bloomfield Hills, Mich.

NOV. 15-18, 1966: **12th Conf. on Magnet. & Mag. Materials**, IEEE, G-MAG et al., Sheraton Park Hotel, Wash., D.C. **Deadline:** Abstracts, approx. **8/19/66.** FOR INFO.: IEEE Headquarters, Box A, Lenox Hill Station, N.Y., N.Y.

Engineering NEWS *and* HIGHLIGHTS

13 RCA EMPLOYEES AWARDED DAVID SARNOFF FELLOWSHIPS

David Sarnoff Fellowships for graduate study in the 1966-67 academic year have been awarded to 13 RCA employees. The Fellowships, established in honor of the Chairman of the Board of RCA, range in value to as high as $6500 each. Although appointments are for one academic year, each Fellow is eligible for reappointment. The David Sarnoff Fellows are selected on the basis of academic aptitude, character, and promise of professional achievement.

New David Sarnoff Fellows in Science are: **James C. Blair,** AED, Hightstown, N. J.—toward a Doctorate in Electrical Engineering at the University of Pennsylvania; **Joseph R. Burns,** RCA Laboratories, Princeton, N. J.—toward a Doctorate in Electrical Engineering at Rutgers University; **Albert W. Weinrich,** CSD, Camden, N. J.—toward a Doctorate in Electrical Engineering at Purdue University.

David Sarnoff Fellows in Science who have been reappointed include: **Maurice W. Cha Fong,** RCA Communications, Inc., New York—toward a Doctorate in Systems Engineering at Polytechnic Institute of Brooklyn; **Geoffrey Hyde,** M&SR Div., Moorestown, N. J.—toward a Doctorate in Electrical Engineering at the University of Pennsylvania; **Chi-sheng Liu,** RCA Victor Home Instruments Div., Indianapolis, Ind.—toward a Doctorate in Electrical Engineering at the University of Illinois; **Louis Sickles,** Applied Research, Camden, N. J.—toward a Doctorate in Electrical Engineering at the University of Pennsylvania; **Frank I. Zonis,** RCA Laboratories, Princeton, N. J.—toward a Doctorate in Electrical Engineering at the University of Pennsylvania; **Joseph P. McEvoy,** Applied Research, Camden, N. J.—toward a Doctorate in Metallurgy at Imperial College of Science and Technology, London, England.

In addition to those in Science, three David Sarnoff Fellows will study toward graduate degrees in Business Administration and one Fellow toward a graduate degree in Fine Arts.

DR. G. H. BROWN & S. W. WATSON ADDRESS STANDARDS CONVENTION

Dr. George H. Brown, Executive Vice President, Research and Engineering, was the keynote speaker at the 16th National Conference on Standards sponsored by the American Standards Association at San Francisco in February. **S. H. Watson,** Director, Corporate Standards, spoke on Management's Utilization of Standardization" at the session devoted to Standardization and Management.

DR. V. K. ZWORYKIN RECEIVES VWOA DEFOREST AUDION AWARD

The 1966 DeForest Audion Award of the Veteran Wireless Operators Association was presented to **Dr. Vladimir K. Zworykin** February 19 at the Awards Banquet of the Association at the Park-Sheraton Hotel, New York City. The award, which was presented by **Dr. James Hillier,** was given to Dr. Zworykin in recognition of "his important contributions to physical and medical electronics."—*C. W. Sall*

W. R. Isom

W. R. ISOM NEW CHIEF ENGINEER

W. R. Isom has been appointed Chief Engineer, Engineering Department in the Victor Record Division, reporting to **N. Racusin,** Division Vice President and Operations Manager. Mr. Isom received his BS degree from Butler University in 1931 and taught there from 1937 to 1944, when he joined RCA at Indianapolis. He developed the first commercially available TV film projector and many special mechanisms for kinescope recording equipment for advancing film during vertical blanking time of a TV system, and for sound-recording equipment for both films and magnetic tape. He also developed precise, high-velocity, large-capacity magnetic-recording systems using tape, drums, and disks. He has pioneered the use of air bearings, air suspensions, and air-floated heads for video recording, tape and drum memories, and military tape and drum systems for broadband recording and radar data processing. Mr. Isom is a fellow of SMPTE, a senior member of the IEEE, and a member of the Editorial Advisory Board of the RCA ENGINEER.

DR. ENGSTROM AND R. D. KELL HONORED BY IEEE AWARDS

Dr. Elmer W. Engstrom, Chief Executive Officer of RCA and Chairman of the Executive Committee, was the recipient of the IEEE 1966 Founders Award for "his leadership in management and integration of research and development programs and for his foresighted application of the systems engineering concept in bringing television to the public." The award was presented at the IEEE International Convention in March. At the same time, **Ray Davis Kell,** a Fellow of the Technical Staff of RCA Laboratories, was presented the Vladimir K. Zworykin Award of the IEEE for his "extensive and significant contributions, papers, and inventions which have been fundamental in the development of both black-and-white and color television."—*C. W. Sall*

EDWARD STANKO RETIRES FROM RCA SERVICE CO.

Ed Stanko, Manager of Service Engineering, RCA Service Company and an Editorial Representative for the RCA ENGINEER, retired March 1, after 28 years of service with RCA. He joined RCA's Installation and Service Dept., Camden, N. J., in 1937. He holds several patents ranging from transformers to aircraft landing equipment. He is a licensed professional engineer in the State of N. J., a licensed amateur radio operator, and a senior life member of the Society of Motion Picture and TV Engineers, the Electron Microscope Society of America, and a member of the Board of Technical Advisors of the RCA Institutes.

M&SR ANNOUNCES WINNERS OF TECHNICAL EXCELLENCE AWARDS

Chief Engineer's Technical Excellence Committee at M&SR, Moorestown, N. J., presented the 1965 Annual Technical Excellence Award to **Tom Howard** in March. Mr. Howard, who was cited for his efforts in advanced radar signal-processing technology, made important contributions to the following programs: high-definition radar, coherent signal processor for the AN/FPQ-6, Floyd Site signal-processing test facility, and restricted bandwith signal-processing technique (REBAT).

Eight other M&SR engineers were the recipients of quarterly Technical Excellence Awards in March. They include:

R. A. Baugh—For devising practical solutions to the complex digital control problems of steering and control of phased array antennas.

D. F. Bowman—For the analytical and experimental development of a transverse feed for a spherical reflector.

C. J. Brown—For leadership in the design, development, and testing of RCA's first phased-array radar.

G. W. Brunner—For contributions to the development of the antennule for the REST antenna system, having high overall efficiency, by utilizing an efficient power divider and advanced element matching techniques.

Y. H. Dong—For contributions to the structural concept and demonstrating the feasibility of utilizing economical cement-asbestos pipe in lieu of steel towers as an antenna support structure for a major proposed system.

J. R. Foglebock—For his contribution to the CAPRI Program in developing a technique of forward gain control of extreme linearity with minimal adjustments for solid-state receivers.

J. L. Sullivan—For his contributions to the AN/FPS-95 radar proposal. His direct responsibility included receiver, signal processor, exciter, display, and radar control subsystems, over 70% of the total low-power electronic equipment proposed.

O. M. Woodward—For technical contributions to a new configuration feed system and reflector, for participation in developing a computer program for the synthesis and analysis of doubly curved reflectors, and for contribution to the modification of the TRADEX antenna feed to incorporate VHF without disturbing existing UFH and L-band performance.

ABOUT THIS ISSUE

Our thanks go to **L. Carmen, K. A. Chittick,** and **C. Hoyt** for their efforts in the early planning, expediting, and coordination of much of this issue of the RCA ENGINEER.—*The Editors*

CSD ANNOUNCES REORGANIZATION OF ENGINEERING AND MARKETING

J. M. Hertzberg, Vice President and General Manager of Communication Systems Division, announced a major reorganization of the Division in February. Under the reorganization, **David Shore,** Chief Engineer, heads the Engineering Department and **W. B. Kirkpatrick** directs the activities of the Marketing Department. Mr. Hertzberg stated that the objective of the reorganization was "to structure the most effective organization to: a) carry out existing commitments—provide organizational motivation; b) obtain new business, including both add-on and new jobs; and c) balance the use of resources and capitalize on our strengths."

The reorganized Engineering Department includes Project Implementation Teams to perform the job, an Advanced Communications Technology Center to provide the new technology, a Product Integration Engineering Center to furnish engineering support, and a Communications Systems Center to provide the technical plan for the Division's growth.

The Project Implementation Teams are divided into four product-line areas: *light communications, heavy communications, digital communications,* and *recording and TV equipment.* Each product-line area is headed by a manager with two deputy managers, one for technical control and the other for business control.

Mr. Shore, with **C. K. Law** as Operations Manager, will have the following staff: **R. Guenther,** Advanced Communications Technology; **K. K. Miller,** Communications Systems; **J. B. Cecil,** Engineering Administration; **R. Trachtenberg,** Product Integration Engineering; **W. B. Harris,** Light Communications Equipment; **E. C. Kalkman,** Heavy Communications Equipment; **T. L. Genetta,** Digital Communications Equipment; **E. Hudes,** Recording and Television Equipment; and **C. G. Arnold,** Staff Engineer to Chief Engineer.

The Marketing Department is reorganized to reflect the same product line areas as Engineering. Each Marketing product line manager is charged with working entirely with his Engineering counterpart on product line expansion and development.—*C. W. Fields*

DEGREES GRANTED

T. S. Kupfrian, DEP-ASD	MSEE, Mass. Institute of Tech.
J. Manning, DEP-ASD	MSME, Mass. Institute of Tech.
J. S. Pepi, DEP-ASD	MSME, Mass. Institute of Tech.
R. Tuft, DEP-ASD	MSEE, Mass. Institute of Tech.
M. A. Savrin, Corp. Standardizing	MBA, Temple University

PROFESSIONAL ACTIVITIES

ASD, Burlington, Mass.: **A. A. Clark** was elected to Sigma Xi by the Case Institute of Technology chapter in 1965. **T. S. Kupfrian, J. Manning, J. S. Pepi,** and **R. Tuft** have been elected to Sigma Xi by the Massachusetts Institute of Technology chapter. **W. J. Gray,** Reliability and Standards Engineering, has been elected Secretary of the Reliability Group, Boston Section of the IEEE, and is Chairman, Arrangements and Facilities Committee for the Group. He is also a member of the Program Committee for the preparation of the 1966 Spring Seminar sponsored jointly by the USAF Electronics Systems Div. and the IEEE Boston Section Chapter on Reliability. **J. W. Vickroy** has been named Publicity Chairman of the IEEE Group on Engineering Management for 1966. **J. E. Mutty** is a member of the NTC Exhibits Committee, Prudential Center, Boston, and was nominated for Vice Chairman, Professional Group on Communications Technology, Boston Section of the IEEE for 1966-1967.—*D. Dobson*

BCD, Camden, N. J.: **R. H. Lee** has been named Associate Editor of the IEEE *Transactions on Engineering Writing and Speech.* **M. K. Wilder,** Chairman of the IEEE Philadelphia Section, participated in the planning of the IEEE International Communications Conference for 1966.—*D. G. Hymas*

RCA Laboratories, Princeton, N. J.: **Dr. Jan A. Rajchman,** director of Computer Research Laboratory, David Sarnoff Research Center, has been appointed to the Newark College of Engineering's Advisory Committee in Electrical Engineering.

PROMOTIONS

to Engineering Leader & Manager

As reported by your Personnel Activity during the past two months. Location and new supervisor appear in parentheses.

DEP Astro-Electronics Division

H. M. Gurk: from T35050 Ldr., Eng. to *D99957 Mgr., Imagery Systems Prog.* (R. E. Hogan, Pr.)
G. K. Martch: from T35001 Eng. to *V99962 Administrator,* Systems Engrg. (R. E. Hogan, Pr.)
D. E. McCandless: from T35050 Ldr., Eng. to *V99964 Adm.,* LEM Testing (C. S. Constantino, Pr.)
E. W. Mowle: from T35051 Ldr., Engrg. Systems Proj. to *V99961 Adm., Equip. Engrg.* (R. E. Hogan, Pr.)
E. R. Walthal: from T35050 Ldr., Eng. to *D95953 Mgr., Missions Analysis* (W. Manger, Pr.)

DEP Missile & Surface Radar Division

C. A. Spurling: from Ldr. to *Mgr., Pedestals & Structures* (J. P. Schwartz, Mrstn.)
E. Staiger: from Adm., Engrg. Oper. to *Mgr., Engrg. Adm.* (F. G. Adams, Mrstn.)
D. Flechtner: AA Engr. to *Ldr., Displays* (M. Korsen, Mrstn.)
D. Pruitt: AA Engr. to *Ldr., Low-Freq. Transmitter Design* (R. N. Casolaro, Mrstn.)
M. Weiss: from AA Engr. to *Ldr., Material Appl.* (I. D. Kruger, Mrstn.)
R. C. Lund: from A Engr. to *Ldr., Spacetrack O & M* (W. L. Hendry, Mrstn.)
W. F. Tester: from A Engr. to *Ldr. Field Support & Depot Oper.* (W. L. Hendry, Mrstn.)
E. B. Darrell: A Engr. to *Ldr. TRADEX O & M* (W. L. Hendry, Mrstn.)
R. Adair: from Proj. Control Adm. to *Adm., Gen. Info. Processing* (E. Staiger, Mrstn.)

RCA Victor Home Instruments Division

I. M. Indiano: from Eng., Prod. D&D, C35721 to *Ldr., D&D Eng., C35700* (L. M. Krugman, Indpls.)

Electronic Components and Devices

R. M. Cullison: from Assoc. Eng., Mfg. to *Mgr., Quality Control* (A. G. Krause, Lanc.)
W. E. Harbaugh: from Sen. Eng., Prod. Dev. to *Engrgs. Ldr., Prod. Dev.* (F. G. Block, Lanc.)
S. B. Deal: from Sen. Eng., Prod. Dev. to *Engrg. Ldr., Prod. Dev.* (R. H. Zachariason, Lanc.)
J. A. Zollman: from Sen. Eng., Prod. Dev. to *Engrg. Ldr., Prod. Dev.* (R. H. Zachariason, Lanc.)
R. A. Donnelly: from Engrg. Ldr., Mfg. to *Mgr., Prod. Engrg.* (R. J. Hall, Findlay)
R. C. Reutter: from Eng., Mfg. to *Engrg. Ldr., Mfg.* (R. A. Donnelly, Findlay)
P. C. Baumann: from Mgr., Finished Goods Quality to *Mgr. Plant Engrg.* (R. J. Hall, Findlay)
J. P. Sasso: from Mgr., Prod. Engrg. to *Mgr. Mfg.* (N. A. Stegens, Cincinnati)
R. E. Ward: from Mgr.. Prod. Engrg. Miniature Tubes to *Mgr. Prod. Engrg.* (P. Sasso, Cincinnati)
J. J. Hanners: from Eng., Mfg. to *Mgr., Prod. Engrg.* (D. Ward, Cincinnati)
R. McFarlane: from Eng., Prod. Dev. to *Mgr., T/E Prod.* (L. Caprarola, Harrison)
R. Buttle: from Eng. to *Engrg. Ldr.* (M. Ammenwerth, Harrison)

DEP Aerospace Systems Division

R. C. Kee: from Senior Proj. Member to *Ldr.* (H. H. Knubbe, Mgr., Burl.)
E. H. Miller: from Staff Engrg. Sci. to *Mgr., Prod. Design* (E. A. Williams, Burl.)
A. Eliopoulos: from Sen. Member, Tech. Staff to *Ldr. Tech. Staff* (G. T. Ross, Burl.)
W. E. Hatfield: from Sen. Proj. Member, Tech. Staff to *Ldr. Tech. Staff* (D. K. Gilbertson, Burl.)
E. M. Stockton: from Ldr. Tech. Staff to *Mgr. Prod. Design* (E. B. Galton, Burl.)
B. T. Joyce: from Proj. Eng. to *Mgr., Computer Equip.* (E. A. Williams, Burl.)
A. Orenberg: from Staff Eng. Sci. to *Mgr., Prod. Design* (E. B. Galton, Burl.)
D. Hamblen: from Sr. Proj. Member, Tech. Staff to *Ldr., Tech. Staff* (L. H. Andrade, Burl.)
W. G. Wong: from Sr. Proj. Member, Tech. Staff to *Ldr. Tech. Staff* (A. Orenberg, Burl.)
R. N. Knox: from Sr. Proj. Member, Tech. Staff to *Ldr., Tech. Staff* (P. M. Toscano, Burl.)
A. Baker: from Staff Eng. to *Mgr., Systems Analysis* (S. S. Kolodkin, Burl.)
R. J. McNaughton: from Sr. Proj. Member, Tech. Staff to *Ldr. Tech. Staff* (H. Brodie, Burl.)
H. Brodie: from Ldr., Tech. Staff to *Mgr., Programming* (E. A. Williams, Burl.)
N. Meliones: from Engrg. Sci. to *Ldr., Tech. Staff* (J. H. Woodward, Burl.)

PROMOTIONS, Cont.

DEP West Coast Division

T. Skelly: from Sr. Member, D&D Engrg. Staff to *Ldr. D&D Engrg. Staff* (F. Worth, Van Nuys)

Broadcast and Communications Division

W. R. Murphy: from Engr. D&D to *Ldr. D&D Eng.* (R. G. Walz, Camden)

Electronic Data Processing

J. Clarke: from Ldr. Tech. Staff to *Mgr., Power Systems & Packaging* (H. N. Morris, Fla.)

R. Spoelstra: from Ldr. Tech. Staff to *Mgr., Logic & Peripheral Design Engrg.* (H. N. Morris, Fla.)

B. I. Kessler: from Ldr., D&D Eng. to *Mgr., Common Design Engrg.* (J. N. Marshall, Camden)

DEP Communications Systems Division

C. Arnold: from Mgr. Engrg. to *Staff Engr.* (C. K. Law, Camden)

M. E. Malchow: from Ldr. Tech. Staff to *Mgr. Integ. Analog Sub-Systems* (E. E. Moore, Camden)

RCA Service Company

Z. S. Hauser: from Ldr., Eng. to *Mgr., Site Engrg.* (O. E. Cole, Thule, Greenland)

R. A. Rose: from Install. & Mod. Eng. to *Mgr., Satellite Tracking* (H. L. Chadderton, Cherry Hill)

J. F. Linsalata: from Eng. to *Ldr., Eng.* (P. J. Barnett, Cocoa, Fla.)

ERRATA

In the previous issue, Volume 11-5, two diagrams were transposed in printing. In the paper "Lunar Excursion Module Radar Receivers," the functional block diagrams of the radar receiver and the transponder on pages 47 and 48 should be transposed. Also, the photographs of authors J. B. Friedenberg and P. Marcellos were inadvertently transposed on page 20 of the article, "LEM Electronics Reliability."

LICENSED ENGINEERS

W. T. Ackermann, ECD, Hr., PE-13007, N. J.

R. L. Boertzel, M&SR, PE-14398, N. J.

L. Campbell, ECD, Hr., PE-12958, N. J.

F. E. Coffey, ECD, Hr., PE-16381, Pa.

H. Eichhorn, ECD, Hr., PE-13486, N. J.

A. Fischer, ECD, Hr., PE-14533, N. J.

J. R. Hays, ECD, Hr., PE-13737

C. Lindsley, ECD, Hr., PE-13355, N. J.

S. N. Nasto, ECD, Hr., PE-13493, N. J.

E. Rudolph, ECD, Hr., PE-13349, N. J.

G. Samuels, ECD, Hr., PE-13506, N. J.

F. E. Scheline, ECD, Hr., PE-14184, N. J.

J. C. Schnitzius, M&SR, PE-14604, N. J.

M. Tuttle, ECD, Hr., PE-12794, N. J.

H. Waltke, ECD, Hr., PE-14483, N. J.

K. Wroblewski, ECD, Hr., PE-14482, N. J.

R. M. Dombrosky, RCA Ser. Co., PE-14495, N. J.

RCA LABS ANNOUNCE ANNUAL ACHIEVEMENT AWARDS

RCA Laboratories announced in March the names of the winners of the Annual Achievement Awards for 1965. This is the 19th year that these awards have been granted. Recipients include:

W. H. Barkow—For continuing contributions to television deflection systems.

M. D. Coutts—For ingenious applications of electron diffraction and microscopy in characterizing electronically active solids, and for demonstrating the resolution capabilities of the RCA electron microscope.

R. E. Enstrom—For materials research leading to significant improvements in the prepartion of superconducting ribbon for solenoid magnets.

G. H. Heilmeier—For ingenuity and aggressiveness in the application of liquid crystal phenomena.

S. Y. Levy—For the conception of machine organizations suited to the technology of large integrated arrays.

A. Presser—For the development of tunnel-diode amplifiers with large dynamic range.

R. L. Rosenfeld—For the conception, design, and implementation of a programming language for the automatic coordinatograph.

R. J. Ryan—For the conception and demonstration of novel techniques in constructing built-up interconnection boards.

A. R. Sass—For research leading to the demonstration of memory systems using structured cells.

A. I. Stoller—For research leading to new methods of fabricating integrated circuits.

H. Wielicki—For the discovery and development of reversal type liquid and hot-melt toners for the Electrofax process.

R. Williams—For research that provided a basic technique for investigating the stability of oxide layers on silicon, and for determination of the energy levels at the oxide-silicon interface.

P. Berger, J. R. Haupt, H. N. Garver, and **S. Skillman**—For team performance in the development of new concepts in management science.

L. J. Berton, J. T. O'Neil, Jr., H. E. Kulsrud, and **T. M. Stiller**—For team performance in simulation of the Spectra 70 computers, prior to their construction, by programs for the RCA 601.

J. R. Burns, J. J. Gibson, A. Harel, K. C. Hu, and **R. A. Powlus**—For team performance in the conception and fabrication of computer memories made of arrays of insulated-gate field-effect transistors.

J. P. Dismukes, L. Ekstrom, D. F. Martin, and **C. C. Wang**—For team performance in the synthesis and evaluation of a superior acicular magnetic material for information storage.

E. C. Giaimo, H. C. Gillespie, and **R. G. Olden**—For team performance in developing an Electrofax transfer process for making multiple copies.

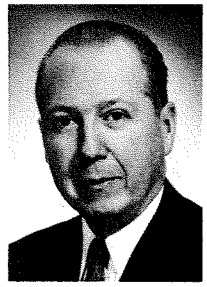
P. V. Goedertier and **R. J. Pressley**—For team performance in the development of an efficient, high-power, optically pumped laser.

R. Hirota, K. Suzuki, and **M. Toda**—For team performance in devising a new class of microwave devices useful over a large range of frequencies for control of microwave and millimeter wave power.

R. J. Ikola and **L. S. Napoli**—For team performance in the demonstration of microwave amplification in avalanche diodes.

G. L. Kasyk and **M. Pradervand**—For team performance in the development of an acoustical-mechanical disc recorder.

A. H. Simon and **D. A. Watters**—For team performance in the design and implementation of RCA SNOBOL, a symbolic manipulation language for computer programming.



L. F. Jones

K. P. Haywood

NEW EDITORIAL REPRESENTATIVES

Two new representatives have been named to the Editorial Board of the RCA ENGINEER. **Loren F. Jones** will serve as the representative for New Business Programs, replacing **N. Amberg. Kenneth P. Haywood** replaces **Ed Stanko,** who retired in March, as the ed rep for Technical Products, RCA Service Company.

Loren F. Jones graduated from Washington University, St. Louis, Mo., and attended Stanford University's Graduate School of Business. After joining RCA in 1930, he undertook engineering assignments in several European countries. On his return to Camden, N. J., he was active in broadcast and associated fields. In 1937-1938 he represented RCA in several nonmilitary engineering projects in Russia. During World War II, he represented RCA in Government R&D projects. He served as chairman of the Direction Finder Committee and as a member of the Communications and Radar Divisions of the Office of Scientific Research and Development, for which he was awarded the Presidential Certificate of Merit. In 1951 he established and organized the New Products Division of RCA. As a result of one of the programs of this Division, RCA established electronic data processing as a new business. Mr. Jones presently is Manager, Engineering, New Business Programs.

Kenneth P. Haywood received his BSEE at North Carolina State College in 1929. Following graduation he was employed by General Electric for 2 years in radio transmitter development. In 1931 he joined RCA Manufacturing Co. and was assigned to test equipment maintenance and calibration. He transferred to the RCA Victor Service Department in 1935, and for 6 years provided field service for theatre sound, centralized sound, and mobile equipment. Mr. Haywood was promoted to Philadelphia District Manager in 1941. Since transferring to the Home Office of RCA Service Company in 1944, he has served successively as Coordinator of Special Services, Technical Products Division; Administrator, EDP Services; Manager, EDPS Administration; Manager, Facilities, Field Support Services; and Manager, Administration, Technical Products Service. He is presently Manager, Administration and Technical Support, Technical Products Division.

1965 RCA TECHNICAL PAPERS INDEX

The annual *Index to 1965 RCA Technical Papers* was mailed with the February-March 1966 RCA ENGINEER to the entire RCA technical staff. The Index categorizes by subject and author some 1200 items of nonproprietary technical literature authored by RCA scientists and engineers during 1965—including all published and presented RCA papers, theses, technical notes, etc. Its scope includes comprehensive coverage of RCA papers in professional society and other outside media, as well as those in the RCA ENGINEER, RCA REVIEW, etc. Also included is a separate list of over 300 U.S. Patents issued to RCA personnel in 1965. The Index presents complete instructions on how to obtain copies of any of the papers. Comments concerning the 1965 Index should be addressed to Ed Jennings, RCA Staff Technical Publications, Bldg. 2-8, Camden, N.J. (PC-3396).

INDEX TO RCA ENGINEER VOLUME 11

This index covers Vol. **11-1** (June-July 1965), **11-2** (Aug.-Sept. 1965), **11-3** (Oct.-Nov. 1965), **11-4** (Dec. 1965-Jan. 1966), **11-5** (Feb.-Mar. 1966), and the present issue, **11-6** (April-May 1966). Since RCA ENGINEER articles are also available as reprints, the reprint number **(PE-000)** is noted throughout. RCA ENGINEER papers are also indexed along with all other papers written by the RCA technical staff in the annual "Index to RCA Technical Papers."

SUBJECT INDEX

Titles of papers are permuted where necessary to bring significant keyword(s) to the left for easier scanning. Authors' division appears parenthetically after his name.

Spectra 70—An Introduction to Basic Design and Philosophy of Operation—A. D. Beard (EDP, Cam.) **11-3**, (in reprint booklets *RCA Advances in Computer Technology, PE-257* and *Computer Progress, PE-258*)

CONTROL SYSTEMS (& automation)

VE-DET: Electronic Vehicle Detector for Automated Traffic Control—E. C. Donald (BCD, Plymouth); **10-5**, (reprint *PE-226*)

DISPLAYS

DIVCON—A Digital-to-Video Conversion System—R. J. Clark (RCA Ltd., Montreal); **11-3**, (reprint *PE-255*)

High-Speed Film Processor and Quick-Access Display System—E. Schwartz, A. A. Blatz (DEP-MSR, Mrstn.); *E&R Note*, **11-5**, (reprint *PE-264-39*)

Selection of Letter Size and Lighting for Display and Control Panel Legends—W. Baskerville, R. N. Gillis (DEP-SEER, Mrstn.); **10-6**, (in reprint booklet *Life Sciences, PE-233*)

DOCUMENTATION

Abbreviations for Technical Terms, New (IEEE) Standard—(R&E Staff, Cam.) **11-1**, (reprint *PE-248*)

Books Looking for Authors (The Engineer and the Corporation series)—C. W. Fields (DEP-CSD, Cam.) **11-2**, (reprint *PE-264-9*)

New Approach to Writing Operator Procedure Manuals—Dr. F. H. Ireland, T. G. Wiedman (DEP-SEER, Mrstn.); **10-6**, (in reprint booklet *Life Sciences, PE-233*)

Scientific Computer Applications Program (SCAPC)—Dr. J. Kurshan (Labs, Pr.); **11-6**, (reprint *PE-279*)

Technical Papers and Their RCA Approval (The Engineer and the Corporation series)—W. O. Hadlock (R&E Staff, Cam.) **11-1**, (reprint *PE-264-1*)

ELECTROMAGNETIC WAVES (theory & phenomena)

Guided Waves in Gaseous Plasmas—Dr. T. W. Johnston (RCA Lt., Montreal); **11-4**, (in reprint booklet *Plasma Research, PE-263*)

Ionized Rocket Exhausts, Interaction of Electromagnetic Waves with—A. Boornard (DEP-AppRes, Cam.) **11-4**, (reprint *PE-264-26*)

HF Signal Statistics; Short-Distance Sky-Wave Rapid-Fading—S. Krevsky (DEP-CSD, N.Y.) *E&R Note*, **11-2**, (reprint *PE-264-17*)

Microwave Emission from Solid-State Plasmas—Dr. R. D. Larrabee, W. A. Hicinbothem, Jr., J. J. Thomas (Labs, Pr.) **11-4**, (in reprint booklet *Plasma Research, PE-263*)

Propagation in Solid-State-Plasma Waveguides—Nonreciprocal Devices—Dr. R. Hirota, M. Toda (Labs, Tokyo) **11-4**, (in reprint booklet **Plasma Research, P-263**)

Resonances in the Ionosphere Magnetoplasma—T. W. Johnston (RCA Ltd, Montreal) Second AAS Symp. on the Interactions of Space Vehicles with an Ionized Atmosphere, Miami, Fla., Nov. 27, 1965

Sunspot HF Disturbance Forecasting Based on Planet Positions—J. H. Nelson (RCA Comm., N.Y.) **11-1**, (reprint *PE-264-3*)

ELECTROMECHANICAL CONVERTORS

Gas-Leak Valve, Electrical Remote-Controlled—M. E. Moi (Labs, Pr.) *E&R Note*, **11-3**, (reprint *PE-264-22*)

ELECTRO-OPTICS (systems & techniques)

Electro-Optical Signal Processing—M. J. Cantella, R. Kee (DEP-ASD, Burl.); **10-5**, (in reprint booklets *Electro-Optics, PE-211, PE-223*)

Infrared Fiber Optics—L. Arlan, N. Aron (DEP-ASD, Burl.); **10-5**, (in reprint booklets *Electro-Optics, PE-211, PE-223*)

ENERGY CONVERSION (& power sources)

Electrical Energy Sources, The Revolution in—A State-of-the-Art Review—P. Rappaport (Labs, Pr.) **11-2**, (reprint *PE-264-12*)

Thermoelectric Water Reclamation for Manned Space Systems—P. E. Wright (DEP-AppRes., Camden); **10-6**, (in reprint booklet *Space Electronics, PE-227*)

FILTERS, ELECTRIC

Filter Design by Digital Computer—T. G. Marshall, Jr. (BCD, Camden); E&R Note, **10-5** (reprint *PE-235-36*)

Three-Resonator Filters, General—R. M. Kurzrok (DEP-CSD, N.Y.) *E&R Note*, **11-4**, (reprint *PE-264-30*)

GEOPHYSICS

Geophysical Phenomena, Laboratory Simulation of—Dr. F. J. F. Osborne, Dr. M. P. Bachynski, J. V. Gore, M. A. Kasha (RCA Ltd., Montreal); **11-4**, (in reprint booklet *Plasma Research, PE-263*)

Resonances in the Ionosphere Magnetoplasma—Dr. T. W. Johnston, Dr. I. P. Shkarofsky (RCA Ltd., Montreal); **11-4**, (in reprint booklet *Plasma Research PE-263*)

Sunspot HF Disturbance Forecasting Based on Planet Positions—H. J. Nelson (RCA Comm., N.Y.) **11-1**, (reprint *PE-264-3*)

GRAPHIC ARTS

High-Speed Film Processor and Quick-Access Display System—E. Schwartz, A. A. Blatz (DEP-MSR, Mrstn.); *E&R Note*, **11-5**, (reprint *PE-264-39*)

Pattern Generation Under Numerical Control—Dr. R. Rosenfeld (Labs, Pr.); *E&R Note*, **11-5**, (reprint *PE-264-41*)

HUMAN FACTORS (physiological & psychological)

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Human Factors Engineering—More Than Just the "Blessing"—Dr. H. B. Matty (DEP-CSD, Camden); **10-6**, (in reprint booklet *Life Sciences, PE-233*)

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(LEM) System Considerations for Landing a Man on the Moon—Dr. L. E. Matson (DEP-ASD, Burl.); **11-5**, (in reprint booklet *The LEM Program at RCA, PE-281*)

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Value Engineering, the Practice of—K. Stoll (DEP-AED, Pr.) 17th Annual Conf., American Soc. for Quality Control, Rutgers Univ., New Brunswick, N. J., Sept. 11, 1965

MEDICAL ELECTRONICS

Biomedical Engineering—Dr. A. N. Goldsmith (RCA Staff, N.Y.); **10-6**, (reprint *PE-225*)

Electronics Technology in Medicine—L. E. Flory (DEP-AED, Pr.); **10-6**, (reprint *PE-230*)

Microcircuit-Microwatt Design Techniques for New Internal Medical Sensors—L. E. Flory, F. Hatke (DEP-AED, Pr.); **10-6**, (reprint *PE-231*)

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Frequency Multiplier for Missiles, Miniature C-Band Solid-State—D. E. Nelson, J. J. Napoleon, C. L. Cuccia (ECD, Pr.); **10-5**, (in reprint booklet *Microwave Systems and Devices, PE-210*; also in reprint booklets *Microwave Components, PE-215, PE-219*)

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Colorimetry and Performance of Color TV Phosphor of Color TV Phosphor Screens—A. E. Hardy (ECD, Lanc.) **11-2**, (in reprint booklets *Color Television and Tube Engineering, PE-253* and *Color Picture Tubes, PE-254*)

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Frequency Multipliers for the LEM Radars, Design of Solid-State—E. Bliss, M. Fromer (ECD, Hr.); **11-5**, (in reprint booklet *The LEM Program at RCA, PE-281*)

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LEM Rendezvous Radar and Transponder—Dr. W. C. Curtis, L. B. Wooten (DEP-ASD, Burl.); **11-5**, (in reprint booklet *The LEM Program at RCA, PE-281*)

LEM Rendezvous Radar and Transponder, Packaging the—J. I. Herzlinger, M. J. Kurina, A. W. Sinkinson (DEP-ASD, Burl.); **11-5**, (in reprint booklet *The LEM Program at RCA, PE-281*)

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Space Track Program 4102-S—E. T. Garner, J. Oseas (DEP-AED, Pr.); **10-5**, (in reprint booklet *Space Electronics, PE-227*)

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Plasma Produced by High-Altitude Nuclear Detonations—Effects on Radar of—Dr. K. Sittel (DEP-SEER, Mrstn.) **11-4**, (reprint *PE-264-27*)

Space Radiation—Its Influence on Satellite Design—Dr. A. G. Holmes-Siedle (DEP-AED, Pr.) **11-1**, (reprint *PE-239*; also in reprint booklet *Space Electronics, PE-227*)

RADIO BROADCASTING
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Wireless (Broadcast) Microphone System Using an Ultraviolet-Light Carrier, A Developmental—J. L. Hathaway (NBC, N.Y.) **11-1**, (reprint *PE-264-7*)

RADIO RECEIVERS
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Design of the RC-1218 Solid-State AM-FM Stereo-Multiplex Hi-Fi Receiver—J. B. Schultz (H.I., Indpls.); **10-5**, (in reprint booklets *Home Instruments, PE-228, PE-229*)

Ferrite Bead Application to AM Transistorized Receiver—E. Diamond, M. Mehta (HI, Indpls.); *E&R Note*, **11-6**, (reprint *PE-264-50*)

Integrated Circuits in Television and Radio Receivers—J. Avins (HI, Indpls.); **11-6**, (in reprint booklet *Television Receivers, Related Circuits and Devices, PE-276*)

Overload Considerations for Low-Cost Transistorized AM Receivers—D. E. Kowallek, M. C. Mehta (H.I., Indpls.); **10-5**, (in reprint booklets *Home Instruments, PE-228, PE-229*)

RELIABILITY
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LEM Electronics Reliability—J. B. Friedenberg, J. J. Landers, P. Marcellos (DEP-ASD, Burl.); **11-5**, (in reprint booklet *The LEM Program at RCA, PE-281*)

SOLID-STATE DEVICES

(Transistor, Overlay): High-Power VHF Overlay Transistor for Single Sideband Applications—R. Rosenzweig, Z. F. Chang (ECD, Som.); *E&R Note*, **11-6**, (reprint *PE-264-46*)

(Transistor, Overlay): The 2N4012—A 2.5 Watt, 1 Gc/s, "Overlay" Transistor—H. C. Lee, G. J. Gilbert (ECD, Som.); **11-6**, (in reprint booklet *Television Receivers, Related Circuits and Devices, PE-276*)

Transistors for Television—R. A. Santeilli, H. A. Thanos (ECD, Som.); **11-6**, (in reprint booklet *Television Receivers, Related Circuits and Devices, PE-276*)

Transistors (RF Power)—Important New Advances in Performance and Economics—Dr. D. J. Donahue, B. A. Jacoby (ECD, Som.) **11-1**, (reprint *PE-246*)

(Transistors): The "Overlay"—A New UHF Power Transistor—D. R. Carley, P. L. McGeough, J. F. O'Brien (ECD, Som.) **11-1**, (reprint *PE-247*)

SPACE COMMUNICATION
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Digital Versus Analog Television Transmission from Spacecraft—J. W. Whelan (DEP-AED, Pr.); **11-5**, (reprint *PE-264-34*)

Erectable Antenna for Space Communications—R. J. Mason, F. S. Wezner (DEP-MSR, Mrstn.); **11-5**, (in reprint booklet *The LEM Program at RCA, PE-281*)

GEMINI Spacecraft PCM-FM Telemetry Transmitter—R. G. Erdmann (DEP-CSD, Camden); **10-5**, (in reprint booklet *Space Electronics, PE-227*)

Ionized Rocket Exhausts, Interaction of Electromagnetic Waves with—A. Boornard (DEP-AppRes, Cam.) **11-4**, (reprint *PE-264-26*)

(LEM) Man-In-Space Communications Subsystem—R. E. Davis (DEP-ASD, Burl.); **11-5**, (in reprint booklet *The LEM Program at RCA, PE-281*)

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Life Beyond the Atmosphere—Origins, Detection, and Support—Dr. A. G. Holmes-Siedle (DEP-AED, Pr.); **10-6**, (in reprint booklet *Space Electronics, PE-227*)

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Cassegrainian Monopulse Tracking Antenna for Space Rendezvous—W. W. Carter, W. E. Powell (DEP-MSR, Mrstn.); **11-5**, (in reprint booklet *The LEM Program at RCA, PE-281*)

Frequency Multipliers for the LEM Radars, Design of Solid-State—E. Bliss, M. Fromer (ECD, Hr.); **11-5**, (in reprint booklet *The LEM Program at RCA, PE-281*)

LEM ATCA and DECA, Packaging the—S. Waldstein, P. J. Sacramone (DEP-ASD, Burl.); **11-5**, (in reprint booklet *The LEM Program at RCA, PE-281*)

(LEM) Attitude, Translation, and Descent Engine Control of LEM—S. S. Kolodkin, P. T. Frawley (DEP-ASD, Burl.); **11-5**, (in reprint booklet *The LEM Program at RCA, PE-281*)

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LEM Rendezvous Radar and Transponder, Packaging the—J. I. Herzlinger, M. J. Kurina, A. W. Sinkinson (DEP-ASD, Burl.); **11-5**, (in reprint booklet *The LEM Program at RCA, PE-281*)

Satellite Orbits, Twenty-Four-Hour—M. I. Aitel (DEP-MSR, Mrstn.); **11-5**, (reprint *PE-272*)

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(LEM) Attitude, Translation, and Descent Engine Control of LEM—S. S. Kolodkin, P. T. Frawley (DEP-ASD, Burl.); **11-5**, (in reprint booklet *The LEM Program at RCA, PE-281*)

LEM Program, History and Management of the—F. J. Gardiner (DEP-ASD, Burl.); **11-5**, (in reprint booklet *The LEM Program at RCA, PE-281*)

(LEM) System Considerations for Landing a Man on the Moon—Dr. L. E. Matson (DEP-ASD, Burl.); **11-5**, (in reprint booklet *The LEM Program at RCA, PE-281*)

Proposed Satellite for TV Observation of Zero-G Effects on Development of the Opossum Fetus—S. H. Fairweather, J. E. Mortimer (DEP-AED, Pr.); **10-6**, (in reprint booklet *Space Electronics, PE-227*)

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LEM Electronics Reliability—J. B. Friedenberg, J. J. Landers, P. Marcellos (DEP-ASD, Burl.); **11-5**, (in reprint booklet *The LEM Program at RCA, PE-281*)

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Thermoelectric Water Reclamation for Manned Space Systems—P. E. Wright (DEP-App Res., Camden); **10-6**, (in reprint booklet *Space Electronics, PE-227*)

TV Camera System for the Command Module of the APOLLO Spacecraft—P. M. Zucchino, J. L. Lowrance (DEP-AED, Pr.); **10-5**, (in reprint booklet *Space Electronics, PE-227*)

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Research in Superconductivity at RCA Laboratories—Dr. F. D. Rosi (Labs, Pr.) **11-1**, (reprint *PE-249*)

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AGC and Gamma Control Amplifiers for TV Broadcast, New Transistorized—V. J. Duke (NBC, N.Y.) **11-2**, (reprint *PE-264-11*)

Color TV Film Camera, The New, Transistorized TK-27—D. M. Taylor (BCD, Camden); **10-5**, (reprint *PE-224*)

Electron Image Magnification in Broadcast TV Cameras Simulates Long-Focal-Length Lenses—F. Himelfarb (NBC, N.Y.); **10-5** (reprint *PE-235-34*)

Parallel Operation of a 50-kW Television Transmitter—*R. E. Winn* (BCD, Gibbsboro) **11-2**, (reprint *PE-264-10*)

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Wireless Broadcast Microphone System Using an Ultraviolet-Light Carrier, A Developmental—J. L. Hathaway (NBC, N.Y.) **11-1**, (reprint *PE-264-7*)

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Digital Versus Analog Television Transmission from Spacecraft—J. W. Whelan (DEP-AED, Pr.); **11-5**, (reprint *PE-264-34*)

Laser Voice and Video Communications—Applying the GaAs Room-Temperature Injection Laser and the GaAs Electro-Optic Crystal—W. J. Hannan, Dr. J. Bordogna (DEP-App-Res, Cam.) **11-1**, (reprint *PE-264-6*)

Scan Converter for Apollo Television—W. G. Gibson (DEP-AED, Pr.); **11-6**, (reprint *PE-264-44*)

Sensor Operating Parameters for Unattended Camera Systems, Regulation of—E. Vaerewych, H. Wittlinger (DEP-AED, Pr.); E&R Note, **10-5** (reprint *PE-235-37*)

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TV Camera System for the Command Module of the APOLLO Spacecraft—P. M. Zucchino, J. L. Lowrance (DEP-AED, Pr.); **10-5**, (in reprint booklet *Space Electronics, PE-227*)

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