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OUR COVER

. . . represents RCA effort in both microwave systems and devices. Top photo: The new CW-60 solid-state microwave relay equipment. A completed unit stands at left in the Camden plant, with several units undergoing assembly to the right; H. S. Wilson (left), Mgr., Microwave Engineering, and E. J. Forbes (right), Ldr., Microwave Communications Engineering, Broadcast and Communications Products Division, Camden, look on. Bottom photo: In a screen room at the Microwave Tube Operations Dept., Electronic Components and Devices, Harrison, one of the solid-state microwave components for the Lunar Excursion Module is tested. Frequency multipliers for rendezvous and landing radars and a transponder for the moon project are being designed and fabricated in Harrison. H. K. Jenny, (right), Mgr., Engineering, MTOD, looks on as E. Bliss, Senior Engineer, demonstrates test procedure. (Cover art direction: J. Parvin. Photography, R. Allen.)

Microwave—The World of Challenge

The past decade has seen a very rapid growth in microwave tubes, devices, an systems. Wide application in the defense effort has caused large sums of money t be invested in meeting these needs. The microwave engineer has had a consta challenge to increase the frequency, up the power, broaden the bandwidth, impro the frequency agility, and to make other demanding technical breakthroughs. Th emphasis has been on technical results; costs, if reasonable, were generally accep able.

The last two years have produced a radical change for the microwave enginee When defense budgets were curtailed, microwave programs were heavily influence Competition became intense. The emphasis now is on minimum cost, but still mee ing the same demanding technical performance. Keen competitive struggle is th order of the day, and no let-up is in sight.

Solid-state devices, long the challenge of the conventional tube and systems eng neer, are moving rapidly into microwave devices and systems; so again, the worl of challenge moves rapidly for the microwave engineer. Additionally, as the growt of electronics continues, the use of microwaves will become more and more commo place. This growth will subject the microwave technology to the same problems o other business areas—change, competition, and challenge. The microwave enginee now as never before, must meet these normal business pressures. From my exper ence with engineers in the microwave field, they too will meet these challenges i an outstanding manner.

C. E. Burnet
Division Vice President and General Manage
Industrial Tube and Semiconductor Divisio
Electronic Components and Devic
Radio Corporation of Americ

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• 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.

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Background — Why IRL?

RCA shares in the ownership and use of a 5-Mw nuclear research reactor located near RCA Laboratories. Called Industrial Reactor Laboratories, Inc. (IRL), it is available for RCA research and development programs requiring a source of nuclear radiation in the study of materials, components, or systems. The swimming-pool-type reactor is complemented by hot cells with remote manipulators, sensitive radiation counters, and other specialized laboratory equipment.

RCA Laboratories coordinates RCA's use of the facility, but engineers and scientists from other RCA locations also undertake projects at IRL. As this paper describes, such work has included radiotracer studies for following the migration and diffusion of materials; radiation damage to tunnel diodes, transistors, fiber optics, vacuum tubes and thermoelectric materials; analysis by neutron activation and neutron diffraction; improvement of laser crystals and superconductors by irradiation.

RCA's participation in Industrial Reactor Laboratories offers RCA engineers and scientists a unique opportunity to use nuclear radiation in furthering their technical objectives. Other facilities, both government and commercial, are available, but the advantages of an in-house facility often spell the difference between doing or avoiding a particular experiment. IRL is geographically convenient for many RCA locations; administrative red tape is a minimum; project costs are nominal; consultation and assistance are available; facilities are outstanding. Unclassified government contract work can be done at IRL provided the work represents a research use of the facilities.

This paper illustrates some of the many ways that IRL has been useful and can continue to be so. Even more important are the ingenious and profitable ways of using nuclear radiation that are still to be devised.

The Engineer and the Corporation

INDUSTRIAL REACTOR LABORATORIES

A Service Available to RCA

Dr. J. KURSHAN and Dr. D. A. ROSS
RCA Laboratories, Princeton, N. J.

Industrial Reactor Laboratories, Inc. (IRL) is a nuclear research reactor installation located in Plainsboro, N. J. and owned by RCA and nine other companies. It provides RCA with a facility where nuclear radiation can be used in the conduct of its research and development activities. RCA Laboratories administers the use of the IRL facilities for RCA and a number of groups throughout the Corporation have taken advantage of the services available.

The ten companies associated in IRL are as follows:

American Machine & Foundry Co.
The American Tobacco Co.
Atlas Powder Co.
Continental Can Co.
Corning Glass Works
National Distillers and Chemical Corp.
National Lead Co.
Radio Corporation of America
Socony Mobil Oil Co., Inc.
United States Rubber Co.

The facility was built and equipped by Turner Construction Company and American Machine & Foundry Company at a cost of $4,500,000. The reactor first went critical in January 1959 and full power operation of 5 Mw was reached in June 1959. Columbia University operates the installation for the participants. The operating staff are Columbia employees and the Director is a Professor of Chemical Engineering in the Division of Nuclear Science and Engineering. Each of the companies and Columbia have individual laboratory space and share in the use of the major experimental facilities. Policy is established by a Board of Directors having one member from each of the ten companies. Dr. J. Hillier, Vice President, RCA Laboratories, represents RCA on this body and is currently President of IRL. Several standing committees of the Board of Directors are responsible for advising it on IRL's activities. Dr. J. Kurshan represents RCA on the Operating Committee which is concerned with the technical aspects of running the facility. H. W. Leverenz, Associate Director, RCA Laboratories, was RCA's original Operating Committee member and served as Chairman during one of its critical formative years. RCA's representative on the Insurance and the Finance Committees is A. N. Curtiss, Manager Administration, RCA Laboratories.

RCA's interest in nuclear radiation lies in its use as a tool to study materials and devices employed in the transmission, reception, storage, and handling of information. Additional interest lies in its use as a source of power to operate such information systems. Specific applications of nuclear radiation of interest to RCA are:

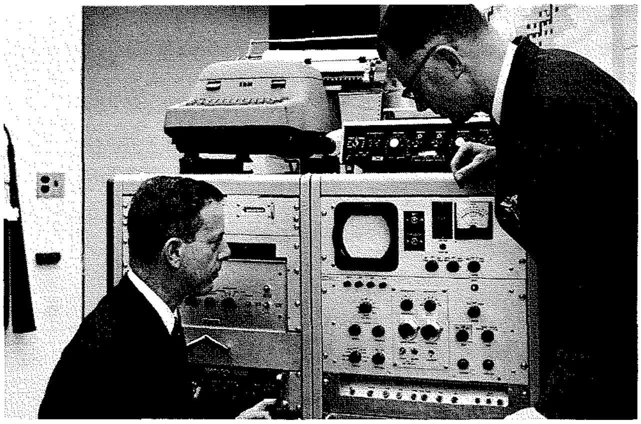
1) Changes in materials and devices caused by nuclear transmutation, photo-ionization, or radiation-produced defects due to knock-on collisions.
2) Activation of materials to produce isotopes for solid state tracer studies or the activation of impurities *in situ* to provide a means for impurity analysis in solids.
3) Development of nuclear instrumentation such as photomultipliers and solid state detectors.
4) Methods of energy conversion from radioisotopes, fission products decay, or other nuclear phenomena to electrical power by thermoelectric, thermionic and photovoltaic devices.
5) Maintaining an active nuclear program which will exploit new techniques which become available for beneficially altering electronic materials.

THE FACILITY

The Industrial Reactor Laboratories is a complete nuclear radiation research center. Its dominant architectural feature, shown in Fig. 1 is the aluminum-sheathed, concrete dome housing the reactor. The primary source of radiation is a 5-Mw swimming-pool-type research reactor. Fig. 2 is a picture of the reactor which is located under the dome. Thermal and fast neutrons are available from the core, and gamma rays without the neutrons can be obtained either from spent fuel elements at one side of the pool or from two cobalt-60 sources of 5,000 and 3,000 curies in separate hot cells.

The reactor core is fueled with completely enriched uranium-235 and moderated and cooled with demineralized light water. When the fuel elements were fresh, it required about 4,000 grams to go critical. The core is located near the bottom of the water-filled pool which is approximately 30 feet deep, 35 feet long, and 20 feet wide. The water serves as the biological shield, the coolant and the moderator to slow the fission neutrons down to thermal energies where the fission cross section is much higher than for the fast neutrons. Access to the core for experimental work is possible by suspending samples in the water near the core, through one of the six beam tubes set in the concrete pool wall, through a graphite-filled thermal column, or through

Final manuscript received October 3, 1964



The authors, J. Kurshan and D. A. Ross, examine a gamma ray spectrum on the 200-channel analyzer at IRL.

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 held the same position 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. Dr. Kurshan was named Mgr., Graduate Recruiting in January 1956; Mgr., Technical Recruiting and Training in June 1956; and Mgr., Employment and Training in January 1958. He was appointed to his present position, Mgr., Research Services Laboratory, in March 1959. Dr. Kurshan is a Senior Member of the IEEE, of which he is a Member of the Group on Engineering Management, a past member of the Education Committee, and of the Subcommittee on Solid State Devices, and a past Chairman of the Princeton Section. He is also a member of the American Physical Society, Phi Beta Kappa, Sigma Xi, Phi Kappa Phi, and Pi Mu Epsilon.

DR. DONALD A. ROSS received his BEEE from McGill University in 1947. He then worked at General Electric X-Ray Corporation, and in 1950 joined the High Voltage Engineering Corporation, where he worked on high-energy particle accelerators. From 1953 to 1957 he attended Yale University, receiving the MSc in 1955 and the PhD in 1957 in physics. While at Yale, he did research on radiation effects in biological and chemical systems, and in nuclear physics. Dr. Ross returned to the High Voltage Engineering Corporation in 1958 and was responsible for work in radiation applications. Dr. Ross joined RCA Laboratories in 1958 as Head of the Nuclear Reactor Group carrying on research in solid-state physics. Since 1962 he has been Manager, Corporate Graduate Recruiting. Dr. Ross is a member of the American Physical Society, Sigma Xi, and the Radiation Research Society.

two pneumatic systems. The latter blow sample containers (called "rabbits") through a tube to within an inch of a core face and then, after a preset time, blow the container back to a shielded receiver outside the pool wall (Fig. 3).

This system is used for short irradiations of up to an hour at a flux of about 2×10^{13} neutrons/cm^2/second. For longer irradiations or for large samples which will not fit into the containers used in the rabbit system, samples are encapsulated in aluminum cans and suspended in the pool next to the core face. Since aluminum has a short half-life, its radioactivity will die out relatively quickly after the irradiation. In addition, it is inert on submersion in the water in a high radiation environment.

The beam tubes are used for semi-permanent experimental installations. The shielding required is extensive and can be moved only when the reactor is shut down. RCA has equipped two beam tubes for experimental investigations of solid materials. One is used for neutron diffraction and the other to investigate changes in electrical properties of materials or devices during irradiation. In the latter case, it is easier to use a beam tube than to suspend the samples in the water since the electrical lead wires are shorter (12 feet versus 30 feet) and do not have to be sealed or insulated against the water environment.

To handle highly radioactive samples, hot cells are used to carry out remote manipulations. Three are available, each equipped with remote manipulators and windows of 3-foot-thick glass. The cells can handle several thousand curies of radioactivity. Two are used exclusively to house the cobalt-60 sources for gamma radiation. The other is for mechanical and chemical manipulation of samples (Fig. 4).

For neutron activation analysis the experimenter wishes to identify quantitatively the radioisotope of interest. A low-level counting room is available with a 400-channel gamma ray analyzer as well as a liquid scintillation counter for low energy beta counting. If the low level counting room with its especially low background is unnecessary, a mobile 200-channel analyzer is available which can be used in any of the laboratories where the experimenter has his own counting crystal and shield. RCA can measure very low activity samples with its 3-inch sodium iodide scintillation crystal housed in a 24-by-24-inch cave with 4-inch-thick walls made out of a special low background lead.

When used in conjunction with the mobile multichannel analyzer, this forms an extremely versatile counting system.

In addition to the facilities available for handling and counting radioactive materials, there are many experiments where the radiation causes changes in properties of materials or in reactions. In many cases the equipment needed to study these phenomena is large and expensive and can be shared by different users. As a result, several pieces of research equipment have been purchased jointly for use by any of the participants, including an electron spin resonance unit and a recording optical spectrometer.

GETTING WORK DONE

Columbia University employs an operating staff of 48 people to operate the reactor, to provide technical services including health physics, and to handle administrative matters. RCA makes use of their services and of the experimental facilities through its staff located at the IRL site. Dr. D. A. Ross was our first staff member in residence at IRL and remained there until 1962. At present R. F. Bailey, a radiochemist, is assigned to RCA's laboratory at IRL and provides liaison between other RCA personnel using the facility and the IRL operating staff. J. G. White is at IRL on a part-time basis. He conducts neutron diffraction studies using the neutron spectrometer that RCA built and he serves as RCA's alternate on the IRL Operating Committee.

Organizationally, RCA Laboratories coordinates technical

Fig. 1—Dome housing the reactor is 87 feet high and made of re-enforced concrete 12 inches thick sheathed with aluminum.

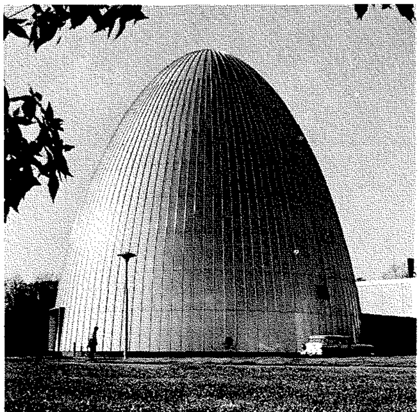


Fig. 2—Reactor core suspended from the instrument bridge which straddles the pool walls.

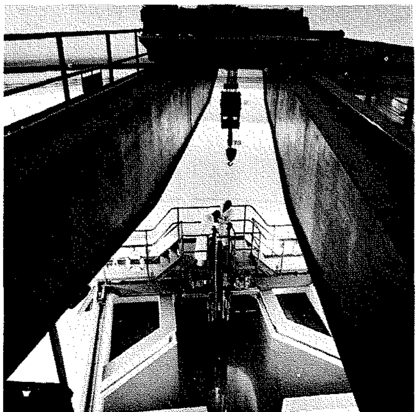
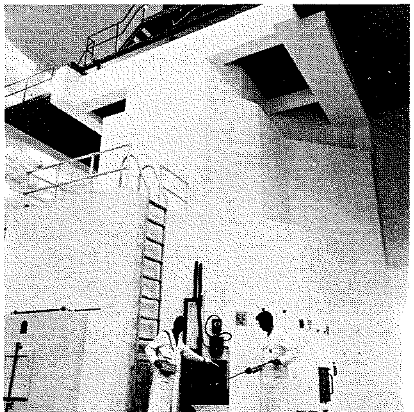


Fig. 3—Receiver of pneumatic tube irradiator being monitored by health physics personnel.



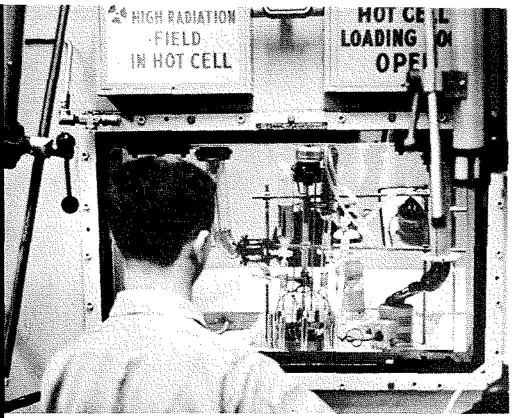


Fig. 4—Chemical experiment in progress in hot cell.

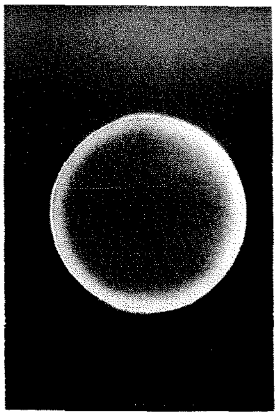


Fig. 5—Autoradiograph showing cesium distribution on target glass assembly of image orthicon tube. (Courtesy of D. J. Shahan)

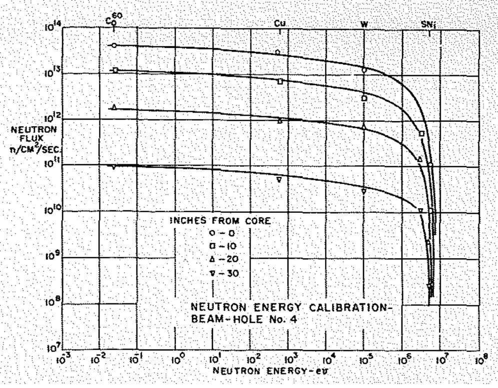


Fig. 6—Neutron energy spectrum of IRL reactor at various distances from the core face.

programs at IRL through the Research Services Laboratory. In this way projects arising throughout RCA receive consultation and limited, specialized assistance. Most RCA research projects at IRL do not orignate within the group responsible for our operations there. Rather, we have established a means for catalyzing projects originating in other RCA research and development groups but requiring nuclear radiation as a means to an end. Where the manhours involved is significant, it is necessary for the project personnel themselves to conduct the experiment at IRL. While the space available to RCA restricts the size of our resident staff, it is feasible at any one time for a few additional RCA engineers from other locations to work there using either available equipment or special apparatus that they bring in for the purpose. Thus a succession of projects, each with its own personnel, normally flows through IRL under RCA auspices. Our resident staff provides guidance, instruction, scheduling, IRL liaison, and supervision of safe practices. Simple irradiations and dosimetry can also be provided.

Several RCA locations, including Princeton, Hightstown, Lancaster, and Somerville, have or have pending facility licenses which enable them to have radioactive materials on the premises. This can greatly facilitate an experiment which requires special equipment that is not easily moved and installed. Material may be activated at IRL, used at another location, and perhaps even moved back to IRL for counting or other processing. A facility license is only granted on the evidence of qualified personnel being employed to supervise the use and handling of radioactive material. It is, of course, important that experiments be done under proper supervision. AEC regulations and IRL procedures require evidence of a proper conveyance and a qualified recipient before radioactive material is released from the IRL premises.

RCA WORK AT IRL

Radiotracer Studies

In the first five years of IRL's operation, RCA has conducted a wide variety of experiments at IRL, even though only a very small percentage of RCA's engineers have been involved. A sampling of these projects will indicate what can be done and may stimulate others to consider the application of nuclear radiation to their work. W. Kern[1] recently described radioisotope work done on Somerville semiconductor projects. IRL's reactor provided some of the isotopes needed and in some cases the work was actually performed at IRL. For example, the contamination of semiconductor surfaces by standard etch solutions was investigated by tagging them with radioactive isotopes of the main constituents. For labelling hydrochloric acid, Cl^{38} was made by neutron bombardment and used at the reactor site because of its short half-life of 36 minutes. Another study determined the contamination of gallium arsenide crystals grown in silica boats. The boats and their impurities were first activated by neutron bombardment at IRL and then transported to Somerville where suitable crystal growing furnaces were available. After the gallium arsenide crystals were grown in these boats, they were sectioned, returned to IRL, and examined in detail for radioactivity. Silicon was traced by the beta emission from the Si^{31} which has a half-life of 2.6 hours and was formed by activation of the Si^{30} in the original quartz. It was found in this work that the entire grown crystal was enclosed in a silicon-rich surface layer, with lesser concentrations in the bulk of the crystal.

Radiotracers have also been used for studying migration of material in image orthicon tubes. D. J. Shahan, Electronic Components and Devices, Lancaster, used IRL's reactor to activate Cs^{133}, producing Cs^{134}, a gamma emitter with a half-life of 2.3 years. The cesium was activated in the same metal channels in which it was introduced into the tubes for processing the photocathode. The tubes were processed in a normal manner, then sectioned, and counted at IRL to determine the amount of cesium getting to the dynodes as well as the photocathode. Counting techniques were supplemented by autoradiography to find the distribution of cesium on a given part when placed in contact with a piece of photographic film. Variations in density of the image on the film indicated considerable nonuniformity of the deposits as shown in Fig. 5.

The diffusion of impurities into semiconductors is a key step in the fabrication of solid state devices and integrated circuits. In this way the electrical conductivity is adjusted and *p-n* junctions are formed for control and isolation functions. The physical process of diffusion is described by a diffusion constant D which is a function of the host crystal, the diffusing impurity, and the temperature. Accurate knowledge of the diffusion constant is essential to the design and quality control of diffused devices. For diffusion from a fixed source, such as an electroplated layer:

$$C(x) = \frac{Q}{(\pi D t)^{1/2}} \exp\left(\frac{-x^2}{4Dt}\right)$$

Where: $C(x)$ is the concentration of diffusant, Q is the amount of source material, x is the penetration, t is the time and D is the diffusion constant. An error-function relationship holds for diffusion from a constant source, such as a vapor. The constant D is measured by determining penetration profiles under fixed conditions of temperature and time.

B. Goldstein[2] has described techniques for obtaining penetration profiles by lapping the sample to remove successive micron-thin layers and measuring the diffusant present. Because of the minute amounts of material to be determined, the use of radiotracers is a preferred way of obtaining penetration profiles. The specific activity of each layer is obtained and plotted as a function of depth. The mechanism of diffusion in compound semiconductors has been clarified by studying self-diffusion, e.g., of Ga in GaAs.[3] Here, radioisotopes offer the only practical means of investigation since the diffusant is not chemically distinguishable from the crystal constituents. While radioisotopes are available from a number of suppliers, the half-life of Ga^{72} is only 14 hours. This radioisotope was therefore prepared by activation at IRL, and then transferred to RCA Laboratories for the experiment with minimum delay.

Radiation Damage

The effects of high energy particles on devices is of particular interest in the communications field where equipment must operate in space or other high energy particle environments. The bombardment can cause damage which may make the equipment inoperative in a short period of time. A number of experiments have been carried out at IRL using both neutrons and high energy gamma rays to study these effects. These experiments were carried out in sample containers in or near the core or in the horizontal beam tubes. Maximum neutron fluxes available in the core are about 2×10^{14} neutrons/cm^2/sec; outside the core, fluxes of 2×10^{13} neutrons/cm^2/sec are available (Fig. 6).

D. A. Gandolfo of the Applied Research group in Defense Electronic Products has carried out neutron irradiations of tunnel diodes. Heavily doped units showed a radiation tolerance to about 10^{17} neutrons/cm^2. The excess (valley) current increased linearly with neutron dose above 10^{16} neutrons/cm^2. Solution grown devices were still operable at 10^{17} neutrons/cm^2 but dot alloy devices failed at lower doses. Some GaAs tunnel diodes were also irradiated under the same conditions and showed a radiation resistance similar to the solution grown germanium units.

Moderately doped minority carrier devices, such as transistors, are extremely sensitive to particle radiation. Very small doses of high energy radiation can produce defects which act as traps and scattering centers and reduce lifetime. Germanium transistors with thin bases and diffused structures were operable in suitable circuits at doses up to 10^{16} particles/cm^2, but more generally the tolerance is only 10^{15}/cm^2. For wide base devices, the lifetime decrease was so serious that they failed at 10^{14}/cm^2. Recently, studies of field effect transistors made at RCA Laboratories have verified that a given amount of damage causes less performance loss in majority carrier devices than in bipolar transistors of comparable geometry (Fig. 7). Probably the most sensitive minority carrier device to high energy radiation is the photovoltaic diode or solar cell. The neutron dose required to reduce output by 25% may be as low as 10^{10} neutrons/cm^2.

During the studies by Applied Research, optical and image recording devices were irradiated with both neutrons and gamma rays. Fiber optic bundles showed typical discoloration due to the formation of color centers. Transmission measurements showed that the opacity was more severe in the violet then the red. Partial recovery of the optical transmission resulted from heating the samples. This annealing presumably released some of the electrons from optically active traps. Gamma rays caused more discoloration than neutrons, indicating the phenomenon was electronic rather than nuclear.

Electrofax paper samples were also irradiated. The sensitivity to gamma rays was so great that irradiations near the reactor core were impractical. In addition the samples were highly radioactive due to neutron activation of the Zn in the ZnO semiconductor. Approximately an hour was needed to allow the radioactivity to decay so that the samples could be handled safely. The images disappeared long before this time. Gamma radiation discharged the samples with doses as low as 170 roentgens. Uncharged samples or images that had been fixed showed no effects with gamma ray doses up to 50,000 roentgens.

Vacuum tubes may show degradation of performance due to particle bombardment in outer space. The most typical cause of degradation is gas evolution. To determine the extent of the problem, special tubes were irradiated at IRL by H. Stern of Electronic Components and Devices. Gas evolution did occur as measured by the rise in pressure in the tubes. Further tests conducted at a pulsed nuclear reactor facility failed to show any pressure rise, however, and it was concluded that the gas evolution under steady state conditions was due to gamma ray heating of the tube elements and tube envelope.

A number of small electron tube types have been exposed to steady neutron fluxes in IRL's reactor to determine their resistance to changes in operating characteristics. As a result of these tests Electronic Components and Devices have been able to provide ratings on the RCA-7586 Nuvistor type in conformity with Military Standard specifications.[4] For example, in a fast neutron flux (mean energy 1 Mev) these tubes withstood an integrated dose of over 10^{17} neutrons/cm^2 with less than 1% change in initial transconductance.

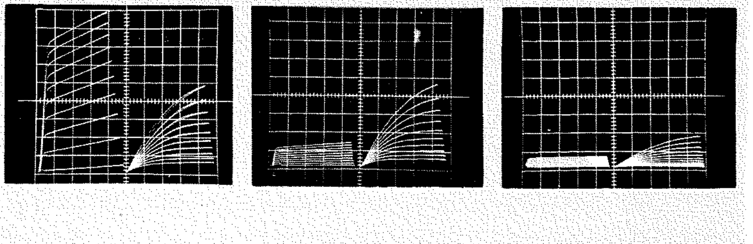
Thermoelectric generators utilizing nuclear reactors as a source of thermal energy have important potential applications in space vehicles. Resistance of thermoelectric alloys to neutron damage is therefore a property of current interest. R. L. Novak has conducted experiments at IRL on the radiation damage to germanium-silicon alloys which undergo partially compensating changes in thermal conductivity, electrical resistivity, and thermoelectric power under neutron bombardment. By shielding the samples with cadmium, a strong absorber of thermal neutrons, he was able to separate the damage effects due to thermal and to fast neutrons. The observed effects on the measured parameters are not only affected by the neutron energy, but also depend on the doping of the alloy. Some of the damage can be annealed out and to investigate this phenomenon it was desirable to minimize the self-heating of the samples in the beam tubes due to gamma-ray capture. This was accomplished by using pool irradiator tubes that are lowered into the pool from above the water surface. This enabled samples to be irradiated while they were simultaneously cooled by the pool water circulating freely around them.

Analytical Applications

Both x-ray and neutron diffraction are used in the determination of crystal structure, and both methods have been employed by J. G. White of RCA Laboratories. Where either might be used, x-rays are to be preferred because of the cost and slowness of neutron beam facilitites. Collecting the data for a single-crystal structure determination by neutron diffraction can take irradiation time worth upward of $10,000.

Neutron diffraction has two principal advantages as compared with x-ray diffraction. The scattering cross section for x-rays is proportional to the atomic number of the atom, while for neutrons it varies randomly throughout the Periodic Table and, hence, the positions of light atoms can be found in presence of heavy atoms. The neutron has a magnetic moment which interacts with ordered magnetic

Fig. 7—Effect of neutron bombardment on current-voltage characteristics of transistor for bipolar (left) and unipolar (right) operation. Fluxes are: left, 0; middle 10^{10} neutrons/cm^2; right, 10^{15} neutrons/cm^2. (Data by S. M. Christian).



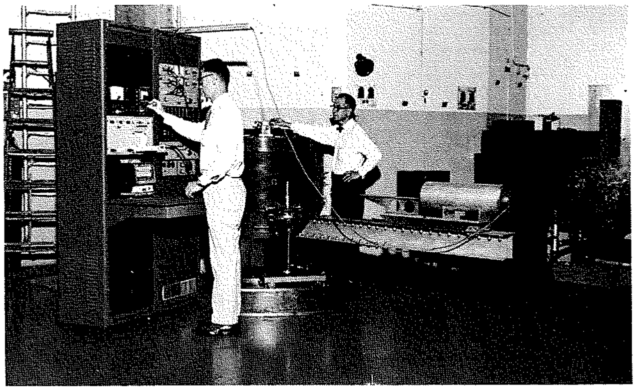


Fig. 8—D. A. Ross and J. G. White adjusting the neutron spectrometer built by RCA and installed at one of the beam ports penetrating the concrete pool wall.

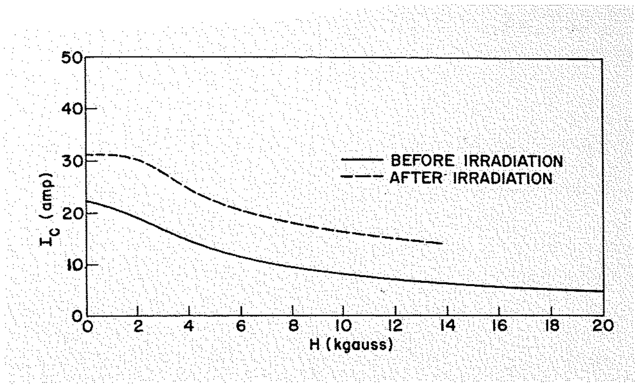


Fig. 9—Critical current versus perpendicular magnetic field for Nb_3Sn before and after irradiation at 6.6 x 10^{17} neutrons/cm^2. (From Cullen and Novak)

spins in a ferromagnet to give an additional scattering pattern from which, in principle, the arrangement of the magnetic spins can be deduced. Effective work on crystal structure determinations should be carried out in close conjunction with the scientist synthesizing the materials. This has proved to be feasible, furnishing an excellent example of mutual reinforcement and recognition between these two activities.[5] Neutron spectrometers are usually made to order. The one at IRL, shown in Fig. 8, was built by RCA in its own shop from designs of Professor C. G. Shull at M.I.T. It uses a lead monochromating crystal allowing easy variation of the neutron wavelength. Both polycrystalline and single crystal specimens may be examined and a cryostat is available for low-temperature measurements.

Neutron activation as a method of analysis for trace impurities has the advantages of high sensitivity, elimination of the blank correction, and small matrix effect. Sensitivity of 10^{-12} gm/gm are feasible for some elements. It is especially useful where a specific impurity is being sought. In RCA, emission spectrography or mass spectrography has been used for broad range analysis with neutron activation reserved for special cases. One such case has been the determination of oxygen in GaAs since this element is hard to determine at the parts-per-million level by other means. Because of the short half-life of the oxygen activation product, a special method was developed suitable for use with the reactor.[6] The reaction O^{16} (T, n) F^{18} was used, the F^{18} having a half-life of 1.87 hours. The tritons were produced by wrapping GaAs wafers in lithium and using the reaction Li^6 (n, T) He^4. By etching the samples after activation, it was possible to separate the effects of bulk oxygen from that of surface oxygen, which had a considerably greater concentration. Sensitivities of 5 ppm-atomic were achieved.

Constructive Applications

While many applications of the neutron source available at IRL degrade the material irradiated or at best are analytical in nature, there is also the potentiality for using neutrons constructively in the improvement of materials or in the fabrication of new devices. Neutron-induced defects in superconductors are currently being studied using the high neutron fluxes available at IRL. McEvoy, Decell, Cullen, and Novak[7,8] have shown that the critical current density in Nb_3Sn can be increased by irradiation with fast neutrons at fluxes of 10^{17} neutrons/cm^2 and greater (Fig. 9). Extended defects produced by other means, such as cold working or grain boundary precipitations, are also known to affect hard superconductivity. Further neutron studies should shed more light on the interrelation of these effects and the usefulness of neutrons in controlling the properties.

Z. Kiss and others at RCA Laboratories have irradiated $CaF:Dy^{2+}$—Dy^{3+} laser crystals at both liquid nitrogen and room temperatures with gamma rays from spent fuel elements. Crystals doped with other rare earths have also been reduced by gamma irradiation.[9] Free electrons ejected from the host atoms were captured by the Dy^{3+} ions reducing them to a divalent state. Since the laser action is due to one of the transitions of the divalent ion, the increase in Dy^{2+} concentration produces an increase in light output for the same pumping power. A gamma ray dose of about 10^6 roentgens at a dysprosium concentration of about 0.02% molar was found to be optimum. Although the simple photoreduction is not a permanent effect, it has provided crystals for research studies while other methods were being investigated. An interesting method for studying the effects of crystalline fields on the rare earth impurities developed from this work. At low temperatures many of the excited electrons remain in their excited states. As the crystal warms up to room temperature, the excited electrons drop to lower energy states emitting visible radiation. By investigating the optical spectra of this thermoluminescence it is possible to obtain information about the location of the atom in the crystal and the effects of the host atoms surrounding it. This information is of value in the continuing search for better laser materials.

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MICROWAVE RESEARCH

Devices for the Future

This discussion of the scope and trends in microwave research explores the future of microwave by examining the present state of the art and problems that must be solved, and by examining the historical trends of earlier research. Information on current RCA Laboratories projects is included and compared with that of two years ago.

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Dr. **LEON S. NERGAARD** attended the University of Minnesota and received the BSEE in 1927. He received the MSEE from Union College, Schenectady, New York, in 1930 and the PhD in physics from the University of Minnesota in 1935. From 1927 to 1930, Dr. Nergaard was associated with the research laboratory and vacuum-tube engineering department of the General Electric Company. He held a teaching assistantship in the Department of Physics at the University of Minnesota from 1930 to 1933. Dr. Nergaard joined the RCA Manufacturing Company at Harrison, New Jersey, in 1933 and transferred to RCA Laboratories as a research physicist in 1942. At Harrison, he worked on microwave measurements, on receiving, transmitting, microwave and pulse-radar tubes. At Princeton he continued on pulse radar tubes, and then worked on transmitting tubes and television transmitters. Next, he worked on solid-state physics, particularly the semiconducting properties of oxide cathodes. He assumed responsibility for the microwave work at RCA Laboratories in 1957. He has 24 issued patents and many published papers, has received two "RCA Achievement Awards" and the "David Sarnoff Award for Outstanding Achievement in Science." Dr. Nergaard is a Fellow of both the APS and the IEEE, and a member of the American Association for the Advancement of Science. He has been active in numerous committees of the IEEE, URSI, and is a member of Theta Kappa Nu, Gamma Alpha, and Sigma Xi.

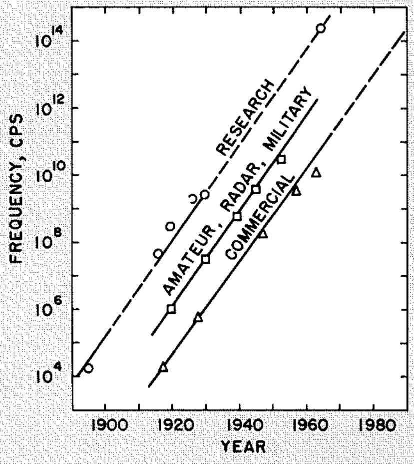

Fig. 1—Utilization of the communication spectrum.

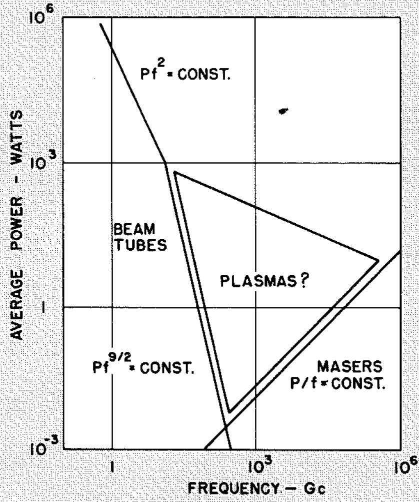

Fig. 2— Power generation.

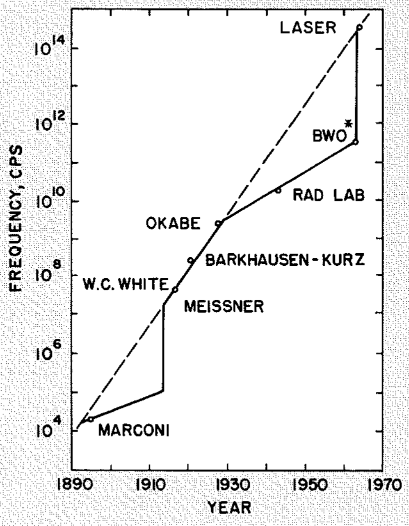

Fig. 3—Frequency vs time research (backward-wave oscillator).

THE title of this paper expresses in a nutshell the dilemma and challenge of microwave research. The dilemma is, first of all: *what is the future of microwaves; what needs can they serve, what can they add to our way of life and in what areas?* Then, *what do we mean by the future—tomorrow, the day after tomorrow—what is the time scale on which we base our consideration of the future?*

These are difficult questions and no simple answers are apparent, at least to the writer. However, the past provides a base from which to speculate on the future and if past progress is meaningful in extrapolating with the future, there is indeed a direction to follow and a need to be met. The course one pursues to meet the need may be upset by the events of tomorrow, but this is the nature of research; in fact, one hopes for the unexpected, the things that will open up new avenues to immediate or long-range goals. The direction pursued in the Microwave Research Laboratory and the reasons therefore are discussed in the next section.

The second part of the dilemma is: *given a sense of direction, how does one choose what to do?* Here *an* answer is easier. One can choose a meaningful future system and see what it takes to implement it. If the system is sufficiently remote, a consideration of the devices required to construct it will expose a host of devices not yet available, some of which can be achieved by known physical principles, some that defy our present knowledge. Thus research must proceed on several fronts. On one front it must explore existing knowledge with a perceptive eye to see what might provide a needed device and then demonstrate that it will indeed do so. On another front it must seek new knowledge in areas that may provide new phenom-

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ena that make possible devices now unattainable. The choice of area is largely a matter of experience, intuition and, hopefully, serendipity. How our present areas of exploratory research are rationalized is discussed in a subsequent section.

The trend of our research and the nature of the devices now under study and sought have made *systems thinking* mandatory. These devices, which will be described still later in the discussion, do not seem to lend themselves to conventional microwave circuitry. Thus, we are led to contemplate completely-integrated systems and subsystems and we have, in fact, undertaken an exploratory experimental program on such systems. (Our initial steps in this direction are described later in the paper.)

WHERE LIES THE FUTURE?

When one considers those properties that distinguish microwave, or, more generally, high-frequency power, one stands out—the ease with which it may be coupled into "free space." A simple half-wave dipole transforms the 377 ohms "per square" of free space into 73 ohms at the antenna terminals. As the frequency increases, the size of a dipole or antenna array shrinks correspondingly, so does its cost for a given aperture. If, in a facetious vein, one extrapolates to a frequency of 10^{13} cps and doesn't care too much about spectral purity, he finds he can get a radiator, complete with power source and an output of about 5 watts/cm^2 of aperture for \$0.04/cm^2. Because high-frequency power is so easily coupled into free space in which the power propagates subject to the laws of physical optics, it provides a wonderful method of communication and, to date, this has been by far the major application of high frequencies.

The past 50 years have wrought no change in this situation. It therefore seems reasonable to proceed on the assumption that this situation will prevail for a few more decades. If so, in what direction should research head? Fig. 1 is pertinent to this question. The line on the right of the figure shows the commercial utilization of the spectrum from the end of World War I to the present. It will be noted that the spectrum has been "consumed" exponentially with a time rate of about 10:1 every 7 years. If the trend continues for a few more years, we will reach the end of the solid line in the figure, the point (approximately) at which atmospheric absorption becomes severe. Is this the end of the road? The writer thinks it is not; as satellite communication opens up, he foresees a need for the compact antenna systems that these frequencies afford for communication between stations above the atmosphere at the least.

The research that engendered the present systems is shown by the line at the left. Note that there is about a 25-year lag between the discovery or invention of the devices that made the systems possible and their use in highly-reliable systems. Note also that the research line has dotted sections. These sections will be examined in more detail in a subsequent section. The uppermost point on the curve shows the achievement of the optical laser, a result promised by the microwave maser. If the extrapolation of the utilization curve holds, the laser will find commercial application in about 1990. There is a lot of engineering and proving between research and highly-reliable service.

The center line shows the utilization of the spectrum by amateurs, radar systems, and the military. Its position shows a different tradeoff between application, urgency, cost and reliability than does the commercial curve.

It is the writer's feeling that past research directed to ever higher frequencies for communications should continue. This is his present sense of direction. Hopefully there will be spin-offs, perhaps another transistor or tunnel diode, that provide better, cheaper and more reliable devices for existing systems. As to the time scale, Fig. 1 shows the time scale of the past. There is persuasive evidence that the time between innovation and utilization is shrinking. Perhaps our needs are out-stripping our ability to innovate or, more likely, research emphasis in the microwave field has been misplaced. More of this later.

WHERE ARE WE AND HOW DID WE GET THERE?

Fig. 2 shows our ability to generate power at various frequencies. On the left lies a line with a knee labeled *Beam Tubes.* It shows that we can generate about 100 kw at 1 Gc, that this power drops rapidly with frequency up to 10 Gc, and then drops drastically to the milliwatt level at about 300 Gc. At this point, the curve showing the power available from masers and lasers crosses the tube curve and rises as the size of the quantum increases. The limitations on both curves seem to be fundamental. There is a wedge, centered at about 300 Gc where we are unable to generate any substantial power. This wedge is labelled *Plasmas?* The question mark might have sufficed. However, *Plasmas* denotes not merely the familiar gas plasmas but a philosophy. The fall-off in power of beam tubes at high frequencies arises from scaling laws, the need to keep two dimensions of a tube less than a wavelength. Hence, as the frequency increases, tube structures tend to become microscopic, incapable of handling much power, and exceedingly difficult to construct. A way out is to discover materials which can oscillate and amplify as a result of bulk properties. Plasmas in gases and in solids do display electrical resonances as the result of bulk properties. Hence the labelling of the wedge. Another year may bring another label.

The route by which the present state of the art was reached is shown in Fig. 3. It is the research curve of Fig. 1 in more detail. From about 1900 to 1914, the principal sources of high-frequency power were spark-gaps, Poulsen arcs, and Alexanderson alternators. The frequency was pushed up to about 100 kc by 1914. Then Meissner invented the vacuum-tube oscillator and opened up a decade of remarkable discovery and innovation. A few high points are noted: in 1916, W. C. White operated a triode at a frequency of 50 Mc, in 1920 Barkhausen and Kurz discovered the oscillator which now bears their names, and 1927 Okabe invented the split-anode magnetron. The next 35 years have shown about as much progress in extending the spectrum as did the years between 1914 and 1927. Admittedly, there has been a period of consolidation and refinement and, lately, a certain amount of innovation. Yet an extrapolation of the 1927-1964 line shows utilization ahead of research by about 1970. The writer cannot escape the feeling that microwave research should have been bolder and more adventuresome in 1949 when the practical fruits of solid-state research began to emerge and should have committed some of its manpower to solid-state physics instead of pursuing what in retrospect looks more and more like engineering. This is water over the dam but perhaps a lesson lurks there.

WHAT ARE WE DOING NOW?

A cryptic but simple and descriptive answer to the question posed is: *we are changing.* There was a time, say 1957, when the program of the Microwave Research Laboratory was entirely tube work. The tube work included microwave tubes and gas-plasma tubes which showed promise of providing microwave power sources. A spin-off of the latter was the thermionic energy converter, which has progressed rapidly and independently to its present advanced state. With advent of the varactor, which made microwave parametric amplifiers feasible, we embarked on a program of solid-state microwave work. This work has included parametric amplifiers, multipliers and mixers, tunnel-diode amplifiers, and Hall-effect amplifiers. The discovery of plasma behavior in semiconductors by Steele and Glicksman (of RCA Laboratories) prompted an exploratory study of plasmas in solids still in progress, both at Princeton and at our Tokyo laboratory. Thus, solid-state devices began to intrude in a field previously dominated by vacuum devices. The trend continues.

The research program as of 1962 is outlined in Table I. The research program of 1964 is similarly outlined in Table II. There are some notable changes:

1) The work on parametric, tunnel-diode and Hall-effect devices has moved up in frequency. It should be noted that the Hall-effect devices represent an attempt to use bulk properties of a semi-conductor to achieve gain. Furthermore, since Hall-effect is a nonreciprocal phenomenon, it represents an attempt to achieve a true two-port device with separate and distinct input and output ports in contrast to the usual parametric and tunnel-diode amplifiers.
2) In another attempt to use bulk properties to achieve gain, a study of the use of superconductivity to achieve gain was undertaken. The result to date is an amplifier with 11-db gain at 6 Gc.
3) The discovery of plasma oscillations at frequencies up to 40 Gc in indium-antimonide has moved it from the purely exploratory to the device category. While much exploration remains to be done, it is much less speculative than it was in 1962. Furthermore, related work at our Tokyo laboratory has resulted in the achievement of an indium-antimonide microwave isolator. The solid-state plasma work seems well on the way to providing useful devices.
4) During the past year the noise figure of an s-band traveling-wave tube was reduced to 1 db, as reported elsewhere.[1] Theoretical work must yet be done to reach full understanding. There also remains some hardware work to reduce the results to commercial form. The latter is probably better left to a product division.
5) The beam-plasma amplifier provided substantial gain (8 db) at 23 Gc. An evaluation of its performance and potential showed limitations similar to those of conventional beam tubes. Work on this tube is being phased out.
6) Work on the crossed-field tube continues. Crossed-field amplifiers have in the past exhibited high efficiency but rather low gain. Hopefully, this tube will exhibit both high efficiency and high gain. If it meets theoretical expectations, we shall continue; if not and lacking another idea, it will be phased out. A spin-off of this work is a new crossed-field electron gun with a beam current ten times as high as previously achieved.
7) Double-stream tubes achieve gain without interaction circuits and in effect display bulk properties. The present study is intended to evaluate their potential at higher frequencies.
8) The "compact klystron" is an evaluation of some new ideas to increase the efficiency of small klystrons for space telemetry systems. It is a short-term project.
9) The cesium hollow cathode is a new cathode which has delivered the phenomenal current density of 800 amp/cm. It is planned to test this cathode in a small klystron to assay the practical problems that may arise in its application.
10) The supporting and exploratory work is just what its name implies. It provides such fundamental data as are needed in the device work and looks for new phenomena that may serve as the basis for new devices.
11) The integrated systems are discussed in the next section.

This discussion of work in progress and trends has been brief and superficial. It is hoped that it conveys more of an idea of the content and intent of the program than would a mere listing of projects.

INTEGRATED SYSTEMS

The trend towards solid-state microwave devices poses problems in the circuitry as well as in the devices themselves. Solid-state devices, particularly those using bulk effects, tend to exhibit very low terminal impedances. Strip lines go part of the way in meeting this difficulty. To go the whole way, very small spacings are called for, spacings that may best be achieved by evaporation techniques. If so, why not make the devices and circuits by the same technique and at the same time? If this could be done, whole subsystems could ultimately be made in very compact and cheap forms. These considerations led us to undertake an exploratory program in this direction. The program is just getting under way and the present work is more in the nature of acquiring background than of pursuing the program itself. We are availing ourselves of the vast experience of Somerville in semiconductor technology and are grateful to them for their forebearance and cooperation. The present work comprises:

1) A study of the overlay transistor to understand and exploit its capabilities. This work is performed at the request of and in collaboration with Somerville and affords us an ideal indoctrination in microwave transistor problems.
2) A study of evaporated strip lines to learn the technology of making them and to determine their properties—their losses, cross-coupling, and other pertinent characteristics.
3) A study of transitions from evaporated lines to standard coaxial lines so that we can use our present instrumentation. Besides, the resistance of a half-wave dipole, evaporated or not, is 73 ohms.
4) A study of ferrites for strip-line use. Ultimately, we must integrate switches, isolators and circulators into the composite system.
5) An investigation of systems, one of which will be chosen as a test vehicle. We plan to choose a system, simple yet complex enough to pose the problems we must solve if we are to achieve an integrated system. The implementation of such a system will provide the unified effort required to achieve the objective.

If this program advances as anticipated, it will soon become a major project and will require a considerable redeployment of the available manpower.

CONCLUSION

What are the devices of the future? The writer does not know, but for communications he is certain that they will be solid-state devices. Apparently others think so, too. The CW-60 all-solid-state relay system is a remarkable achievement.[2] Perhaps the next generation of this system will be integrated as well.

TABLE I
Microwave Research Program, 1962

	Frequency, Gc
Solid-State-Device Research	
Parametric Amplifiers	3
Tunnel-Diode Amplifiers	3
Hall-Effect Amplifiers	3
Tube Research	
Low-Noise Traveling-Wave Tubes	3
Crossed-Field Tubes	3
Double-Stream Tubes	3
Beam-Plasma Amplifier	23
Cesium Hollow Cathode	
Propagation of Space-Charge Waves	3
Klystron Window Failure	3
Supporting and Exploratory Research	
Recombination in Cesium Plasmas	
Gas-Plasma Diagnostics	
Plasmas in InSb	

TABLE II
Microwave Research Program, 1964

	Frequency, Gc
Solid-State Device Research	
Parametric Amplifiers	55
Tunnel-Diode Amplifiers	55
Hall-Effect Amplifiers	Infrared
Superconducting Parametric Amplifiers	55
Solid-State Plasma Oscillators	≤ 40
Oscillation in Intermetallic Compounds	5.0
Tube Research	
Low-Noise Traveling-Wave Tubes	3
Crossed-Field Tubes	3
Double-Stream Tubes	3
Compact Klystron	3
Cesium Hollow Cathode	—
Supporting and Exploratory Research	
Gas-Plasma Diagnostics	
Phonon Propagation in CdS	
Laser Modulation and Deflection	
Wave Propagation in Solids	
Integrated Systems	
Overlay Transistor	2
Evaporated Strip-Lines	3
Transitions	3
Ferrites	—
A System	—

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RECENT RESEARCH ON LOW-NOISE TRAVELING-WAVE TUBES

The low-noise traveling-wave tube is a microwave receiving tube possessing broad bandwidth, high gain and good stability — all without sacrifice of sensitivity. These tubes thus find application in military radar and missile guidance systems as well as in telemetry and commercial radio-relay systems. This paper emphasizes noise-reduction research conducted at RCA Laboratories in recent years which has resulted in the lowest TWT noise figures yet obtained.

Dr. S. BLOOM

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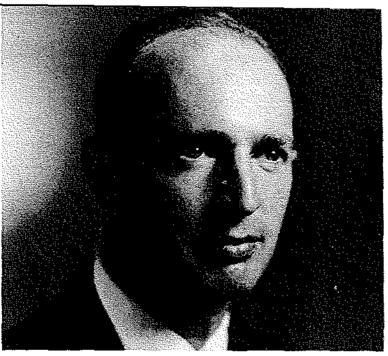
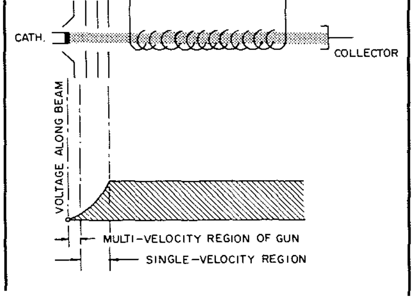


DR. STANLEY BLOOM received his BS in Physics from Rutgers University in 1948, and PhD. in Physics in 1952 from Yale University, where he worked on gaseous electronics and on the quantum theory of spectral line broadening. He joined RCA Laboratories in 1952, where he has worked on traveling-wave tubes, beam electronics, molecular amplification, plasmas and parametric amplifiers. He is at present Head of the Electron Physics and Devices Group in the Microwave Research Laboratory of RCA Laboratories. Dr. Bloom has received two RCA Achievement Awards for his work on the noise properties of traveling-wave tubes. He is a Senior Member of the IEEE and a member of the APS, Sigma Xi, and Phi Beta Kappa.

We start with a reminder of how the traveling-wave tube works. Fig. 1 shows a narrow electron beam originating from a hot cathode and accelerated by three or four gun electrodes up to a final potential of several hundred volts. The beam then goes through a helix and is finally collected. The beam is focused by a strong magnetic field. The signal to be amplified is coupled into one end of the helix, and microwave power is taken out at the other end. The electromagnetic wave on the helix travels with the speed of light along the wire; however, because of the helical pitch the axial component of the field has a velocity which is only about 1/20th as fast. By accelerating the beam to very slightly above this axial wave velocity, the beam and wave can be synchronized and cumulative interaction can occur. Because electrons tend to bunch where the wave is in a retarding phase, more electrons give up their kinetic energy to the wave than extract energy; hence, the signal wave is amplified. So too, however, is the noise on the beam.

The main sources of noise in the TWT arise at the thermionic cathode. Because the electrons leave at random times the beam current exhibits shot noise. In addition, the electrons are emitted with a wide thermal-distribution of velocities; this causes velocity-fluctuation noise. The product of the magnitudes of these two noise sources is called the *beam noisiness* and labeled S. The mutual dependence of one noise source upon the other is called the *noise correlation* and is labeled π. The tube noise factor F, when optimized, is—reasonably enough—proportional to the difference S-π. Because the current and velocity fluctuations are independent at the cathode, π is zero there. For a cathode at emperature T_c emitting a Maxwellian beam with full shot noise in each velocity class, the beam-noisiness power has the value $S = kT_c$ at the cathode.

Fig. 1—TWT and potential profile along beam. Low-voltage, multivelocity-flow region of the gun is bounded by the cathode, beam-forming electrode, and first-anode electrode. At higher voltages further downstream, flow is single-velocity. Lowest noise figures require each region to be optimized; the former to reduce noise on the beam, the latter to minimize coupling of this noise to the helix.



SINGLE-VELOCITY AND MULTIVELOCITY REGIONS SEPARATELY OPTIMIZED

Suppose, now, that the beam is accelerated abruptly (which it isn't) to a volt or more just in front of the cathode. The mean velocity of the beam is then sufficiently greater than the thermal velocity that multivelocity effects can be ignored. Under this condition the noisiness S and the correlation π are constant along the beam; π remains zero and S remains kT_c. With these noise powers thus fixed, the best one can do is to minimize their coupling to the helix. This optimization is done by so adjusting the gun-electrode voltages and spacings as to affect the best compromise between noise launched on the helix due to velocity fluctuations and noise due to current fluctuations. The result is a minimum TWT noise figure of $F = 1 + [(S\text{-}\pi)/kT_{290}] = 1 + [T_c/290]$, or about 6 db for a cathode at 1,000°K.

Beams, however, are *not* abruptly accelerated and measured noise figures are better than this 6-db "minimum." The gradual acceleration in front of the cathode provides a more-or-less extended region of very low voltage. In this region, the velocity distribution effects can not be ignored. Computer analyses made at Stanford University,[1] which were later[2] put into closed form and linked to the general phenomenon of Landau damping, show that because of the space-charge interaction between beamlets in adjacent velocity classes neither the overall noisiness S nor overall correlation π are any longer constant. These quantities vary with distance in a multivelocity beam and so the object of proper gun design is to decrease S and increase π. This is done by varying the voltages on especially the electrodes closest to the cathode. Having produced a beam with a small S and large π by optimizing the multivelocity region, one must still optimize the single-velocity region, as previously described, so as to realize a noise figure proportional to S-π. In this way the noise figure can be made, for all practical purposes, arbitrarily small.

The increase in π under gun-voltage conditions typical of present-day ultra-low noise TWT's has been observed.[3] In this mode of operation, as the voltage V_{BF} on the beam-forming electrode (see Fig. 1) is made sufficiently *positive*[4] the noise figure drops from about 6 db to about 3 or 3.5 db. Simultaneously, the correlation π is observed to rise from zero at negative V_{BF} to around 0.4 at optimally-positive V_{BF}. These values of π were inferred from noise figure measurements on a special TWT having an adjustable gun-to-helix distance. This dramatic increase in π indicates that the drop in noise figure is due to multivelocity effects occurring in the cathode region, in accordance with present-day theory. The positive beam-forming voltage draws current from the cathode and so allows a lower voltage V_1 to be used on the first anode. This results in a longer multivelocity region of low voltage than if negative V_{BF} and therefore high V_1 were used. The strong magnetic field prevents electrons from being collected on the highly-positive beam-forming electrode. Other benefits—to be pointed out later—accrue from the use of a high magnetic field.

MORE-VERSATILE TWO-ELECTRODE GUN

Improved flexibility is possible through use of a second beam-forming electrode.[5]

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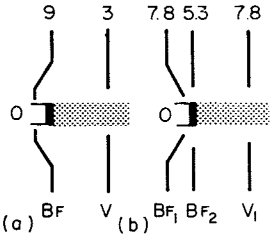


Fig. 2—a) Conventional gun with one beam-forming electrode. (Electrodes following first anode not shown.) b) Gun with two beam-forming electrodes provides independent control of fields around cathode over fields nearer first anode.

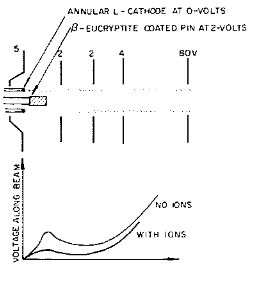


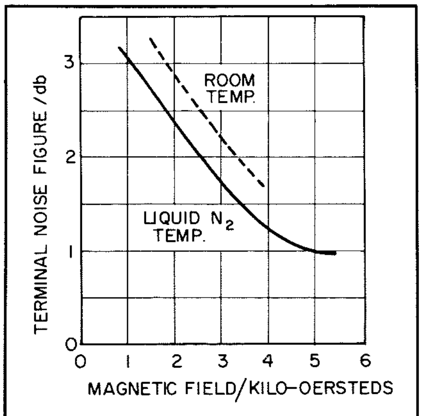
Fig. 3—Tube with plasma-cathode gun uses lithium ions to reduce space potential without need for high radial fields of conventional guns.

Fig. 2a shows a conventional gun; a single beam-forming electrode BF and a first anode V_1 are shown, at typical voltages of 9 volts and 3 volts. Fig. 2b shows two beam-forming electrodes, BF_1 and BF_2, at 7.8 and 5.3 volts, with a first anode at 7.8 volts. The gun in Fig. 2b allows the low-voltage region adjacent to the cathode to be adjusted independently of the single-velocity portion of the cathode-first anode space lying further downstream. Electrolytic tank plots show that changes in BF_1 voltage cause only small changes in the field pattern beyond electrode BF_2. Furthermore, noise figure measurements in which voltages BF_1 and BF_2 were varied indicated the independent control over the cathode-edge fields afforded by this gun. Finally, the gun in Fig. 2b with its higher V_1 value is less prone to the formation of a noise-enhancing virtual cathode.

PLASMA CATHODE—LOW-NOISE BEAM

Further evidence that the multivelocity, low-voltage region near the cathode plays an important role in noise reduction was obtained from experiments on a plasma-cathode TWT.[6] An otherwise standard TWT gun was equipped with an annular L-cathode and a center-pin electrode as shown in Fig. 3. The center pin was coated with β-eucryptite which thermionically emits lithium positive ions. These mix with the electrons emitted from the L-cathode to form a plasma. This tube, with normally a best noise figure of 5 db, gave about 4 db when the lithium-ion source was turned on. The noise figure steadily decreased to this value as the ion current was raised or as the length of the plasma region in front of the cathode was increased. These results indicate that such a synthesized plasma—aside from not being inherently a "hot" medium—can actually be used as a noise-reducing medium. This reduction is attributed to the fact that the plasma produces the desired low-voltage region in front of the cathode without the need for strong radial fields for current injection. This is in contrast to conventional cathode and gun structures in which strong radial fields (due to positive V_{BF} voltages) lead to high electron temperatures at the beam edge.

Fig. 4—A 1.0-db noise figure for an S-band TWT at 77°K and a magnetic field of 5,000 oersteds. Cooling both prevents solenoid burnout and reduces helix-resistance noise contribution by about 0.5 db.



A REFRIGERATED VACUUM TUBE

The temperature of the beam electrons is not the only temperature affecting the TWT noise figure—that of the wire helix and other circuitry is also important. An analysis showing, first, that the thermal noise generated in the resistance of the helix contributes to the noise figure an amount no longer negligible at the present state of the art and, second, suggesting the use of circuit refrigeration was made.[7] An experiment in which an entire TWT was immersed in liquid nitrogen showed a noise figure decrease of about 0.5 db below the value at room temperature. This study also points up the importance of using low-loss material and low-loss structural design.

1.0 DB

Although a 0.5-db improvement in noise figure due to refrigeration is significant, it is dwarfed by the results of another more-recently uncovered technique.[8] A larger-than-usual magnetic field was used to reduce the TWT noise figure to only 1.0 db. *This is the lowest noise figure yet obtained at any frequency with any beam-type microwave amplifier.*

In these experiments, a commercial RCA 3-Gc tube that ordinarily has a noise figure of about 3.5 db was immersed, together with its focusing solenoid, in liquid nitrogen. This had two effects: First, cooling the helix reduced the thermal noise, as described above, and improved the noise figure by about 0.5 db; the second, and more significant, effect of cooling was that it permitted the current in the solenoid to be raised, without overheating, to provide a focusing field of over 5,000 oersteds. This high field, compared with the field of 800-oersted attainable with the uncooled solenoid, was found with several tubes to give an additional reduction in noise figure of about 2.0 db. (See Fig. 4.)

The fact that curves essentially parallel to, but at higher ordinates than, those of Fig. 4 are obtained for non-optimal voltages on those electrodes which affect the multivelocity region of the gun, indicates that the high magnetic field is reducing a new, additive noise term in the noise figure. Furthermore, the fact that the excess noise figure, F-1, falls with increasing magnetic field roughly as $1/B$, indicates that this added term may be due to transverse, or cyclotron-frequency dependent, fluctuations of the beam, in contrast to the axial plasma-frequency dependent fluctuations causing the old S-π term. Further experiments are needed for clarification.

The attainment of a 1.0-db TWT noise figure represents the culmination of over a decade of research at RCA and many other industrial and university laboratories. Not surprisingly, though, it also raises many new questions calling for continued research, as well as many practical problems requiring added efforts in advanced development. For example, *what is the actual transverse-noise mechanism, and where along the beam is it occurring? Can this noise reduction be obtained by simpler means not requiring large magnetic fields?* On the practical side, there is a need for extending these techniques to X-band and higher frequencies, and a need for high-field permanent-magnet focusing systems that are compatible with low-noise.

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MICROWAVE DEVICES— A SURVEY OF BUSINESS POTENTIAL

In this review of the business potential for microwave, it is pointed out that a continued expansion of applications can be expected. Explosive invention will offer much stimulation to the developer and a challenge to management attempting to perform profitably despite rapid obsolescence. The microwave component designer can now offer the systems engineer technically highly skilled service to support him in the development of complex, sophisticated systems. Break-throughs in performance will continue in all areas (power, frequency, noise, life, size, etc.), especially stimulated by the rapidly developing microwave solid-state art, and dramatic cost reductions will generate a quantity of new applications.

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During 1964, sales in the microwave-component industry will exceed 200 million dollars. On the basis of the strong past and excellent future growth possibilities (Fig. 1) this industry looks like one of the most interesting and promising of those in the electronic product areas. However, because of three main factors, the microwave-component industry is one of the most difficult in which to participate with a steady profit performance.

First, microwave components include a very wide range of diverse products, such as magnetrons, traveling-wave tubes, klystrons, crossed-field amplifiers, triodes, quantum devices, TR and ATR devices, parametric amplifiers, tunnel-diode circuits, varactor multipliers, masers, ferrite and crystal devices, and others. Fig. 2 shows some of the products included in the RCA lines of microwave components. These components differ from each other as much as do receiving tubes from transistors or image converters from transformers. A very wide scope of knowhow and extensive engineering effort is therefore required for adequate coverage or specialization in one or a few selected component areas.

Second, the microwave-component industry is still very young; only during and since World War II has it been of an appreciable size. As a result, new inventions still far outshadow established products, and steady production programs covering several years represent the exception.

Finally, the bulk of the microwave business, both developmental and production, represents military applications; sizeable commercial uses are only now starting to establish themselves. Because military systems are in a continuous struggle for superiority between offensive and defensive capability, they are in a constant state of redevelopment requiring much innovation and only modest production.

Microwave components are highly integrated, that is, they incorporate a considerable portion of the system circuitry and thus are suitable for one or a few applications only. Those attempting to form a profitable microwave-component business find themselves forced into the intricate juggling act of picking the proper product areas and user systems at the right time. This situation leads to individual participation by system rather than by product, with the entire microwave business fluctuating according to over-all military strategy. The result is a strongly fluctuating business performance among individual microwave-component manufacturers with waves of ascending or descending cycles superimposed on the entire industry. One could then sum up the microwave-component business as technically most stimulating and rewarding and financially subject to uncontrollable excursions that are usually larger in the negative direction because government profit regulations act as limiters on positive excursions.

WHAT ARE THE MOST PROMISING NEW MICROWAVE AREAS TODAY?

There are many promising areas for the use of microwave devices in both military and commercial applications. Table I lists a number of military uses for microwave devices. Commercial uses at present are not nearly so varied or extensive; however, there is a definite and progressive growth of the potential in this area. Table II lists several commercial applications for which microwave devices would be required.

The broad span of military uses, particularly in space and phased-array applications, and the expanding commercial possibilities promise an excellent future for the microwave-component industry. The varied and unusually demanding requirements of these applications, however, also forecast that device capabilities will have to be advanced to new levels, without any significant increase in the costs of development and production, if this promise is to be realized.

Space Applications

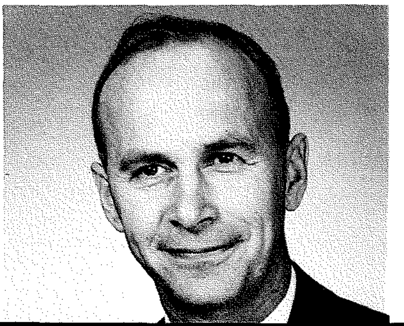
Space applications today impose the most stringent demands on microwave components. The vibrational conditions of launch and the subsequent exposure of the components to vacuum and temperature extremes dictate severe environmental requirements. Moreover, each pound lifted into space is essentially equivalent in cost to its weight in gold; thus, small size, light weight, and high operating efficiency are premium requirements for the components used in space systems. Because no maintenance or repair is possible during the operation of a space system, new concepts of reliability are required to provide adequate assurance that no operational failures will occur and, very often, that extremely long life performance will be obtained.

Conventional-design microwave devices usually prove inadequate for space use, and extensive engineering efforts are required to develop components capable of meeting the stringent require-

TABLE I—Military Uses of Microwave Components

Radar — early warning, search, acquisition, guidance, fire control, navigation, surveillance, AICBM, instrumentation, telemetry, phased array, marine, weather.
Communications — radio relay, walkie-talkie, scatter, satellite.
Countermeasures — jamming, decoy, augmenters, ECC.
Satellites — communications, navigation, research, military.
Beacons
Missiles — launching, guiding, firing.

HANS K. JENNY received the MSEE from the Swiss Federal Institute of Technology in 1943. Later, he was associated with the Institute where, for a period of two years (1944 and 1945), he did research work on Klystrons and their application in radars. Mr. Jenny joined the RCA Microwave Tube Operation Department in 1946 and was given the responsibility for the development of CW magnetrons and frequency-modulating schemes for magnetron oscillators. This work also included studies of the concept of coupled-cavity tuning. He holds four patents on magnetrons. He was promoted to his present position, Manager, Microwave Engineering, in 1950; since then he has had the over-all responsibility for all the engineering activities carried out in the various departments of the RCA Microwave Tube Operations. Mr. Jenny is a Senior Member of the IEEE.



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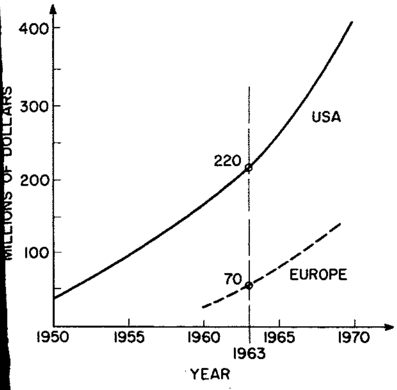

Fig. 1—Microwave component sales.

. 2e-g—Representative RCA traveling-wave
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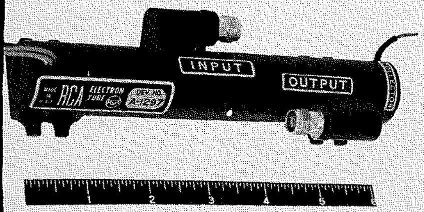
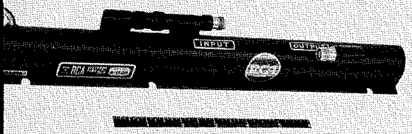
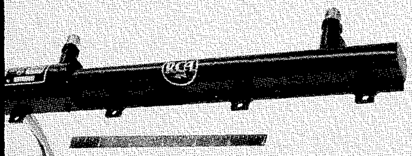


2e) A-1297 miniature tube



2f) A-1309 1-watt S-band ruggedized tube

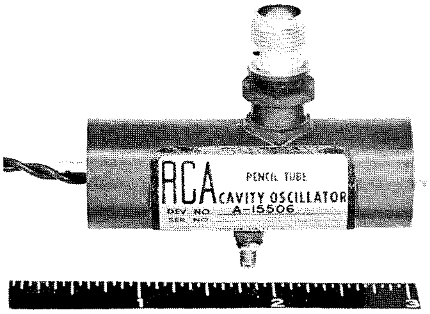


2g) Type 4056 S-band tube for satellites

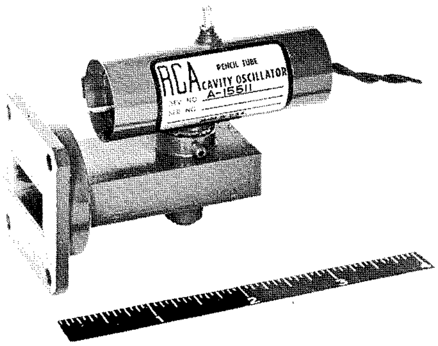
Fig. 2a-d—Representative RCA pencil tube and cavity oscillators.

2a) A-15481 miniature pulsed, S-band oscillator

2b) A-15487/15488 L-band MOPA chain

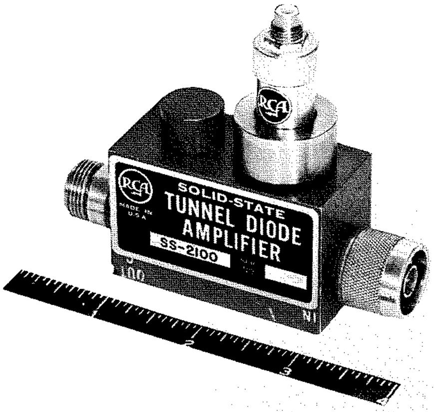


2c) A-15506 S-band oscillator

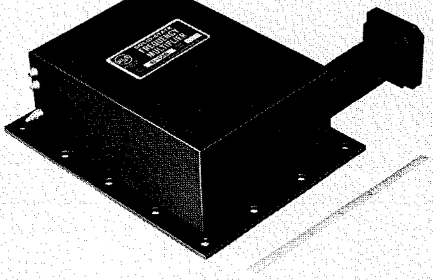


2d) A-15511 X-band microwave signal source

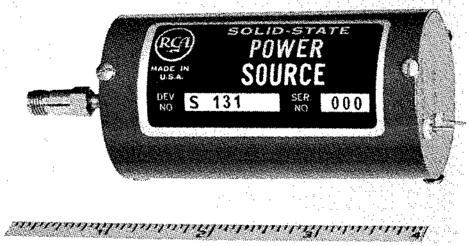
Fig. 2h-k—Representative RCA solid-state microwave components.



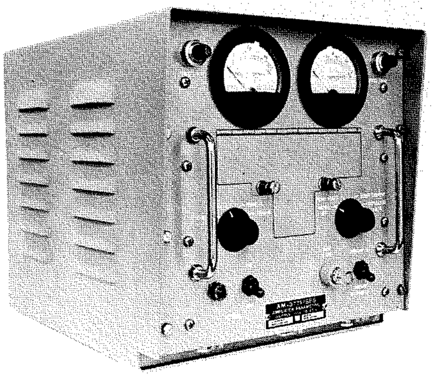
2h) SS-2100 tunnel-diode amplifier with circulator



2i) S-126 X-band frequency doubler



2j) S-131 L-band power source



2k) AM-3771/SPS parametric amplifier system

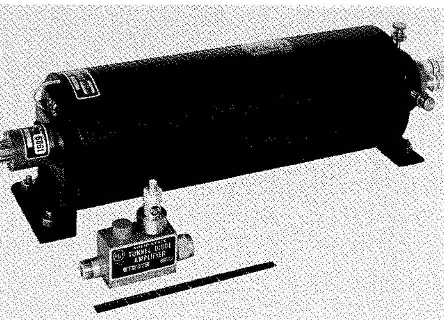


Fig. 3a—Size comparison: conventional and solid-state G-band low-noise amplifier.

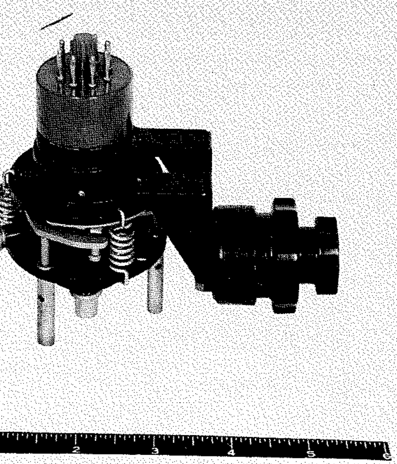


Fig. 3b—Size comparison: conventional and solid-state X-band low-level oscillator.

TABLE II—Commercial Uses of Microwave Components

Radar
Cooking
Data Processing
Radio
Television
Radio Relay
Therapy
Industrial Heating
Accelerators — sterilization, induction of chemical reactions, nuclear research, generation of x-rays.
Science — spectroscopy, atomic clocks.

ments of such applications as: communications, navigational, and weather satellites; scientific space probes and space observatories; and manned space vehicles and specialized types of military satellites.

Phased-Array Applications

Phased-array systems have moved to the forefront of radar development because they permit the simultaneous processing of multiple targets that is required for space and defensive surveillance as well as other modern radar applications. Although the phased-array systems are more complex and their initial cost is substantially greater, they by far outperform conventional radars in relation to performance capabilities versus overall cost. Nevertheless, phased-arrays require microwave components in vast quantities (up to several thousand per installation), and such systems can become practical only if a reasonable relationship can be maintained between performance and cost. This requirement necessitates the attainment of the optimum combinations of such factors as sensitivity, phase control, life, and cost for receiver elements and of efficiency, life, size, and cost in transmitting devices.

Cost must be reduced by at least an order of magnitude in systems that are required to have an assured operating life of many thousands of hours. The impact of reliability and life requirements is evident from the estimate that installation and 5-year maintenance costs for a phased-array system will be five times the initial cost of the equipment! The prime uses of phased-array systems will be in the following applications:

1) *Space surveillance:* rapidly increasing number of objects to be tracked.
2) *Missile defense:* phased array is the dominant element of a missile defense system
3) *Shipborne air defense:* only practical solution for restricted space available
4) *Tactical land-based air defense:* requires considerable miniaturization for practical phased-array system

Other possible applications might be:

1) *Missile detection:* BMEWS now doing job. Phased arrays may be eventual replacement, or addition.
2) *Air defense:* Elaborate system presently in use. Phased arrays may be eventual replacement, though not too likely.

Commercial Applications

The dream of every microwave-component business is to participate in a commercial system having a life span of several years:

1) *Communications satellite systems* offer such an opportunity, as do radio relay systems which consist of line-of-sight repeater chains having an over-all span of several thousand miles and appearing all over the globe.
2) *Computers,* too, will require microwave techniques to handle the sharply increasing amount of information at extremely high speeds.
3) *Commercial radar,* such as airborne-weather, airport-surveillance and navigational-marine radar systems, represents a prime application.
4) *Microwave devices for cooking* in both homes and commercial establishments is also a growing market.

HOW IS THE INTERPLAY BETWEEN SYSTEM AND COMPONENT DESIGNER DEVELOPING?

As the nature of military electronics changes to much larger, more sophisticated equipments, which often include entire interrelated defense systems, the systems engineer finds himself fully occupied with the task of specifying the multitude of black boxes and their interplay. Where possible, he will attempt to procure the black boxes, because he cannot hope to lead with the latest developments in every one of the many specialties required in his system. He will therefore look for a supplier who:

1) specializes to the point that allows him to continuously advance the state of the art and is in a position to deliver advanced devices;
2) has a sizable business in the specialty and can provide reinvestment in development, excellent facilitation, and the necessary upkeep of skills;
3) and finally can reproduce the product uniformly and economically with high standards of quality.

The microwave-component manufacturer is such a supplier who can work with and support the system designer. The microwave industry, more than any other electronic specialty, has required device-circuit integration since its inception. The early use of microwaves represented an extension of the capabilities of lower-frequency devices, by shrinking the circuit elements, to obtain performance at higher frequencies. As the limits of circuit elements placed outside the electron tubes were reached, the microwave-component engineer took over and designed a portion of, or all, the circuit inside the vacuum envelope thus allowing a considerable step-up in performance. This integration of tube and circuit elements, however, had shortcomings which have determined the nature of the microwave-component industry. While the integration of circuitry and electron-tube elements permitted optimum performance to be achieved and relieved the systems engineer of considerable circuit design, fabrication, and integration effort, it also restricted the performance of the integral package to the design objectives, and a multitude of types was now needed to cover the broad scope of applications. As a result, the business became very much one of "customized devices" for specific applications. It was not possible (as it is with receiving or power tubes) to develop a line of microwave tubes, and then to standardize on a few types and produce them in large quantities to be sold from inventories all over the country. The microwave tube was customized for a special system, built on request, and shipped directly to the system for which it was designed.

TABLE IV—Performance Characterists of High-Efficiency Medium-Power Microwave Amplifiers

Typical performance: 10-watt power output at 5 Gc

Type	Gain, db	Efficiency, %	Band width	Life, hours
Traveling-Wave Tube	50	20 (wideband) 40 (narrowband)	octave	>30,000
Klystron	30	35	½%	∾ 20,000
Crossed-Field Amplifier	20	60	10%	∾ 10,000

As the microwave art grew, the performance requirements increased, adding to device complexity and further reducing the probability of multiple-system use.

A few years ago, the solid-state art had advanced sufficiently to permit operation at microwave frequencies. Potential gains in performance, size, weight, reliability, and cost attracted the attention of systems and components engineers. Because the component engineers required some lead time to evolve new products, the systems engineer was forced to develop his own components to remain competitive.

Today, the microwave-component supplier is ready, having become a specialist in three fields: microwave circuitry, electron tubes, and solid-state devices.

TABLE III—Comparison of Low-Noise Microwave Amplifiers

Type	Noise*, °K	Bandwidth,* Mc	Size, ft³	Approx. Cost	Operating Temp.
Maser: at 4.2°K)	10	15	2	$10,000	super-cooled
Parametric Amplifier:					
at 4.2°K	15	100	1	$10,000	
at 77°K	100	100	1	$8,000	
at room temperature	200	100	½	$4,000	room temp
Tunnel-Diode Amplifier	500	600	1/300	$400	
Traveling-Wave Tube	300	2000	1/10	$1000	

* Typical performance at 5 Gc

He not only has the knowhow to develop complex and sophisticated black boxes for the system's designer and, with his specialized facilities, to deliver economical, high-quality units, but he can also continue their evolution and advance the state of the art.

Thus, the interplay between system and component designer is rapidly changing from one of competition to that of cooperation.

WHAT PERFORMANCE IS EXPECTED OF MICROWAVE DEVICES IN THE NEAR FUTURE?

The attainment of ever lower system noise figures in receivers and of higher and higher operating efficiencies in transmitting devices is the continuous goal of the systems designer. Additional frequency channels must be made available, at a faster rate than ever before, to accommodate new types of systems. There is an increasing trend in system design for equipment and components which are able to perform several functions simultaneously. Life and reliability requirements have already advanced far beyond points that only a few years ago were considered the ultimate for practical devices. These and other factors indicate that the performance capabilities of microwave devices will have to be greatly extended if they are to meet the requirements of future systems. However, the largest factor in expanding microwave applications will be spectacular reductions in costs of new devices.

Noise

The strong competition that now exists among traveling-wave tubes, parametric amplifiers, masers and tunnel-diode amplifiers for sockets in low-noise microwave receivers is expected to continue. (In the near future, transistors are also expected to offer competition for input-amplifier sockets in the lower microwave region.) Recent experiments at RCA Laboratories have demonstrated traveling-wave-tube noise figures as low as 1 decibel. This low noise figure, when coupled with the outstanding gain, power, and bandwidth performance of which it is capable, should enable the traveling-wave tube to capture some sockets that now would normally be filled by parametric amplifiers or masers. Tunnel-diode amplifiers, because of their small size, simplicity, and low cost, will find use in many sockets for applications in which noise figures of 3 to 5 decibels are adequate, while masers and parametric amplifiers will continue to be used in applications that require very high sensitivity.

Table III lists typical performance data and other features of the four types of low-noise microwave amplifiers.

Efficiency

The near future will see a real competitive runoff between long-life, lightweight traveling-wave tubes and klystrons for space applications. These devices operate with efficiencies of about 30 per cent today, and that the efficiency will be improved to the 40 to 60 per cent range seems very probable. Present space applications do not require the wide bandwidth that the traveling-wave tube can offer, but they do require more bandwidth than klystrons can now efficiently provide. Table IV lists typical performance characteristics for traveling-wave tubes, klystrons, and crossed-field amplifiers.

Size

Solid-state technology offers the most exciting promise for the generation of microwave devices of greatly reduced weight and size and at a substantially lower cost. For example, combinations of transistors and varactors can be integrated into very compact, lightweight power sources and amplifiers capable of supplying tens of watts of power output at frequencies up to about 1,000 Mc and several watts up to about 10,000 Mc. In these units, circuit elements and solid-state components are integrated in an optimum arrangement that will provide high efficiency, gain, and power levels at dissipation levels compatible with long life. The circuits will capture a large portion of the microwave market in receivers and other low-level applications as well as in moderate-power amplifiers and transmitters. The integration of passive and active components will be greatly advanced in these circuits, and their low cost potential will initiate a number of new microwave uses and applications of considerable quantity.

Fig. 3 shows the sizes of various types of solid-state microwave devices relative to those of other types of devices used for similar applications.

Frequency

The growing electronics industry had not nearly used all the available microwave spectrum when explorations of optical frequencies opened a huge new spectrum to future applications. The availability of this new spectrum has resulted in a partial leap-frogging over the millimeter-wave range, which has thus been left for later consideration. One of the most significant features of optical frequencies is their realistic potential, through microwave modulation and deflection, to provide a super information capacity. The availability of frequency spectrum will no longer be a limiting consideration for new applications for a long time to come.

Multiple Performance

Sophisticated systems because of ever-increasing performance requirements must often employ complex devices capable of multiple-function operation. A single device may be required to perform functions such as low-noise wideband amplification over wide dynamic ranges, variable signal delays, phase control, and high-level detection. Because of multiple-performance requirements, solid-state integrated-circuit techniques will eventually be used to produce complete modules for most microwave systems.

Life

Gone forever are the days of World War II when one considered himself lucky to get a few tens of hours of life from a microwave device. Not only has the life of the devices been greatly increased, but their capability to withstand very severe environmental conditions has been substantially improved. Entirely new concepts of reliability have been developed to assure that *every* device launched into space will perform as required, and these concepts will have beneficial effects on all future devices, including those for optimum-economic-cost commercial systems.

The real need for long life and high reliability of microwave solid-state devices is now starting to be fulfilled.

Until recently 10,000 hours of life was considered outstanding for a microwave device. Tubes developed for long-life space applications will have field (or rather space) lives of 50,000 to 100,000 hours. These tubes will subsequently be replaced by solid-state devices having life capacities that are multiples of the tube life-hour values.

CONCLUSIONS

It has been shown that the microwave business will see continued expansion of applications. Explosive invention will offer much stimulation to the developer and a challenge to management attempting to perform profitably despite rapid obsolescence.

The component designer can now offer the systems engineer technically highly skilled service to support him in the development of complex, sophisticated systems. Breakthroughs in performance will continue in all areas (power, frequency, noise, life, size, etc.), especially stimulated by the rapidly developing microwave solid-state art, and dramatic cost reductions will generate a quantity of new applications.

LOW- AND MEDIUM-POWER TRAVELING-WAVE TUBES AS VERSATILE BROADBAND MICROWAVE AMPLIFIERS

A Review

This paper discusses the extensive role of the traveling-wave tube—specifically of low- and medium-power types—as a microwave frequency amplifier. The exceptional capabilities of this tube in comparison to conventional microwave amplifiers are pointed out, the operation of the tube is briefly explained, and the major applications and the environmental capabilities of the device are described. The emphasis is on the exceptional versatility of this tube which has resulted in its extensive use as a microwave-systems amplifier in communications, radar-surveillance, electronic-countermeasures, and a variety of other complex microwave applications.

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R. W. McMurrough

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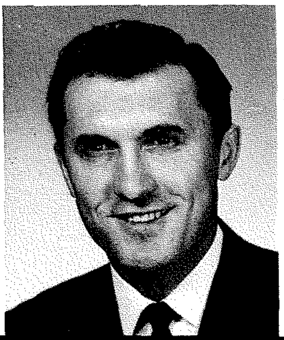

H. J. Wolkstein

The large gain-bandwidth product and the excellent high-frequency capabilities of traveling-wave tubes make these devices ideally suited for use as microwave-frequency amplifiers. In these tubes, the limitations that restrict the gain and bandwidth of the conventional gridded type of amplifier at microwave frequencies have been overcome. Traveling-wave tubes are exceptionally versatile and therefore can provide the required amplification in many widely different microwave-systems applications. These tubes have contributed immeasurably to the progress made in successfully meeting the ever-growing requirements of complex microwave systems. Conversely, the increased systems demands have greatly accelerated technological advances in traveling-wave tubes and thus have directly contributed to significant improvements in their performance characteristics, environmental capabilities, and reliability.

TWT's VS. CONVENTIONAL AMPLIFIERS

The important figure of merit of a microwave-system amplifier is its gain-bandwidth product. Prior to the development of the traveling-wave tube, practical microwave amplifiers rarely provided bandwidths greater than 5%. The traveling-wave tube can provide microwave-frequency bandwidths of more than 100%. Table I lists typical performance characteristics for gridded amplifier tubes (including conventional, frame-grid, and pencil tube) and for traveling-wave tubes. These data show that the figures of merit of typical, commercially available traveling-wave tubes are several orders of a magnitude higher than those of the best gridded tubes.

Electron transit time, the high output impedance necessary to assure adequate circuit gain, the lead inductance, and the plate and grid shunt capacitance all combine to limit the response of conventional gridded-tube amplifiers to a relatively low frequency. Several techniques may be used to partially overcome these limitations, but one factor is improved at the expense of another. For example, the response of conventional-tube amplifiers can be extended into the 1-to-3-Gc range by the use of resonant input and output circuits to "tune out" the shunt capacitance; the circuit bandwidth, however, will be severely limited (to less than 2.5%) by these resonant circuits.

The pencil tube uses special construction techniques to improve its high-frequency capabilities. The electrodes are very closely spaced to reduce transit time and to increase transconductance, and coaxial circuits and tuning cavities

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are added to the tube as integral parts of the vacuum envelope to reduce the shunt capacitance. As a result of these features, the RCA line of gridded pencil tubes will operate satisfactorily at frequencies up to 6 Gc and provide instantaneous bandwidths of about 50 Mc.

The klystron uses the transit-time effect, which limits the high-frequency response in conventional gridded amplifier tubes, to obtain high-level microwave-signal amplification. The high gain results from the mutual and cumulative interaction between the microwave signal and the electron beam in a very short interaction gap, or cavity, in the region between the cathode and the anode. The amplifier bandwidth, however, is again severely limited, because of the large voltage-to-energy ratio and the high circuit impedance of the klystron. If staggered tuned cavities are used in the klystron amplifier, an instantaneous bandwidth of about 10 to 20% is possible.

In the traveling-wave tube, high gain over microwave bandwidths in excess of an octave is achieved as a result of the cumulative, nonresonant interaction between the electron beam and the RF signal wave over an extended interaction region. This tube, which was developed during the early 1940's (as a result of separate efforts by N. E. Lindenblad, R. Kompfner, and J. R. Pierce), has very capably filled the void for a broadband, high-gain microwave amplifier.

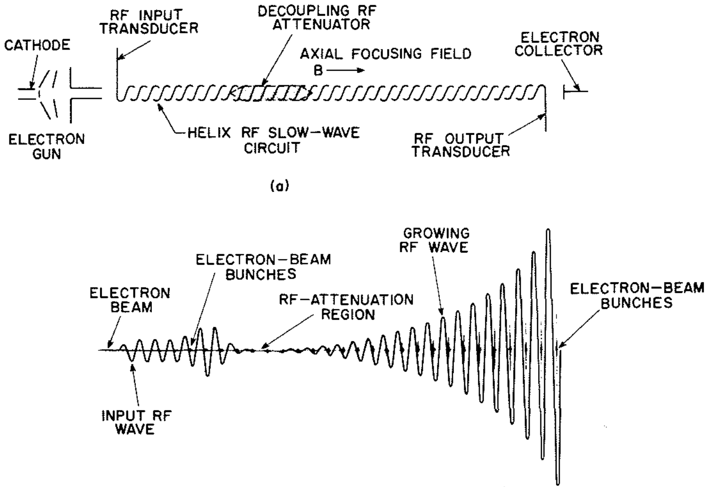


Fig. 1—Operation of TWT: a) schematic of the main parts; b) bunched beam and the growing RF wave caused by interaction between the beam and the wave.

OPERATION OF THE TWT

A schematic of the basic traveling-wave tube is shown in Fig. 1a. The tube consists of an electron gun which provides the electron beam, a helical transmission line (slow-wave interaction circuit) having the desired RF-propagation characteristics, input and output RF-energy ports, and an electron-beam collector. Amplification is achieved as a result of the mutual interaction between the RF signal wave and the beam, which is uniquely accumulative over the entire length of the helix. Because the interaction occurs in a low-impedance circuit, the amount of energy stored is negligible; broadband frequency response and high gain can, therefore, be achieved simultaneously.

The input signal is applied to the helix which is wound so that the signal wave is propagated along its length at a slightly slower rate than the beam velocity. This wave causes velocity modulation of the beam as shown in Fig. 1b. As the beam moves through the extended interaction region, this velocity modulation results in beam density bunching. The bunched beam, as indicated, causes a growing rf wave to appear in the output section of the helix. The RF amplification is realized through extraction of kinetic energy from the electron beam.

To insure stable operation, a decoupling attenuator is used between RF input and output ports. This device, applied to the helix as shown, provides RF isolation between the input and output ports, but does not affect the bunched beam which again initiates an exponentially growing RF wave after passing under the attenuator.

The electron beam, unless restrained, tends to diverge due to the mutual repulsion of the electrons. Therefore, an axial focusing field is required to maintain a constant beam diameter in the slow-wave interaction region. The focusing circuit may be any one of several types (e.g., solenoid, permanent magnet, periodic permanent magnet, or electrostatic) depending on the specific parameters and application. As it interacts, the wave grows exponentially at the expense of the dc energy of the beam. Since the helix is a low-impedance broadband circuit, this high gain is achievable over an octave or more. Typical gain-bandwidth products of 10^5 to 10^6 Mc are common.

TABLE I — Typical Performance Characteristics of Gridded Tubes and of Traveling-Wave Tubes

Conventional Tubes	Type of Construction	Transconductance, μmhos	Total Equivalent Shunt Capacitance, pf	Figure of Merit Gain-Bandwidth Product, Mc
6AG7	Gridded	12,000	20	90
6AK5	Gridded	5,000	7	120
6JC6	Frame Grid	17,600	11.5	264
6EJ7	Frame Grid	19,000	12.8	236
Pencil Tubes				
A15288	Gridded	13,500	2.76	780
7554	Gridded	14,000	3.1	720
Traveling-Wave Tubes	Small-Signal Power Gain, db	Voltage Gain	Bandwidth of Actual Operation, Gc	Voltage Gain-Bandwidth Product, Mc
A-1203	30	31.6	4	126,000
A-1300	35	56.2	2.3	130,000
A-1206	50	316	5	1,580,000

MAJOR TWT APPLICATIONS

The characteristics of the traveling-wave tube have been widely adapted to systems requiring broadband, high-gain characteristics and a large dynamic range as will now be reviewed.

Front-End Receiver Components

The large instantaneous bandwidth of a traveling-wave tube makes it particularly attractive as a front-end component for microwave receivers. The minimum detectable signal, input-signal drive range, and instantaneous bandwidth requirement of the system largely determine the type of active amplifier (or detector) that can be used. Table II indicates the capability of various traveling-wave tubes relative to other low-noise amplifiers or detectors.

The dynamic characteristics of the traveling-wave tube are shown in Fig. 2. The threshold of useful operation is determined by the bandwidth and noise figure of the tube. The dynamic range is that region between the threshold input level and the input at which there is departure from small-signal or linear gain. The gain continues to decrease for approximately 6 db to the point of saturated power output. In the region of additional drive, power output will decrease with periodic minor peaks, which are generally lower than that of the saturation point. This fall-off in power output can be minimized in specially designed "limiter" traveling-wave tubes.

The noise figure (*NF*) for any circuit

TABLE II — Relative Capabilities of Traveling-Wave Tubes and Other Low-Noise Amplifiers and Detectors

Device	Noise Figure, db	Gain, db	Instantaneous Bandwidth, % of octave	Saturated Power Output	Special Requirements
Traveling-wave tube A-1207V	4.5	25	30	3 mw	Solenoid—weight 20 lbs. Permanent magnet.
Traveling-wave tube A-1207V4	1.0	20	10	1 mw	Solenoid in liquid nitrogen. 4,500 gauss field.
Traveling-wave tube A-1173	11-15	30	100	10 mw	Periodic permanent magnet.
Triode 416A	14-20	10	1-5		
Crystal mixer	8-15	-6 to -8	0.5		
SS parametric analyzer	1-6	15-25	1-10	-15 dbm	Requires pump RF generator.
Maser	0.03	30	0.1	3 μw	Requires cryostat.
Tunnel diode amplifier	4-5	20	10	-20 dbm	DC supply (low impedance).

element is defined by two signal-to-noise ratios (S/N):

$$NF = \frac{(S/N)_{input}}{(S/N)_{output}}$$

The equivalent noise input power of the traveling-wave tube, which is a measure of the threshold sensitivity, is the product of the thermal noise which is inherent in a well-matched source at room temperature ($KT/Mc = -114$ dbm/Mc), the noise figure (NF), and bandwidth in megacycles (BW) of the tube. If it is expressed in decibels, the integrated noise input power is given by

$$Noise\ input\ power =$$

$$KT/Mc + NF + BW$$

It is reasonable to assume that the minimum detectable signal of a receiver would be approximately equal in amplitude to the total integrated equivalent noise power of the tube. (This condition is known as the *tangential sensitivity.*) The large bandwidth of the traveling-wave tube penalizes the system, because the large bandwidth results in additional integrated noise and thereby reduces signal sensitivity.

The gain of the tube, of course, also adds to the equivalent output power; this condition (Fig. 2) reduces the dynamic range of the tube.

Cascaded TWT's

In many systems several traveling-wave tubes are used in cascade. For this application, it is necessary to determine several factors which affect the overall system noise figure and system dynamic range. The system noise figure (in decibels) is given by

$$NF_{system} =$$

$$NF_{1st\ stage} + \frac{NF_{2nd\ stage} - 1}{G_{1st\ stage}}$$

With a large first-stage gain, the effects of noise produced in the second and subsequent stages are negligible.

The integrated power output of a cascaded chain has been increased by the gain of each stage in the chain. For example, Fig. 3 shows a typical application, in which three traveling-wave tubes are cascaded to provide 114 db of gain and which can provide 1-kw output power. The integrated noise power output over a 1-Gc bandwidth is approximately 100 watts. This output power indicates that dynamic range to saturation for the entire system is only 10 db, while the input stage has an overall range of 34 db to saturation. If wider dynamic range is required before saturation is reached, the gain, bandwidth, or noise figure of the system would have to be reduced, or a tube with an output capability larger than a kilowatt would have to be used. For example, a reduction in the noise figure of the input stage by 10 db would add 10 db to the dynamic range. It is apparent, however, where an optimum linear dynamic range is sought, that cascaded stages in an amplifier chain must be designed so that all tubes approach saturation together.

In many applications linear performance can be sacrificed, and the signal detection capability is preserved over an extremely wide input drive range while a minimum power-output capability is maintained. This characteristic is important for many radar and countermeasure receiver-transmitter chains and for certain microwave frequency repeaters. This characteristic will be achieved if good power overdrive characteristics are obtained for individual tubes and if the stages are designed so that their respective power output characteristics are complementary.

A two-stage traveling-wave-tube limiter chain, designed for X-band operation, which uses tubes having complementary characteristics that emphasize a wide power overdrive capability, is shown in Fig. 4. (A 10-db isolation pad is used between stages to augment overdrive characteristics and enhance stability.) In this chain, the second tube must be driven to saturation before the first tube, and the low-level first stage must pass through its peak power range before the last stage is driven far enough into the overdrive region so that its power again diminishes.

Fig. 2—Dynamic characteristics of TWT.

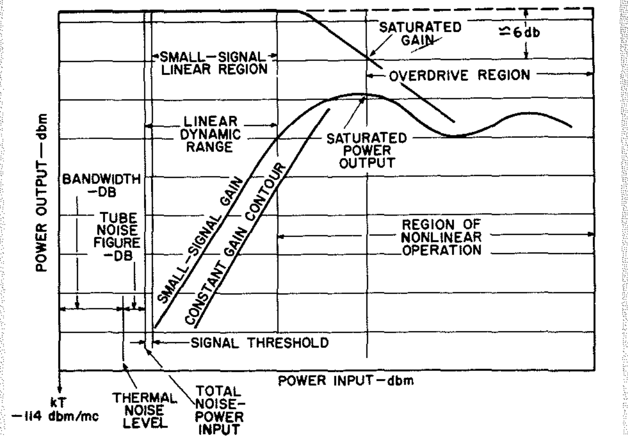
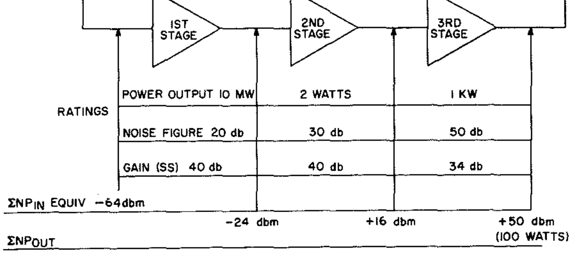


Fig. 3—TWT's operating in cascade, showing buildup of integrated-noise-power output.



Complementary circuits designed in this fashion can produce a substantially constant power output within reasonable limits (about 3 db) over a 50-db input-power drive range. In addition, noise power does not significantly interfere with the dynamic operation of such nonlinear systems.

Communications TWT's

Communications traveling-wave tubes are a class of tubes used as transmitter amplifiers in microwave radio relay systems. The tubes are used in each link of the relay system. The links may be ground-based or may be orbiting satellites. Since a microwave radio link must amplify an incoming rf signal with a minimum amount of distortion and reradiate the amplified signal to the next link, the communications traveling-wave tube, therefore, must be a high-fidelity amplifier. To insure high-grade transmission performance, the communication traveling-wave tube must have the following characteristics:

1) The wideband input and output couplers must provide an excellent hot match.
2) A small intrinsic noise factor without any spurious noise signals in the vicinity of the carrier is usually required.
3) The gain should remain essentially flat over the signal channel.
4) The gain response should be linear in the operating region.

Ground-Based Applications

Communication traveling-wave tubes designed specifically for ground-based applications have certain basic differences when compared to tubes designed for satellite applications. Because ground-based systems are multi-hop systems and satellite systems are usually only one-hop systems, the electrical performance requirements of ground-based systems are the more stringent. In addition, because of the larger number of tubes required, low cost and ease of adjustment and replacement are important considerations in ground-based equipment. Long life and good reliability are also of major importance.

The RCA-7642 traveling-wave tube (Fig. 5) is used in the MM-600-2 microwave relay system, a 600-channel, multi-hop system designed by the former RCA Industrial Electronics Products Division and now being produced by RCA Victor Ltd. of Canada. The specifications for the 7642 are listed below to illustrate the type of performance required of a ground-based communication traveling-wave tube:

Frequency range	1.7 - 2.3 Gc
Power output	18 watts min.
Gain (small-signal)	30 db
Gain variation	0.02 db/Mc
Input match (hot)	1.4:1 max.
Output match (hot)	1.5:1 max.
Noise @ 18-watt output	48 db max.
Harmonics @ 18-watt output	9.5 db below fundamental

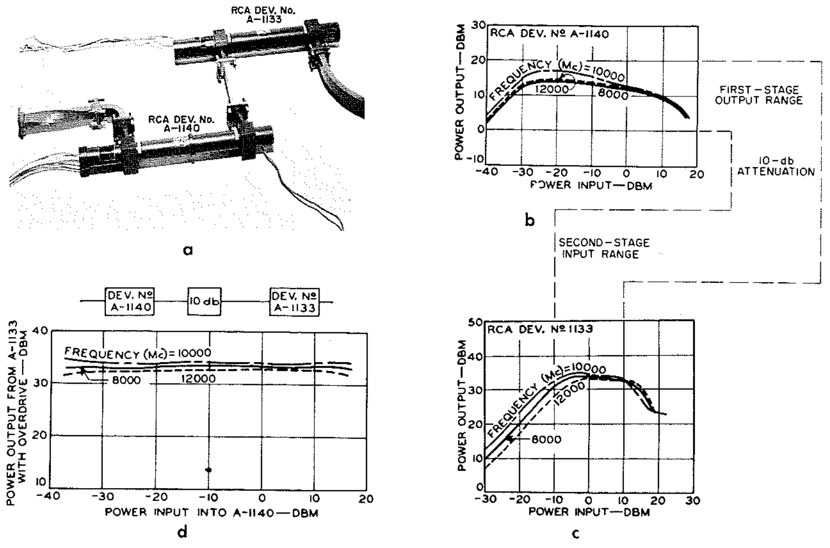


Fig. 4—Two-TWT limiter chain: a) the two TWT's with a 10-db attenuator between them; b) first-stage (A-1140) power output vs. power input; c) second-stage (A-1133) power output vs. power input; output range of first stage is aligned to show input range of second stage reduced 10 db by attenuator; d) power output of the second stage, A in (c), plotted against the range of power input to the first stage; the Chain limits the 50-lb variation of the power input to a variation of 3 db.

Satellite Applications

There are four basic requirements that satellite traveling-wave tubes must satisfy to much greater degree than tubes designed for ground-based communication systems. These are: 1) extreme ruggedness, 2) excellent reliability and a long life expectancy, 3) high overall efficiency, and 4) light weight.

The satellite traveling-wave tube must be extremely rugged to withstand the rigors of launch and of operation in the space environment. To operate successfully under these conditions, the tube must be designed to withstand severe g-forces of vibration and shock, hot and cold temperature extremes, and Van Allen belt radiation.

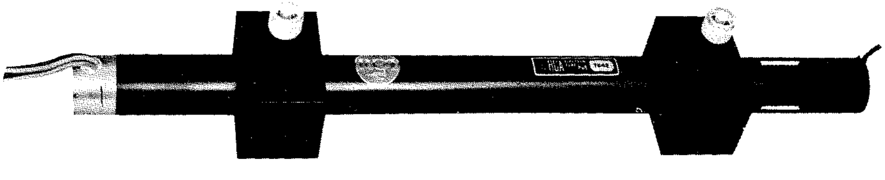
Good reliability and long life are requisites because of the costliness of launching a satellite and the inaccessibily of the equipment for repairs.

The efficiency of the traveling-wave tube must be high because primary power is at a premium in satellites. Good efficiency has been obtained by careful and painstaking design of the collector, helix circuit, and heater. Tubes having overall efficiencies greater than twenty-eight per cent are now being produced.

The reliability of the satellite traveling-wave tube has been proven by the performance of the tubes in the various communication satellites that have been orbited and successfully operated. The tubes used in the RELAY satellites which were designed in 1961 have performed well since launch and have shown very little performance degradation. The two tubes in RELAY I have been in orbit for 17 months and the two tubes in RELAY II have been in orbit for 4 months, as of May 18, 1964. The specifications (presented in Ref. 1) of the RELAY satellite tube provide as an indication of the performance required.

Specifications for the most modern satellite tubes, which have recently been developed or are currently under development at RCA and other companies, show the degree of sophistication attained by these tubes in a few years.

Fig. 5—A 20-watt, S-band TWT (RCA 7642).



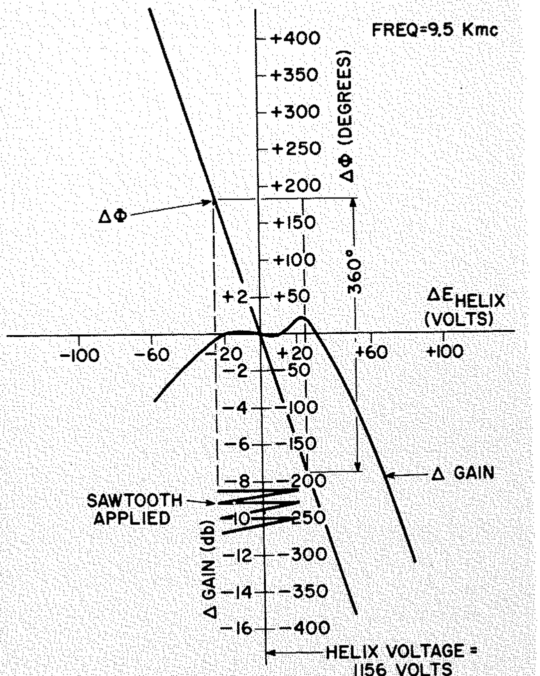


Fig. 6—Change in small-signal gain and phase vs. changes in helix voltage applied to a TWT. Phase shift and gain compression in serrodyne operation.

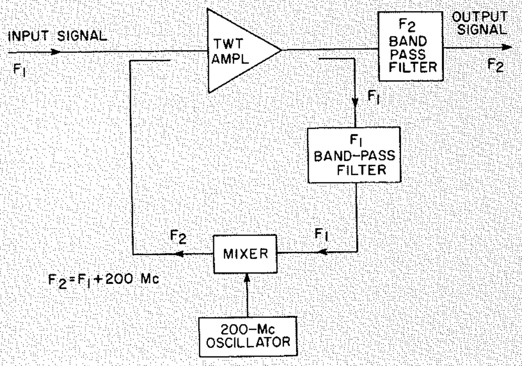


Fig. 7—Typical communications receiver using a TWT in a reflex circuit.

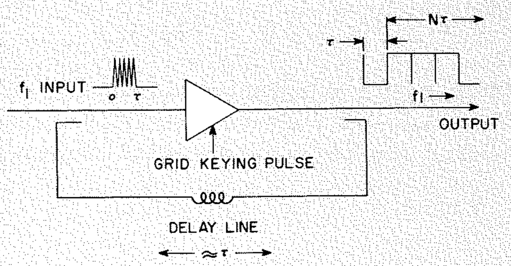


Fig. 8—RF memory system using a TWT.

Tubes that supply up to 10 watts of power in s-band at an overall efficiency of 30% are being made in packages that have a weight in the order of one pound. The tubes are extremely rugged and have life expectancies of 5 years or more.

Phase-Tracking Applications

The use of traveling-wave tubes as active elements in phased-array radars is particularly attractive because of the large gain-bandwidth products of these tubes. The array radars are used in the search, detection, and tracking of a large number of targets; therefore, they require a very close tube-to-tube reproducibility of RF characteristics. The main problem in the design and production of traveling-wave tubes for phase-tracking applications is that each tube in the array must provide essentially the same amount of phase delay. Because traveling-wave tubes have electrical lengths of many thousands of degrees, the problem of insuring reproducible tube-to-tube phase characteristics is of considerable magnitude. The problem is solved by very closely controlling all mechanical dimensions of the tube—especially those of the slow-wave circuit—and by attaching external phase compensators to each tube.

SPECIAL APPLICATIONS

Traveling-wave tubes are often required to provide unique types of operation to meet specialized requirements of modern microwave systems. The modes of operation described in the following paragraphs are but a representative sample of the specialized applications of traveling-wave tubes.

Serrodyning and Synchrodyning

In some applications (e.g., radio relay systems, countermeasures systems, and certain types of reflex circuits), traveling-wave tubes are used to shift the frequency of an RF signal; either the serrodyning or the synchrodyning process is used to effect this frequency translation. In the serrodyning process, the frequency shift is produced by the sawtooth modulation of the helix voltage; in the synchrodyning process, the frequency shift in the result of a sinusoidal modulation of the helix voltage. Each process varies the transit time of the electron bunches in the electron beam and, in this way, changes the frequency of the output signal obtained from the helix. A negative increase in the modulating voltage results in a higher frequency, and a positive increase in this voltage causes the frequency to decrease. Although the frequency translation is accomplished by either method, the serrodyning process is the more widely used, primarily because it produces better carrier frequency suppression.

Fig. 6 illustrates the effects of serrodyning on the output signal of a traveling-wave tube. Because helix-voltage variations modulate the transit time of the electron bunches in the beam, serrodyning is essentially a phase-modulation process. The amplitude of the sawtooth serrodyning voltage determines the amount of phase shift in the output signal, while the frequency of the sawtooth voltage determines the rate at which the phase is shifted. The flyback time must be extremely short so that distortion is held to a minimum. For the case shown in Fig. 6, a 40-volt peak-to-peak sawtooth signal produces a phase shift of 360° with gain compression <1 db.

Reflex-Circuit Application

Traveling-wave tubes are used in many different types of unique reflex, or feedback, circuits. One such circuit for a typical communications receiver is shown in Fig. 7. This application is made possible by the wide bandwidth and the large dynamic range of the traveling-wave tube. Because of these characteristics, the tube can accept and amplify an initially low-level input signal and then can re-amplify this signal after it has been shifted in frequency. This stable feedback circuit, thus, amplifies the input signal twice before it is applied to the next stage.

The wideband traveling-wave amplifier has been used extensively as an RF-storage reflex memory device. In this application, an RF pulse of good signal purity and of the proper duration is applied to the traveling-wave tube; the output of the tube is then fed back and recirculated through an appropriate delay line to the input of the tube. For the circuit shown in Fig. 8, an RF pulse of duration τ, is applied to the tube. while an output pulse having a duration that is a multiple of τ is obtained. Memory recirculation devices covering octave bandwidths and capable of several microseconds of storage have been adapted to special system applications.

Multiple-Coupler TWT's

Traveling-wave tubes that have more than a one output coupler have been built to provide several outputs at widely different linear gain levels.[2] The characteristics of a tube having two output couplers—one of which provides an output at a gain of 30 db and the other, an output at a gain of 10 db—is shown in Fig. 9a. Multiple-output tubes are particularly useful in cases where the input signal varies widely in dynamic range while amplitude linearity of the output must still be preserved. One of the clear advantages of this type of device is that the both signals can be extracted simul-

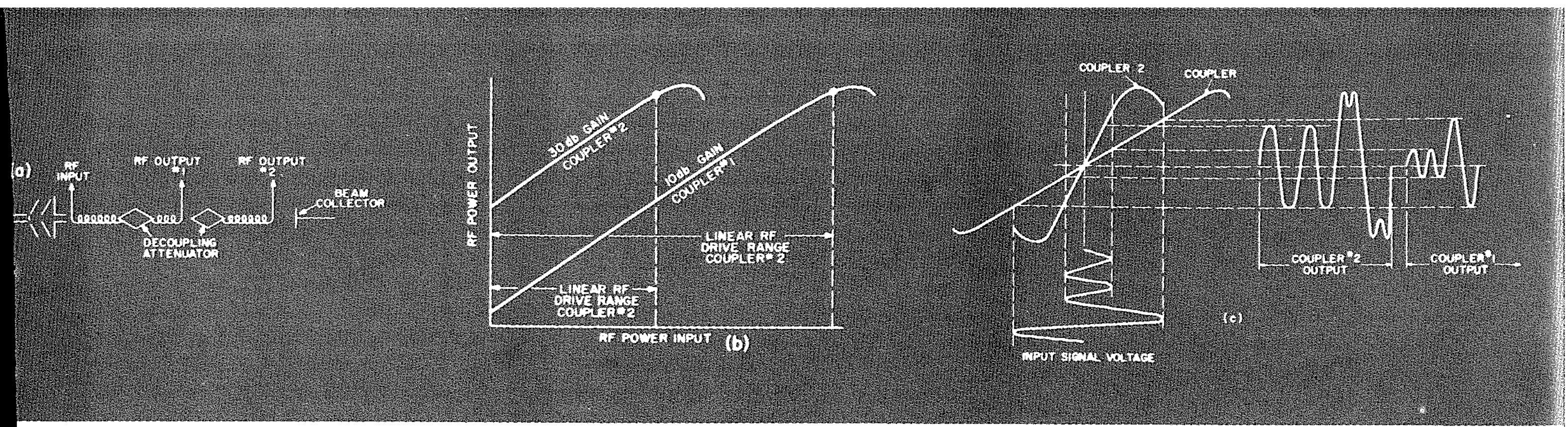


Fig. 9—RCA multicoupler TWT: a) location of couplers; b) RF output at the two couplers plotted against the RF input signal to the tube; c) transfer curves of the two couplers: at coupler 1, undistorted voltage outputs are available from large input signals; at coupler 2, high-gain, undistorted output is taken only from small input signals.

taneously to the same ultimate saturation level without one output effecting the other if the proper design precautions are exercised (i.e., outputs are adequately isolated from one another).

Fig. 9b shows the input voltage-output voltage transfer curves for the traveling-wave tube having two output couplers. These curves clearly show the clipping and harmonic distortion that results when the linear drive-range characteristics are exceeded. The results are somewhat similar to those expected for a conventionally gridded tube in the overdriven state.

If the proper sensing mechanism is used, a system can be designed to shift RF output couplers dynamically so that low-level gain output is used for large overdriving signals (to reduce distortion) and the high-gain output is used for weak signals. This feature would be particularly useful in search-tracking radar systems, because the search function requires good sensitivity and high gain, while the tracking function requires good dynamic range and large-strength signal tolerance.

Multisignal Environment

The operation of any amplifying device is dependent on the entire signal environment included in its band-pass response at any instance. This condition can be particularly troublesome for the traveling-wave tube because of its wide band-pass capabilities. While significant data have been published for the operation of the traveling-wave tube, the multisignal case is usually avoided and the data only deal with a single signal in the band. Such data are clearly inadequate for the device expected to handle multiple signals.

Fig. 10 illustrates the effect that a growing signal has on other constant low-level signals in the band. As indicated, an increasing signal is applied to the traveling-wave tube at 3,600 Mc, while other low-level signals of constant input power (−20 dbm) are monitored. The results clearly show that the gain of the low-level signals are significantly suppressed as the magnitude of the driving signal (at 3,600 Mc) is increased. This phenomena causes amplitude and cross modulation distortion in traveling-wave tubes and must be avoided where linear performance is required.

ENVIRONMENTAL CAPABILITIES

As the electrical performance of the traveling-wave tube has improved, the environmental performance capabilities have also improved markedly. Tubes are now supplied for satellite, missile, and aircraft use that can withstand all the rigors imposed upon them. The three most important factors are discussed below:

Vibration

Because the traveling-wave tube is basically a relatively long, small-diameter cylinder, early tubes were very susceptible to failure during vibration. Constructional techniques that trap and prevent movement of the helix and heater and all other parts have solved the vibration problem. Modern, ruggedized traveling-wave tubes can withstand the following levels of vibration in all three phases:

Sinusoidal	20 g, 20-4,000 cps
Random	0.1 g^2/cps, 20-2,000 cps

Shock and Linear Acceleration

The same techniques used to enable the traveling-wave tube to withstand vibration also increase the resistance of the tube to damage by mechanical shock and linear acceleration. The following levels of shock and acceleration are now easily met by ruggedized traveling-wave tubes:

Shock, 40 g, 6 msec, sawtooth
Linear Acceleration, 50 g

Temperature

The magnet used to focus the electron beam has been the major limiting factor in determining the operating temperature range for the traveling-wave tube. The solenoids that supplied the magnetic field for early tubes had a very narrow operating temperature range. The use of periodic, permanent ferrite magnets resulted in a substantially broader temperature range than was possible with the solenoids. However, the magnetic field strength of these magnets varied with changes in temperature, and the temperature range was therefore limited by the stability requirements for the focusing magnetic field. The use of temperature-compensation techniques with these magnets substantially improved the stability of the magnetic field and, thereby, greatly extended the temperature range of the traveling-wave tube.

In most modern traveling-wave tubes, the magnets are now made of platinum-cobalt or Alnico. The characteristics of these magnets are relatively insensitive to variations in temperature, and it is now possible for traveling-wave tubes to operate over the same temperature ranges of other equipment and systems components. Present-day tubes can operate without any significant change in performance over a temperature range from −65°C to +150°C.

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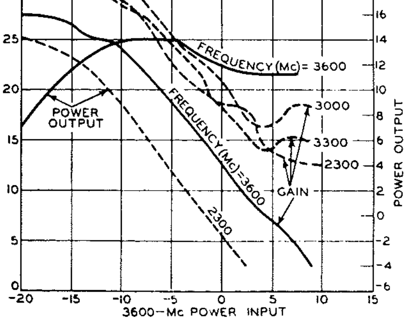
1. The specifications for the RELAY satellite tube are given in the paper, "Satellite Communications Traveling-Wave Tubes at RCA," by F. E. Vaccaro, P. R. Wakefield, and M. J. Schindler, RCA ENGINEER, *this issue*.
2. Traveling-wave tubes of this type are now being developed by the RCA Microwave Tube Operations Dept. under BUSHIPS Contract NObsr-87535.

Fig. 10—Power-output and gain characteristics of a TWT at various frequencies vs. input at a fixed frequency.



RECENT ADVANCES IN TRAVELING-WAVE TUBES FOR COMMUNICATIONS SATELLITES

Satellite communications systems have introduced new requirements for high-efficiency, lightweight, reliable microwave power amplifiers. Among available microwave amplifiers, the traveling-wave tube provides the most favorable overall balance of the particular characteristics required for these satellite applications. During recent years, the RCA Microwave Tube Operations Department has conducted continuous programs to develop high-efficiency, rugged, lightweight, reliable traveling-wave tubes having the performance and environmental capabilities, and long life expectancy required for space use. These programs have led to recent achievements, discussed herein, that only a few years ago were considered unattainable in practical traveling-wave tubes.

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Microwave Tube Operations Department

Electronic Components and Devices, Harrison, N. J.

The assurance of exceedingly long operating life and the inviolable reliability requirement have dictated an evolutionary approach to the development of traveling-wave tubes for space. Each design innovation or improvement considered must be thoroughly evaluated and proved satisfactory in relation to performance, reliability, and life expectancy, before it is incorporated into the overall tube design.

Time and economic limitations of the development program do not usually permit large-quantity statistical evaluations and life tests before the tubes are operated in the satellite system. Short-term tests, selection techniques, and theoretical analyses must, therefore, be devised that provide accurate predictions of reliability.

Since 1958, the RCA Microwave Tube Operations Department (MTOD) has conducted several continuous programs for the development of both groundbased and satellite communications traveling-wave tubes and have established an ultra-modern tube-fabrication facility especially tailored to the production of very reliable, long-life tubes for use in satellite systems.

The chart in Fig. 1 shows the continuing evolution of RCA satellite communications traveling-wave tubes from the development of the 7642 tube for a ground-based communications system to the newest type A1292 satellite tube now in production (commercial type 4056). Because of the overlap of the later programs with the preceding ones, they may all be considered as a single evolutionary process.

EARLY SATELLITE TWT's

The development of three early glass-envelope tubes, the 7642 and the developmental types A-1228 and A-1245, provided the experience necessary to establish the basic criteria and techniques for the design and production of modern satellite traveling-wave tubes.

RCA-7642 Communications TWT

The history of satellite-tube development at RCA began with the development of the RCA-7642. This tube, together with a similar one developed by the Bell Telephone Laboratories, was the first traveling-wave tube designed specifically for application in a microwave communications system, the RCA-manufactured Western Union MM-600 ground-based microwave relay system. Information obtained from system evaluations of this tube proved extremely important, and formed the basic starting point, in the design of reliable, long-life tubes for satellite communications systems. With respect to cathode life, for instance, theoretical calculations had indicated that a cathode loading of 80 milliamperes per square centimeter would yield a long cathode life span. The validity of these calculations has now been amply demonstrated by extensive field operation of many 7642 tubes beyond 2 years, indicating a probability life of several additional years.

First Satellite Communications TWT

The development of the first satellite communications traveling-wave tube, RCA Developmental No. A-1228 was begun (under prime contract AF04-(647)-595) before the 7642 development program was fully completed. The A-1228 tube, which was scaled from the 7642, is operated in the depressed-collector mode to obtain higher efficiency. The tube is cooled by direct conduction and is specifically designed for use in a pressurized transmitter. Because sufficient cathode-loading life-test data had not been acquired before the development of the A-1228 was begun (the cathode life tests on the 7642 tube had not been completed at that time), the cathode loading in the A-1228 tube was limited to 57 ma/cm^2 to provide added assurance of the cathode life required in the satellite tube.

Although the system for which the A-1228 tube was designed was never launched into space, the development of this tube was completed, and it became the first traveling-wave tube qualified to withstand the launching of a satellite and then to provide reliable operation in the space environment. Ten of the finished tubes were placed on life tests, and ten others were delivered, as flight-model tubes.

The ten life-test tubes are operated at an elevated temperature of 80°C at saturated RF power output levels. As of April 1964, these tubes had operated for more than 150,000 hours (total operating hours for all tubes). With but one exception, the tubes are still operating after two years of testing, and practically no change in their performance has been observed. The A-1228 tubes have thereby demonstrated a mean time before failure (MTBF) of 9 years with a 90% confidence level. (The MTBF for the A-1228 tube was determined from the combined life-test hours accumulated by A-1228 tubes and by the A-1245 tubes developed for the Relay communications satellite. This approach is valid because the Relay tube has the same electrical and thermal stress levels as the A-1228.)

RELAY Communications Satellite TWT

The development of the traveling-wave tube, RCA Developmental No. A-1245, for the Relay communications satellite[1,2] was started (under Prime Contract NASA-5-1272) at about the time the A-1228 development program was nearing completion. The A-1245 tube was required to operate at a center frequency of 4.2 Gc and its efficiency was to be higher and its weight substantially less than those of the A-1228 tube. The program allowed only one year before the scheduled flight of the Relay satellite. This time was not sufficient to permit the accumulation of significant life data on the A-1245 tube prior to launch. The approach, therefore, was to scale the A-1245 from the A-1228 in such a way that the electrical and thermal stress levels would be maintained. The life and reliability data accumulated on the A-1228 could then be applied with maximum confidence to the A-1245.

One important step that was taken to

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assure equivalent thermal stress, at least until more data could be accumulated, was to provide the tube with a simulated atmospheric environment. The tube was designed so that it could be enclosed in a lightweight container that was pressurized to one atmosphere of nitrogen. These pressurized-capsule tubes have been operating in the Relay I satellite since December 1962 without any detectable change in performance.

A parallel program was also carried out to reduce the thermal stresses in an unpressurized version of the A-1245 tube. In this way, tube weight could be reduced, and tube reliability would be improved because the possibility of a leak in the pressurized capsule would be avoided. A combination of factors were used to reduce thermal stresses of the unpressurized tube by approximately 100°C, resulting in a tube having thermal stresses comparable to those in the pressurized version and lighter in weight by one pound. A tube of this type has been operating in the Relay II satellite without any detectable change in performance since the satellite was launched in January 1964.

The A-1245 life tests are conducted on tubes operated in a vacuum chamber and with a programmed variation in baseplate temperature. More than 30,000 hours of life and flight operation of the A-1245 tubes have been accumulated without failure. A combination of the A-1245 data with those of the A-1228 results in the 9-year MTBF referred to in the discussion of the A-1228.

METAL-CERAMIC TECHNIQUES

The development of the 7642, A-1228, and A-1245 glass-envelope tubes provided important experience related to the design and production of traveling-wave tubes for satellite communications systems. This experience indicated, however, that a continuous development of new and improved techniques is necessary to insure that the technology for future satellite tubes keeps pace with the requirements of new systems.

The following paragraphs point out the improvements that have been made possible as a result of incorporating metal-ceramic technology. In these studies, an advanced-design tube was developed to prove-out proposed metal-ceramic design techniques; the proven techniques were then applied in the design of a metal-ceramic tube for a specific satellite system. The objective was to develop a tube of this construction capable of operation at significantly higher power densities than an equivalent glass tube, thus permitting higher efficiency and power output to be obtained in a lighter weight tube. Metal-ceramic tubes are also more rugged.

The use of the metal-ceramic structure permits direct conduction of heat from the vacuum envelope through a short metal ceramic section, whereas the glass tube envelope is cooled by radiation or by conduction through gas. The power-handling capabilities of a metal-ceramic tube surrounded by a vacuum can be made approximately an order of magnitude greater than that of an equivalent glass tube. As will be described later, the improved thermal properties of metal-ceramic construction permits smaller size, which leads to a significant reduction in focusing-magnet weight.

All the advantages given above were verified during the development and evaluation of a 10-watt metal-ceramic traveling-wave tube, RCA Developmental No. A-1261, designed for continuous-wave operation from 5.5 to 11 Gc.

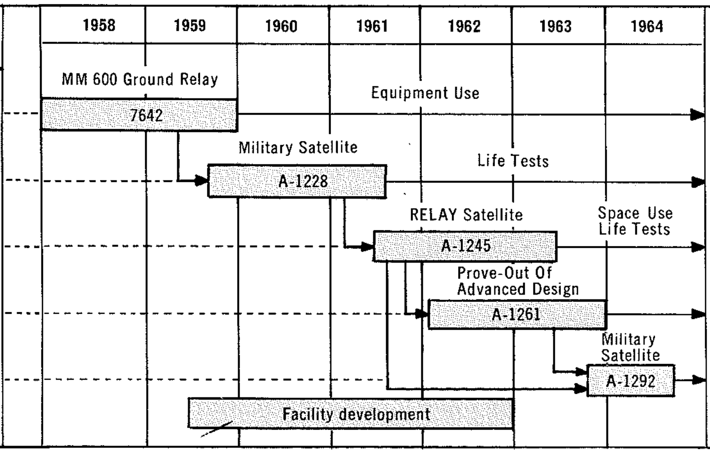
The experience gained during the A-1261 program made possible the development of the RCA A-1292; an extremely rugged, lightweight metal-ceramic traveling-wave tube, capable of supplying 12.8 watts of output power in the s-band frequency range, with an operating efficiency of 30% to 35%. Important design considerations with regard to efficiency and weight are discussed below, followed by the design and performance of the A-1292 tube.

Efficiency Design Considerations

The following discussion assumes that the reader is familiar with the basic operation of a traveling-wave tube. (The basic operation of a traveling-wave tube is briefly described in Ref. 3).

Fig. 1—Evolution of TWT's for satellites at RCA. (Bars indicate period of development.)

RF Power Output (watts)	Center Freq. (Gc)	Overall Eff. (%)	Weight
18	2.0	14	6
11	2.2	20	8
11	4.3	21	3.6
5-10	7.5	15	1.0
2-13	2.2	30	1.7



Major Objectives

Demonstrate microwave communications capabilities of TWTs. Starting point for satellite tubes.

Establish reliability, space ruggedness and high efficiency.

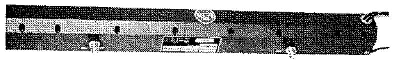
Lighter weight, higher frequency, maintain reliability.

Higher efficiency, lighter weight, wide power range.

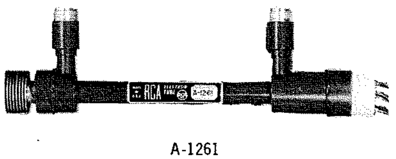
Demonstrate and prove out advancing design and technology features for general use.

Provide space-tube manufacture and test facility, etc.


7642

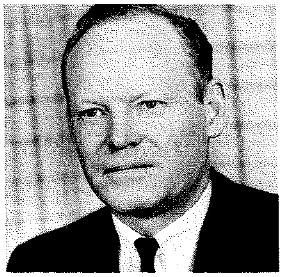

A-1228


A-1245


A-1261


A-1292

F. E. Vaccaro



P. R. Wakefield

Dr. M. J. Schindler

F. E. VACCARO received the BSEE from the University of Tennessee, Knoxville, in 1948 and the MSEE from the Stevens Institute of Technology, Hoboken, N. J., in 1953. He joined RCA as a specialized trainee in 1949, and was assigned to the Microwave Tube Operations Department, Harrison, N. J., in 1950 as a development engineer. In 1953, he became a member of the Technical Staff in the Microwave Applied Research Group at Princeton, N. J., studying the characteristics of wide-range, high-power tunable magnetrons. In 1956, he became an Engineering Leader in the Group, directing the activities of all advanced development projects on magnetrons, traveling-wave tubes, and related fields. In 1959, he was promoted to Manager, Microwave Applied Research, where he supervised work on microwave beam devices. In 1963, he assumed his present position as Manager, TWT and Pencil Tube Engineering. He is a Senior Member of the IEEE.

P. R. WAKEFIELD received his BSEE and BS in Business Administration, both in 1949, from the University of Colorado. He has also taken graduate courses at the Polytechnic Institute of Brooklyn. Mr. Wakefield joined the RCA Microwave Tube Operations Department in 1949. He has done major design work on a series of traveling-wave tubes, including the first commercial traveling-wave tube sold in the United States. In 1955, he was appointed Engineering Leader of the traveling-wave tube design group. In 1957, he was promoted to Manager, Microwave Engineering Services and Administration, and given the responsibility for all project planning and control, including the planning of personnel assignments, control of engineering expenditures, and other general administrative services. He served in this capacity until promoted to his present position in 1960. Mr. Wakefield is currently Manager of Special Projects and Research and Development Liaison. He served as project manager of the program to develop the RCA A-1245 traveling-wave tube used in the RELAY communications satellite. He is a member of the IEEE.

DR. M. J. SCHINDLER received his MS in 1951, and his PhD in 1953, both from the Institute of Technology of Vienna, Austria. From 1951 to 1954, he was a research assistant at the Institute of Technology of Vienna. From 1954 to 1957, he worked at the Tungsram-Watt tube plant in Vienna. From July 1957 to August 1958, he was a research scientist at the Wright Air Development Center. Dr. Schindler joined the RCA Microwave Tube Operations Department in 1958, and has since worked on basic technical problems of traveling-wave tubes and magnetrons. He is an authority on magnetics. Among his major contributions in this field are improvements in periodic-permanent-magnet focusing structures. He has also made unique contributions in a study of thin magnetic films. He is responsible for the design of the magnet structure for the RCA klystrons of the new Stanford Linear Accelerator. As an Acting Group Leader, his major work was development of a hydraulically-tuned magnetron. Since 1963, first as Project Engineer then as Senior Engineer, he has led a group working on development of a traveling-wave tube for a communications satellite. Dr. Schindler is a Senior Member of the IEEE.

The overall efficiency, η, of a high-gain traveling-wave tube is defined as:

$$\eta_{overall} = \frac{P_o}{P_c + P_H + P_I} \times 100\%$$

Where: P_o is the RF power output, P_c is the power dissipated at the collector, P_H is the heater power, and P_I is the power dissipated by beam interception on the helix and electron-gun electrodes.

For a fixed beam power, the RF power output P_o of a traveling-wave tube is maximum when the beam and circuit parameters of the tube are adjusted for the maximum electronic efficiency. This relationship indicates that a specific beam diameter, voltage, and current together with a particular helix diameter and pitch must be chosen to obtain the optimum transfer of energy from the electron beam to the growing RF wave on the helix. The electronic efficiency is furthermore highly dependent upon the manner in which the helix is supported. Ideally, the helix should be suspended in free space to avoid the shunting effect of the helix support structure on the RF wave. In practice, however, the helix is supported by thin ceramic rods that minimize the amount of dielectric material in contact or close to the helix. The electronic efficiency is also lowered by the RF losses of the helix circuit and the output coupler. It is also important to have a gain of 30 db or more in the output section of the tube to achieve maximum efficiency.

The largest factor in the denominator of the expression for efficiency is the power dissipated at the collector, P_c. This power can be minimized by operating the collector at the lowest possible voltage. For a good design this is approximately half the helix voltage. For collector voltages lower than this value the slow electrons in the beam (those that have given up the greatest energy) no longer enter the collector but are returned to the helix and result in excessive heating. They also reduce the overall tube efficiency. Care must therefore be taken in the design of the collector shape and the magnetic field configuration in this region to prevent the return of secondary electrons.

The heater power P_H is an important factor in the overall efficiency of a medium-power tube because it is typically 5 to 10% of the total power input. The exact heater power required for a traveling-wave tube is determined by the operating temperature, size, and means of support of the cathode. Because the cathode temperature is fairly well fixed by the type of emissive coating, this parameter is usually not a variable. The designer therefore selects a cathode size that operates at a safe current-density level for the tube life needed. Efficient designs of indirectly heated cathodes for traveling-wave tubes require a heater power of less than 9 watts/cm^2 of emitting surface. In the A-1292, for instance, heater power is 2.5 watts or 8 watts/cm^2 of cathode emitting area.

The power intercepted by the helix and gun elements, P_I, is usually less than 7% of the input power and is mostly dissipated on the helix. The primary reasons for interception are imperfections in the magnetic focusing field and tube geometry. Care must also be taken to design the collector geometry and magnetic field configuration so that secondary electrons are trapped and prevented from returning to the helix.

Size and Weight Design Considerations

Once the design of a traveling-wave tube that provides the high efficiency required has been determined, methods for obtaining the minimum size and weight of the tube must be considered. A major contributing factor in the weight of a traveling-wave tube is the focusing structure, which in tubes for satellite applications is usually a lightweight, periodic-permanent-magnet type. The size and weight of such structures are primarily determined by the way in which the tube is constructed and by the type of material used for the magnets.

For a given type of magnet material, the more closely the inner diameter of the magnet pole piece can be made to approach the diameter of the delay line (helix), the lighter the weight of the focusing structure will be. The importance of a small pole-piece diameter is indicated in Fig. 2, which shows that the weight of the tube rises from 0.4 to 3.2 pounds as the inner diameter of the magnet shim increases from 0.24 to 0.42 inches. These results were calculated for a periodic-permanent-magnet type of Alnico VIII focusing stack, 10 inches long, having a flux density of 700 gauss on the axis. This increase in weight corresponds to an increase in the outer diameter of the magnet from 0.4 to 1.29 inches. If platinum cobalt is used as the magnet material, instead of the Alnico VIII, the weight is reduced approximately by a factor of 2; however, the cost of the magnet stack is increased approximately 30 times.

Design and Performance of the A-1292

The prime objectives in the design of the A-1292 traveling-wave tube were high overall efficiency, light weight, and extreme ruggedness. A metal-ceramic construction and the design considerations discussed in the preceding paragraphs were employed to satisfy these somewhat conflicting requirements. Fig. 3 shows a cross-sectional view of the A-1292 tube.

The helix support structure in the A-1292 tube is designed to provide a maximum conduction of heat away from the tungsten helix, and at the same time, to insure a minimum shunting effect on the RF interaction field. Three beryllia support rods spaced at 120° intervals about the helix circumference are used to secure the helix in the metal envelope. The helix is copper-plated to reduce RF losses. The helix lead wires are brought out through ceramic insulators to provide low-loss coupling to the tube.

The elements of the electron gun (cathode structure, beam former, and anode) are brazed to metalized ceramic spacers to form a very rugged subassembly. The gun subassembly is inserted and brazed into the metallic cylinder that is used to form the vacuum envelope. Stem-lead connections through ceramic feed-through seals are used to apply the voltages to the gun electrodes. The length of the metal-ceramic seal in this gun is more than an order of magnitude shorter than that of the seal in the well-known "stacked gun." The reliability and ruggedness of the A-1292 gun are therefore substantially greater than those of the stacked gun.

The cathode in the A-1292 gun is supported on a tripod structure that assures both rigidity and a low loss of heat by conduction. This cathode operates at a loading density of 100 mamp/cm^2 requires 2.5 watts of heater power. Ion bombardment of the cathode is held to a minimum by operating the anode at a potential higher than that of the helix. This potential profile prevents ions formed in the helix region from striking the cathode.

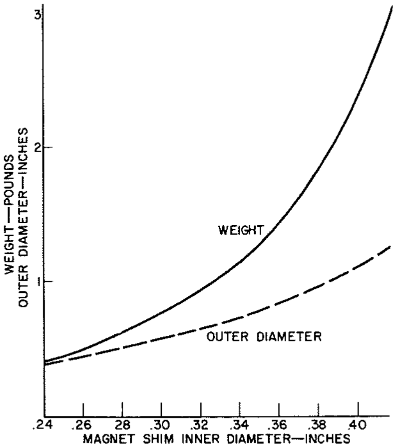


Fig. 2—Weight and outer diameter of a periodic-permanent-magnet focusing stack vs. inner diameter of the magnet shims.

The collector for the A-1292 tube was designed to dissipate the beam energy and trap secondary electrons. Its diameter was made as large as possible in keeping with the requirement that it still be small enough to permit the periodic ring magnets and shims to be slid onto the tube after exhaust. If a larger diameter had been chosen, it would have been necessary to split the ring magnets because no magnetic material can retain its magnetization during the exhaust bakeout at 600°C. Splitting of the magnets is undesirable because it increases magnet cost and creates problems in control and reliability.

One of the main advantages of metal-ceramic traveling-wave tubes discussed in the preceding section is that of mounting the focusing structure directly on the vacuum envelope. This direct-mounting technique has made possible the use of conventional Alnico magnets in the A-1292 tube, instead of the expensive platinum-cobalt alloy used in the RELAY tube, without any undue increase of the stack weight. In the final package, magnet structure, tube, and base plate are bonded into a rigid unit which can stand very high vibration and temperature levels without ill effects.

TABLE I—Comparison of A-1245 and A-1292

	A-1245	*A-1292*
Length, inches	17	13
Weight, lbs.	3.6	1.7
P_{out}, watts	11	13
Minimum Overall eff., %	21	31
Frequency, Mc	4,050-4,250	2,200-2,300
Heater Power, watts	5	2.5
Gain, db	35	35

A comparison between the A-1245 (RELAY tube) and the new A-1292 (Fig. 3) illustrates the progress achieved in the satellite-tube technology in only a few years. In spite of its operation at a much lower frequency (2.2 compared to 4.2 Gc), the A-1292 tube is much smaller and weighs only half as much as the A-1245 tube. The power output was increased slightly from 11 to 13 watts, but the required DC power input simultaneously was decreased from 52 to 42 watts. The major characteristics of the two tubes are compared in Table I.

CONCLUSIONS

Impressive as the advances have been, even more stringent requirements are programmed for future satellites, and RCA is continuing to investigate techniques to further improve its traveling-wave tubes for these applications.

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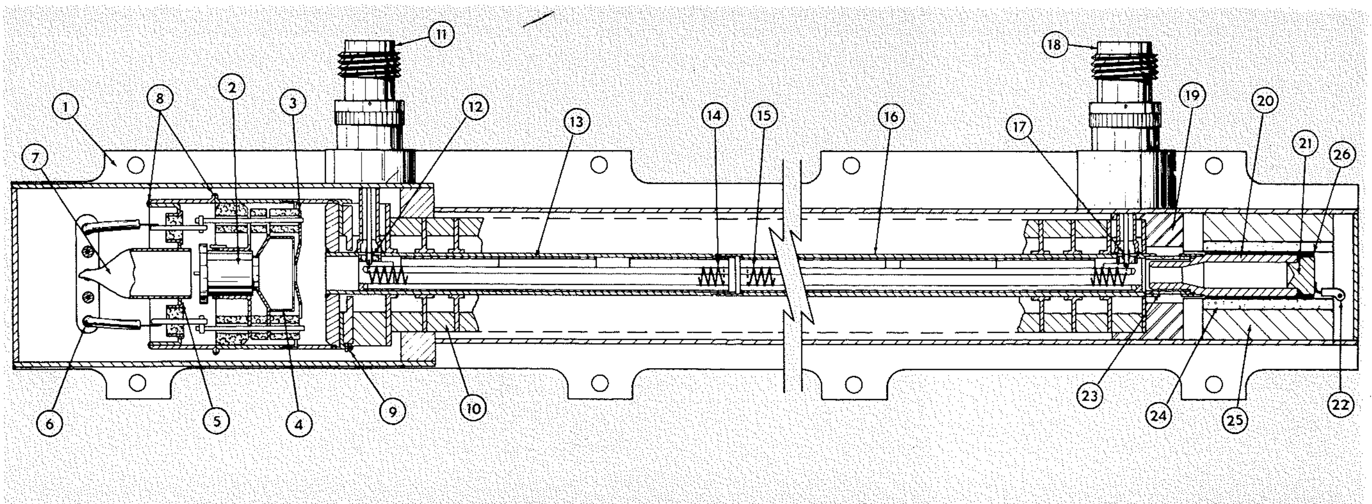
Fig. 3—RCA A-1292 metal-ceramic satellite communications traveling-wave tube; 2.2-3.0 Gc, 13 watts, 30% overall efficiency.

1. base plate
2. cathode assembly
3. anode
4. beam former assembly
5. stem
6. flying leads
7. tipoff
8. seal-in welds
9. focusing screw
10. periodic permanent magnet
11. input coupler
12. input-pin seal
13. compression ring
14. input helix
15. output helix
16. helix envelope
17. output pin seal
18. output connector
19. magnet stack lock
20. collector
21. collector plug
22. collector lead
23. collector isolator ceramic
24. beryllia insulator
25. heat sink
26. solder fillet



TEST AND SPECIFICATION ENGINEERING FOR MICROWAVE DEVICES

"Thorough product evaluation is the key to the successful exploitation of a creative design." This statement reflects the basic operating principle of the Test and Specification Engineering group of the EC&D Microwave Tube Operations Department. This paper describes how that group plans the test programs, performs the special tests, and provides and maintains the test facilities required for the complete evaluation of microwave components from their initial development through final production.

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The members of the Test and Specification Engineering group are essentially independent arbiters of product capability. In this role, they work very closely with the design-engineering, manufacturing, and quality-assurance groups—a close rapport which requires that test engineers and technicians be assigned on the basis of their special training and skills to meet the specific needs of each group. In development programs of sufficient size, a test engineer and technician are assigned full time to service the development of the product from inception to final production.

Responsibilities of the Test and Specification Engineering group include: the design and implementation of test-engineering programs for all products in development; operation and maintenance of the life-test-and-rating laboratory, the environmental-test laboratory, and the calibration laboratory; and the establishment and maintenance of instrument and specification centers. From these activities, the group provides six services:

1) Establishes accurate test specifications and procedures that fully define each product type;
2) Determines and obtains the test facilities necessary for a complete evaluation of capabilities of each product;
3) Assures that all test equipment is accurately calibrated and operates satisfactorily within tolerance limits;
4) Performs qualification-approval and engineering-design evaluations of products under various ambient-temperature, environmental, and life-test conditions;
5) Provides accurate estimates of the cost of the test program required in developments of new products;
6) Maintains control of those characteristics of a product that cannot be evaluated economically during manufacturing.

TEST ENGINEERING

Engineer-technician teams, who are specialists in the testing of the particular type of product being developed, are assigned to each development program to provide the following services:

1) Prepare the test program.
2) Prepare test specifications and procedures.
3) Develop the necessary test instrumentation.
4) Evaluate the components which are to be used in the product.
5) Conduct tests on preliminary models.
6) Perform acceptance tests on samples delivered to the customer.
7) Plan and perform the necessary qualification-approval tests.
8) Report the results of the qualification-approval tests.
9) Perform sample tests for a limited time on production models.
10) Assist manufacturing personnel, as needed, in their routine tests.

LIFE-TEST-AND-RATING LABORATORY

The life-test-and-rating functions of the Test and Specification Engineering group are provided mainly to support the manufacturing operations; however, life and rating tests are also performed on special request for engineering and quality-assurance groups. In general, the life-test-and-rating group tests samples from a typical production lot and makes wide distribution of all data accumulated. Such data is particularly useful for a statistical analysis of production. Most of the data is now fed into a computer system[1] which provides an immediate evaluation recorded on tape to assure easy access to averages or distributions of characteristics.

The rating equipment in the Life Test and Rating laboratory is very modern and similar to that used in manufacturing except that it provides much more accurate and comprehensive data than the *go* or *no-go* indication common in production. Although the laboratory has small vibrators, many design tests re-

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Final manuscript received October 22, 1964

Fig. 1—The Rating Laboratory.

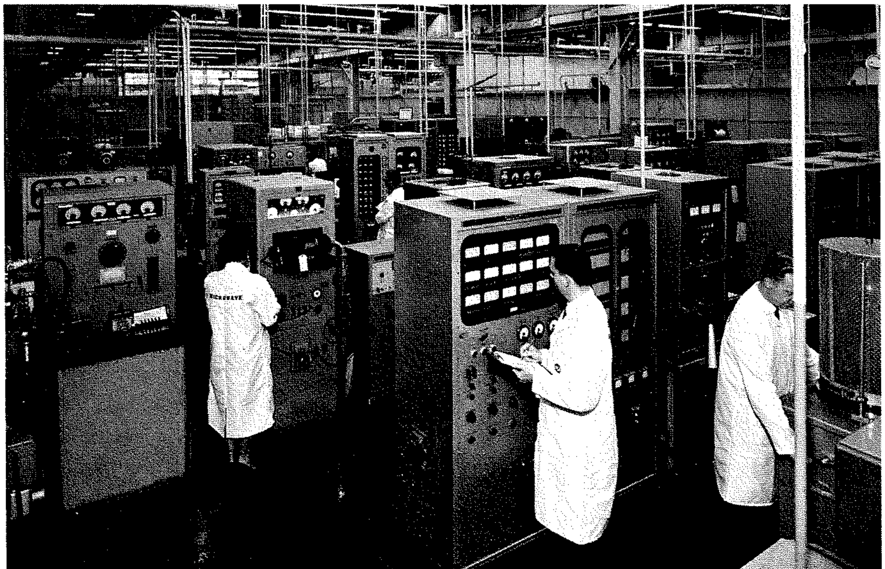
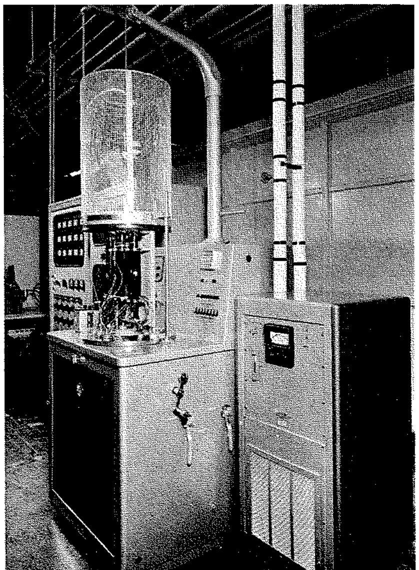


Fig. 2—Thermal-vacuum life-test chamber for RCA satellite TWT's.



quire the additional equipment located in the environmental laboratory.

Fig. 1 shows some of the variety of equipment required to perform the rating and life tests on microwave components. The life-test facilities consist of 30 items of equipment which can handle 8 magnetrons, 50 traveling-wave tubes, 500 pencil tubes, and 19 solid-state devices, under various operating conditions. The magnetron facilities include two C-band positions and seven X-band positions. Three of the X-band positions can be used for life tests of hydraulically tuned magnetrons. The 50 traveling-wave-tube positions include 8 high-vacuum positions (for life test of satellite tubes), 20 positions with automatic temperature cycling, and 22 positions in which tubes are tested at ambient room temperatures. The pencil-tube positions mainly provide oscillator life tests at frequencies from 500 to 4,000 Mc; 200 high-temperature positions are available for life test of the special, high-temperature ceramic pencil tubes.

Fig. 2 shows a typical life-test rack used for testing satellite traveling-wave tubes under high-vacuum conditions with a temperature-cycled heat sink. At present, 10 satellite tubes are undergoing life tests, and have already accumulated more than 160,000 life hours. Equipment is currently being constructed that will provide, in addition to the high-vacuum and temperature-cycled heat sink, a temperature-controlled shroud around the vacuum chamber to simulate the temperature conditions under which the tubes must radiate their heat while in a space environment.

For the solid-state devices, 7 ambient room-temperature life-test positions and 12 high-vacuum life-test positions are available.

ENVIRONMENTAL LABORATORY

The environmental laboratory has under its control environmental test equipment to assess the capabilities of a microwave device under variations of temperature, altitude, vibration, shock, and humidity. The laboratory staff not only operates this equipment but also must continuously compare the quality of their equipment with the current state-of-the-art in testing. If presently installed equipment falls behind, they recommend ways to rectify the situation.

TABLE I—Environmental Test Equipment

Type	*Limits*
MB C25H Vibrator	*Force:* 3500 lbs. or ½″ DA *Freq.:* 5–2000 cps
MB C6E Vibrator	*Force:* 200 lbs. or ½″ DA *Freq.:* 5–5000 cps
All American Vibrator	*Freq.:* 1–60 cps *Disp.:* 0.125 in.
Hyge HY3401 Shock Machine	*Force:* 15–50 g *Shape:* half sine *Time:* 11 msec
Barry 15575 Shock Machine	*Force:* 200 G max. *Shape:* half sine *Time:* 11 msec
IRC Temp-Alt-Hum Chamber	*Temp.:* —100 to +500° F *Alt.:* 0 to 100,000 ft *Hum.:* 20 to 95% over range 35 to 185°F dry bulb
Tenney Temp-Alt Chamber	*Temp.:* —100 to +360°F *Alt.:* 0 to 100,000 ft
ALT A-4-L Alt Chamber	*Alt.:* 0 to 100,000 ft
Mantec D102 Temp Chamber	*Temp.:* —75°C (with CO_2) to +325°C
Statham Temp Chamber	*Temp.:* —75°C (with CO_2) to +325°C
ATL CO_2 Temp Chamber	*Temp.:* —100°F to +300°F
ATL Temp-Hum Chamber	*Hum.:* 20% to 95% over range 0 to 300°F dry bulb

The laboratory staff works very closely with both the test-engineering and the rating-laboratory groups, who are the greatest users of the equipment. When special tests involve the environmental equipment, the test is planned jointly with the environmental laboratory. When a special environmental test exceeds equipment capabilities, this section arranges for an outside facility. Table I is a list of the environmental equipment located in the laboratory.

INSTRUMENT CENTER

An instrument center is maintained by The Test and Specification Engineering group. This center stores, catalogs and issues all engineering test equipment.

The catalog lists not only the equipment, but also all the pertinent engineering facts which would be needed in determining the utility of a particular test instrument. Equipment returned to the center is immediately checked so that only properly operating instruments are placed on the shelves. This central "clearing house" for all engineering test equipment provides an easy reference as to what equipment is available or when it will be available.

CALIBRATION LABORATORY

The calibration laboratory maintains the secondary standards for the Microwave Tube Operations Department. It is primarily a radio-frequency standards laboratory in that it provides only frequency, RF-power, noise-figure, attenuation, and VSWR calibrations. It does not provide DC or AC voltage or current calibrations, which are provided by another standards laboratory within Electronic Components and Devices.

Some of the standards with which the calibration laboratory is equipped are as follows:

1) "hot and cold" noise sources for calibrating noise figures,
2) calorimeters for power calibrations,
3) precise frequency-measuring counters,
4) precision capacitors, and
5) precision attenuators.

By use of these and other standards, the laboratory can calibrate:

1) frequency from 100 Mc to 12.4 Gc,
2) power up to 10 watts from 0 to 12.4 Gc,
3) attenuation up to 55 db from 2 to 12.4 Gc,
4) noise figure from 0 to 4 Gc, and
5) VSWR from 500 Mc to 12.4 Gc.

SPECIFICATION CENTER

The specification center is responsible for scheduling and managing all efforts concerned with the generating and updating of specifications covering all finished products and all purchased items other than raw materials and machined parts. The center also maintains a central data file for all finished-product specifications, both RCA and customer, and for related government specifications. When required, the specification section represents the Microwave Tube Operations Department in discussions and negotiations with customers, vendors, government agencies, and trade associations in areas concerning specifications.

While the specification center is the newest part of the Test and Specification Engineering group, it is expected that it will soon become one of its most important subsections.

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M. E. DeVITO received the BEE in 1953 and the MEE in 1960 both from the City College of New York. He joined RCA as a specialized engineering trainee in 1953; before completing this training, he was called into the Army, serving until 1955. He returned to RCA in 1955, and from 1955 to 1958, designed cavities for ceramic pencil tubes and developed techniques for measuring characteristics of ceramic pencil tubes. He was also an applications and customer-liaison engineer. In 1958, he joined the Microwave Life Test and Rating Laboratory, and subsequently was made project engineer in charge of the life-testing and rating of traveling-wave tubes, magnetrons, and pencil tubes. In 1962, he was promoted to his current position, Engineering Leader, Microwave Test and Specification Engineering. In this position, he has supervisory responsibility for all design and life tests, environmental evaluations, and qualification-approval programs on traveling-wave tubes, magnetrons, and pencil tubes and solid-state devices; for solving calibrating and setting up calibration standards for all equipment used in the design and manufacture of RCA microwave devices. Mr. DeVito is a Member of the IEEE.

AUTOMATIC TECHNIQUES USED IN THE DEVELOPMENT AND MANUFACTURE OF TRAVELING-WAVE TUBES

The helices used in traveling-wave tubes must be selected on the basis of very precise masurements, because even minute deviations from the design objectives in over-all helix characteristics or in turn-to-turn uniformity severely affect tube performance. The resistance of a traveling-wave tube to shock and vibration is a direct function of how firmly the helix is supported in the vacuum envelope. The support arrangement, however, must not disturb the precise alignment of, create an unbalance in the loading on, or distort the helix, because such conditions would result in a significant change in helix electrical characteristics. The stability of electrical characteristics and the life of traveling-wave tubes are critically dependent upon the processing methods and schedules employed during tube exhaust. Automatic techniques, which employ improved processing methods, have been developed to supplant or supplement the highly technical manual skills required for these critical operations. These techniques can be used to effect substantial increases in product yield together with appreciable reductions in the costs incurred in the development and manufacture of traveling-wave tubes.

F. ULRICH, M. FROMER, K. KAROL,* D. MAWHINNEY, and E. THALL

Microwave Tube Operations Department
Electronic Components and Devices
Harrison, N. J.

In the complex tasks related to the development and subsequent manufacture of traveling-wave tubes, the design-engineering effort is immeasurably aided by the supporting services provided by other engineering groups. For example, tool engineers design the tools and assembly jigs required in tube construction, both for the early stages of development and for the subsequent higher-volume production in the manufacturing operation. Methods and process engineers specify processing methods and facilities and develop new assembly techniques. Test and specifications engineers are responsible for all qualification-approval, design-verification, and life tests and for the development of test equipment. These support engineering groups work very closely with tube-design engineers from the very beginning of and throughout the tube-development program.

The development of new techniques and equipment for use in the assembly and processing of the tubes is often necessary for new designs to become practically realizable. Such developments are frequently the basis for state-of-the-art improvements in the traveling-wave-tube industry. When the development of a new product requires new processes or equipment, certain logical steps must be followed in the development and evaluation of the new process or equipment from conception to final release, and the interaction of many groups of engineering personnel is required. The PERT chart in Fig. 1 shows the project plan for such operations in the Microwave Tube Operations Department (MTOD).

This paper discusses new processes and automatic equipments developed by MTOD methods, process, and equipment engineers for use in the development and manufacture of traveling-wave tubes that have contributed substantially to the solutions to the problems in three critical areas: measurement of helix parameters, assembly and support of the helix in the vacuum envelope, and exhaust processing of the tube. An equipment that provides high-precision measurements of traveling-wave-tube helices, the development of a new process for securing helices to glass envelopes, an automatic system for securing the helices to the glass envelopes by this process, and an automatic adaptively controlled exhaust-processing machine are described.

HELIX MEASUREMENTS

One of the major components of a traveling-wave tube is the helix. A typical helix is made of 0.005-inch-diameter tungsten wire, has 90 turns per inch, and is 8 inches long and 0.090 inch in diameter. Evaluations have shown that deviations in the helix pitch of only a

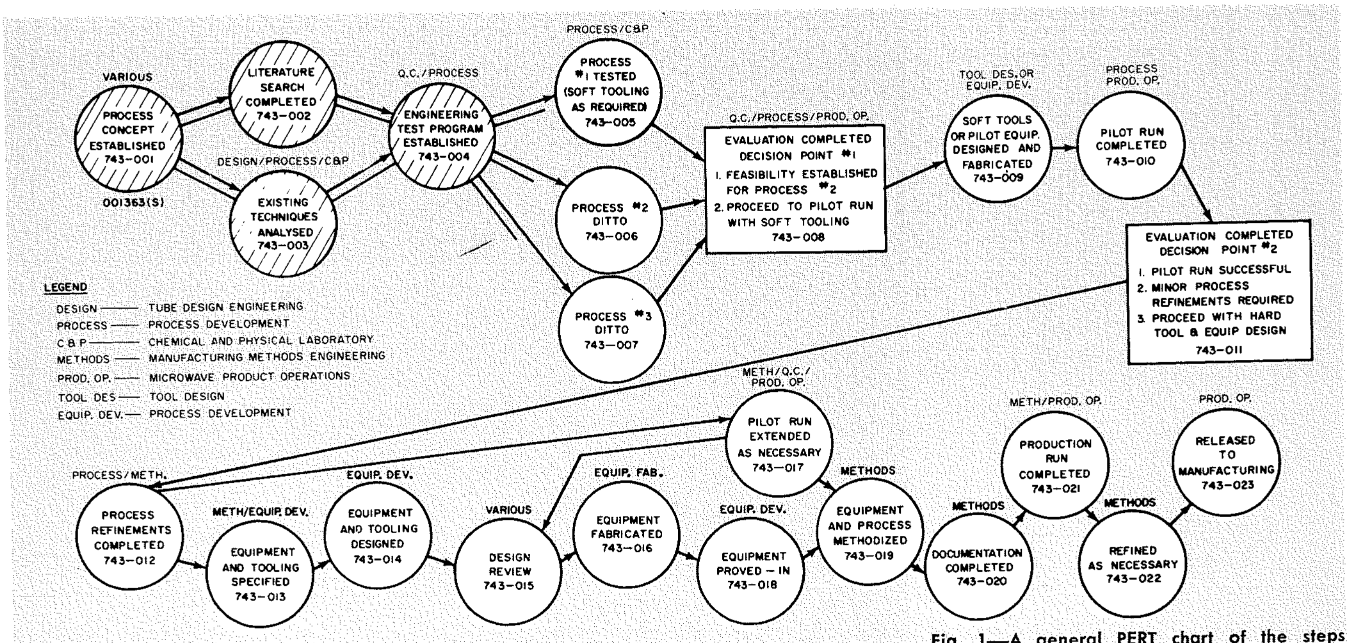


Fig. 1—A general PERT chart of the steps MTOD uses in evolving a manufacturing process from its inception to its final release.

Final manuscript received October 22, 1964

* Mr. Karol is no longer with RCA, and his biography and photo were not available for publication.

few microns can cause severe variations in the gain of traveling-wave tubes. Consequently, there is an urgent need for precision measurement and control of this parameter during helix fabrication. Instruments such as toolmaker's microscopes and optical comparators can be used, but are impractical for this application because of the strain they impose on the operator. For example, 12 to 16 hours may be required to obtain a complete set of data on a single helix by these methods, and because of operator fatigue, data may not be dependable.

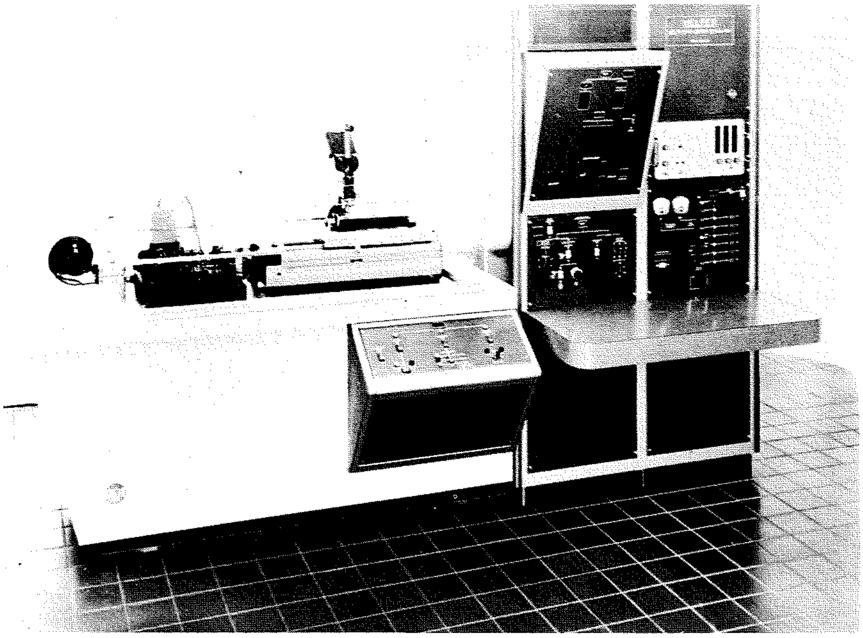


Fig. 2—The *HELix Pitch Error Resolver* (HELPER). An MTOD-developed instrument for measuring pitch conformities in traveling-wave-tube helices.

At MTOD, some of the problems inherent in the control of helix fabrication have been solved by the development of a new, automatic helix-pitch measuring instrument. This instrument, called HELPER (*HEL*ix *P*itch *E*rror *R*esolver) is shown in Fig. 2. HELPER is used to measure traveling-wave tube helices for pitch conformity on either a turn-to-turn or a cumulative basis. It is a hybrid analog-digitally controlled instrument which uses adaptive programming techniques and which provides an overall system precision of better than 2 microns.

To minimize measuring errors and machine maintenance, the instrument is located in a special controlled-environment room in which temperature and humidity are precisely regulated and which is free from lint, dust, or fumes. In addition, the actual mechanisms are housed in clear plastic enclosures which are kept under positive pressure with filtered air. All major heat-dissipating components are located in the control cabinets with the exception of the instrument motors located on the gear train. These motors, which are used intermittently, are in the stream of the filtered air to minimize effects of their heat.

HELPER was designed as a comparative (as opposed to an absolute) measuring apparatus because of practical needs and economics. Its absolute accuracy is in the neighborhood of ±10 microns, while its precision is in the order of ±2 microns. The calibration of the instrument for absolute accuracy would require a calibrated grating freshly verified from the Bureau of Standards or equivalent. A true-accuracy test would be expensive and time-consuming, and is unwarranted for the practical use of HELPER.

The basic system operation is best explained by an example. Assume that the first turn of a helix is positioned to an arbitrary index using a transducer for detection and that the theoretical pitch is known. Then a hybrid two-step action takes place: First, the helix is moved precisely one increment of theoretical pitch under the transducer (a digital operation). Second, the arbitrary index is sought, and the helix moves to this position (analog motion). The amount of deflection made in the second step is directly proportional to the pitch and is scaled to a display-counter unit which is read by an operator.

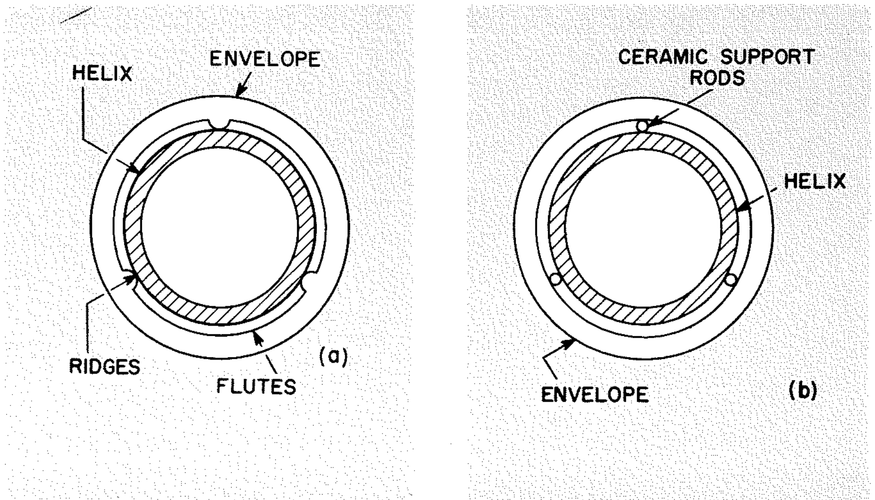
Because the over-all objective was to improve accuracy of helices used in traveling-wave tubes, the HELPER data were first used to improve existing helix-winding equipment. Specifically, the analysis of the data from this measuring instrument was responsible for machine changes which improved the helix-winding machine capability from approximately ±10-micron deviation in pitch to ±6-microns on certain helices. Helices for traveling-wave tubes used in the RELAY satellite were measured on HELPER, and chosen on the basis of the data generated.

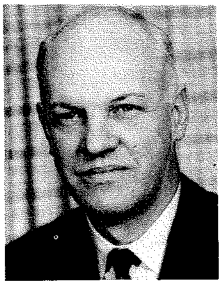
THE HELIX-SECURING PROCESS

After helices have been wound and annealed and measurements have proved them to be satisfactory, they are ready to be inserted into and secured to the traveling-wave-tube envelope. A successful securing operation must insure that the helix will be very precisely aligned within the vacuum envelope, that the dielectric loading will be balanced, and that the finished assembly will be sufficiently rugged to withstand the rigors of shock and vibration in the given application. The two basic methods of supporting a helix in a glass envelope are shown by the cross-sectional drawings in Fig. 3. Sketch *a* shows a helix supported by the ridges of a fluted-glass type of envelope; sketch *b* shows a helix supported by three ceramic rods.

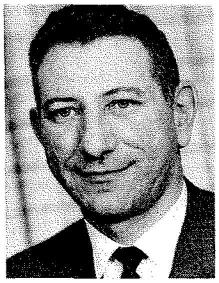
Until recently, the main process used at MTOD to affix helices to traveling-wave-tube envelopes was furnace-embedding. In this process, properly aligned helix and fluted-glass envelope assemblies, mounted in a special compression holding fixture, are heated inside a furnace, in a hydrogen atmosphere, just enough to cause a slight softening of the glass. The compression force then causes the helix wire to be embedded in the ridges of the fluted-glass envelope. With this process, however, it was extremely difficult to achieve uniform embedding in all ridges of the glass over the entire length of the helix.

Fig. 3—Cross section of a traveling-wave-tube helix within a vacuum envelope. (a) Helix supported by the ridges in an internally fluted envelope. (b) Helix supported by three ceramic rods.





F. Ulrich



M. Fromer

D. Mawhinney

E. Thall

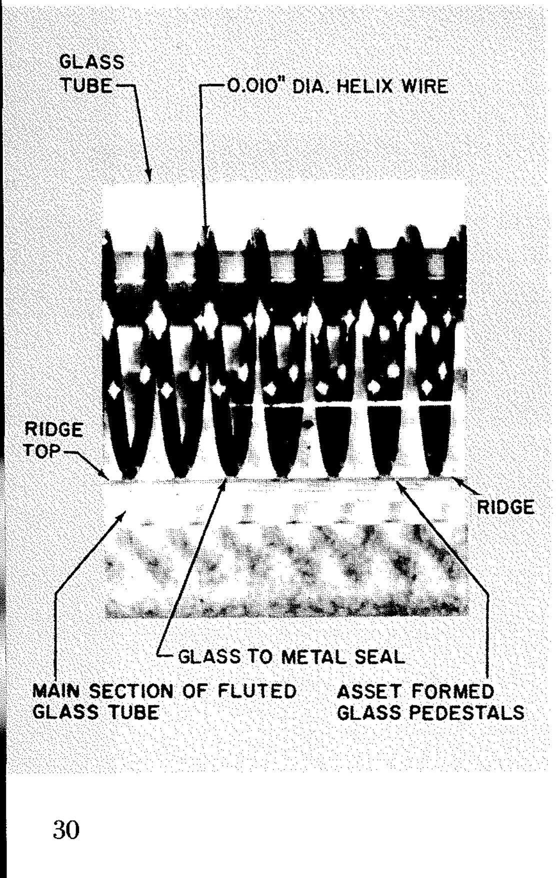
F. ULRICH received his BSME from the Newark College of Engineering, in 1942, and his MSIE from Stevens Institute of Technology in 1953. From 1942 to 1944, he worked for the Conmar Products Corporation. From 1944 to 1946, he was Assistant Production Officer at the Naval Ordnance Plant in Louisville, Ky. From 1946 to 1947, he worked as a planning engineer for the Western Electric Corporation. From 1947 to 1951, he supervised projects in tool procurement and design and manufacturing planning for Bendix Aviation Corporation and General Electric. From 1951 to 1956, he worked as a production project engineer at the Elastic Stop Nut Corporation. Mr. Ulrich joined the RCA Microwave Tube Operations Department in 1956 and was placed in charge of the Tool Engineering activity of the Microwave Equipment Development group. In 1959, he was promoted to Engineering Leader responsible for development of special fabrication facilities for production of microwave tubes. In 1961, he was promoted to Manager, Manufacturing Methods Development, and was responsible for the development of assembly, process, and test methods, and product improvement to establish manufacturability of new tube types. Mr. Ulrich was promoted to Manager, Manufacturing Support Engineering in 1962. In this position, he directed the work of the Tool Engineering activity, the Life Test and Rating Laboratory, and the Manufacturing Methods group. Since April 1964, he has been Manager of Process Development directing the activities of the Process Development, Chemical and Physical, and the Tool and Mechanical Equipment Design groups. Mr. Ulrich is a Registered Professional Engineer in the State of New Jersey and a Senior Member of ASME and ASTME.

M. FROMER received his BSEE from the Newark College of Engineering in 1950. From 1950 to 1953, he worked for the Weston Electric Instrument Corporation on the design of industrial instruments. From 1953 to 1955, he worked for the Winslow Company. He was an Engineering Leader on advanced development of servo equipment and computer machinery, at both the Helipot Corporation (1955 to 1957) and the Kearfott Company (1957 to 1959). Mr. Fromer joined the RCA Microwave Tube Operations Department, in 1959, as an equipment development engineer. The helix pitch error resolver (HELPER), which he designed and put into operation, has contributed substantially to the traveling-wave-tube art. He also made systems engineering studies on equipment requirements and techniques used in manufacturing RCA microwave tubes. Mr. Fromer was appointed Engineering Leader, Electrical Equipment Development, in 1961. In 1962, he was a system engineer in the Data Systems Development group, and in 1964, became Engineering Leader of the Solid-State Engineering Mechanical Design group. This group develops and refines electro-mechanical aspects of solid-state microwave devices. Mr. Fromer holds a patent for his work on ultra-precision servo systems. He is a member of the IEEE and a Senior Member of the Instrument Society of America. He received a 1963 Electronic Components and Devices "Engineering Achievement Award" for "ingenuity in the development of a practical tuning system for hydraulically tuned magnetrons."

D. MAWHINNEY received his BSEE from Polytechnic Institute of Brooklyn in 1957. He was in military service from 1948 to 1952, where he attended the Signal Corps Radar School and was later assigned to the Naval Research Laboratory to work on developmental programs for the AEC. In 1952, Mr. Mawhinney joined the RCA Microwave Tube Operations Department. He has contributed to the development of numerous special microwave devices used in the test and evaluation of microwave tubes. In 1959, he became Engineering Leader of Microwave Equipment Design with the responsibility for the development of all electrical test equipment. He is currently Engineering Leader, Solid-State Systems Development and is assigned to the Lunar Excursion Module team engaged in the design and fabrication of solid-state frequency-multiplier units. Mr. Mawhinney has been issued a patent for his work and has published several papers. He is member of the IEEE and a Licensed Professional Engineer in the State of New Jersey.

E. S. THALL received his BA in Science from the School of Engineering of the University of Toronto in 1945. He then joined the technical staff of the Ontario Research Foundation. In 1950, he returned to the School of Engineering where he served as special lecturer in physical metallurgy until 1952, while earning his MA in Science. From 1952 to 1953, he was assistant plant metallurgist for the John Inglis Company, Toronto. In 1953, he joined the RCA Electron Tube Division as an engineer in the Methods and Process Laboratory, working on processing and materials development and general metallurgical problems. In July 1959, he was assigned to the Chemical and Physical Laboratory of the RCA Microwave Tube Operations Department. From that time until June 1960, he conducted vapor-pressure and cathode-poisoning studies. In his recent work, he has developed a novel helix-flashing technique for embedding the helix in glass traveling-wave tubes, a method of making pyrolytic carbon attenuator coatings from plastic, and a method of adapting the technique of film cathodes to traveling-wave tubes. He has applications for patents on this work. Mr. Thall is a member of the American Society for Metals and is registered with the Association of Professional Engineers, Province of Ontario.

Fig. 4—The result of the MTOD-developed ASSET process of sealing helix to glass at points of contact on ridges of internally fluted glass tubing.



Work by K. N. Karol and E. S. Thall led to the development of a new method of securing the helix in the tube envelope. With this new method, the helix wire is sealed, at the points of contact, to the ridges of the internally fluted-glass envelope, or to ceramic rods in tubes that use the three-rod helix-support technique. The helix is operated as an electric heater in a forming-gas atmosphere to effect the glass- or ceramic-to-metal bonds at the localized areas in which ridges of fluted-glass or ceramic support rods contact the helix wire.

The idea of using the helix as a direct heater for shrinking the glass and securing the helix to the envelope was not new. However, in earlier attempts, it had always been assumed that the whole envelope should be heated slowly, so that the surface tension of the envelope would gradually pull the glass around the helix and, in this way secure it in place. This approach necessitated the use of an extended heating period, low voltages, and corresponding high currents. The result was always an uneven embedding of the helix and a grossly distorted envelope.

During the course of the process investigation by Karol and Thall, a 1-inch length of helix wire was placed in a corresponding length of glass tubing and an electrical voltage was impressed across the ends of the helix. It became obvious that if the time of heating were sufficiently reduced, distortion and melting could be restricted to those areas of the glass in contact with the helix wire. It was found that the heating period had to be limited to a fraction of a second and that relatively high voltages and currents were required.

A laboratory setup was devised to evaluate the process. This setup was initially designed to secure 10-mil-diameter tungsten wire, wound at about 50 turns per inch, in 6-inch lengths of glass tubing; however, its capabilities were later extended to handle assemblies 10 inches in length. About 150 volts per inch of helix length and a heating time that corresponded to two cycles of 60 cycle AC were required for the sealing operation. A synchronous welder was used for the timing device. This timing device was necessary not only for accurate timing, but also to avoid the huge surge currents that would result if the magnetic fields of the power transformer were to collapse in the middle of the half cycle.

Oscilloscope studies indicated that currents as high as 50 amperes were being drawn during the first half of the initial cycle, that input voltages of 1,500 volts would be required for a normal 10-inch traveling-wave tube, and that the wire temperature would rise to about 3,000°C during heating. Fortunately, the length of time of the flashing pulse is short so that considerable overload of the equipment could be tolerated.

The resulting "flash" technique for securing helices forms a strong helix-to-glass bond and has the added advantage of a very low dielectric loading on the helix because the glass is pulled into small support pedestals at the points of contact, as shown in Fig. 4. In order to keep glass strain at a minimum, a short preheat step before and an annealing step after flashing are necessary.

AUTOMATIC SECURING OF HELICES

The basic process-development engineering and resulting pilot equipment for flash-securing of helices to envelopes, described in the previous section, set the stage for the development of an automatic helix-securing system for use in the manufacture of traveling-wave tubes.

A machine called ASSET (*A*utomatic *S*ystem for *S*ealing *E*lements in *T*ubes) was recently developed at MTOD (Fig. 5). In the operation of the system, the program components are first determined and inserted, and the operator need only load the work and push a button to start the process. Once started, the process continues automatically to its predetermined conclusion. Protection circuits are incorporated in the machine which will halt the process if any malfunction exists in the machine itself or the elements acted upon by it.

There are five basic process requirements in the use of ASSET to seal helices in traveling-wave tubes: atmosphere control, preheat, seal (or flash), anneal, and soak anneal. ASSET can be programmed to accomplish any of these.

The first step, atmosphere control, is accomplished through the use of metering valves, flow switches, and time-delay relays which are permanently set to provide a purge cycle and steady flow of nitrogen. The air in the atmosphere chamber is thereby replaced with nitrogen, and the nitrogen atmosphere is maintained throughout the process.

The second and fourth steps (preheat and anneal) require the use of an individual power supply for each work station. (Three work stations are provided at the ASSET console.) This supply or constant-current regulator is cam-controlled and will supply a constant current although the load resistance changes with time as the helix heats or cools. The cam is positioned on the controller, and a cam follower actuates a potentiometer which results in an input signal to the constant-current regulator. Thus, the setting of the cam determines that the value of current that passes through the helix is that required to produce the desired temperatures for the preheat and anneal stages. Selection of various current and voltage ranges is made possible by the use of a plug program that is inserted in the front panel.

The actual sealing of the helix is the third step of the process. This step requires the use of a pulse power supply that is common to all work stations. This pulse power supply is used to create the high-energy flash current which determines the size of the seal area and shape of the pedestals formed during sealing. The pulse power supply is programmed through the insertion of a plug-in resistor in the work-station controller. The value of this resistor determines the peak voltage of the pulse. The timing of the pulse power supply is programmed by the insertion of a resistor at the timer cabinet as part of the plug-in program that is inserted in the front panel of the work station. The value of the resistor determines the time duration of the flash.

Common use of the pulse power supply by all ASSET work stations requires a logic system to prevent a simultaneous demand for the pulse by the three positions. This logic system insures that no more than one station at a time will insert into the power supply circuitry the signals for voltage setting, duration of flash, or flash initiation. No work station can start if it demands a flash-volt-

Fig. 5—The MTOD-developed Automatic System for Sealing Elements in Tubes (ASSET).

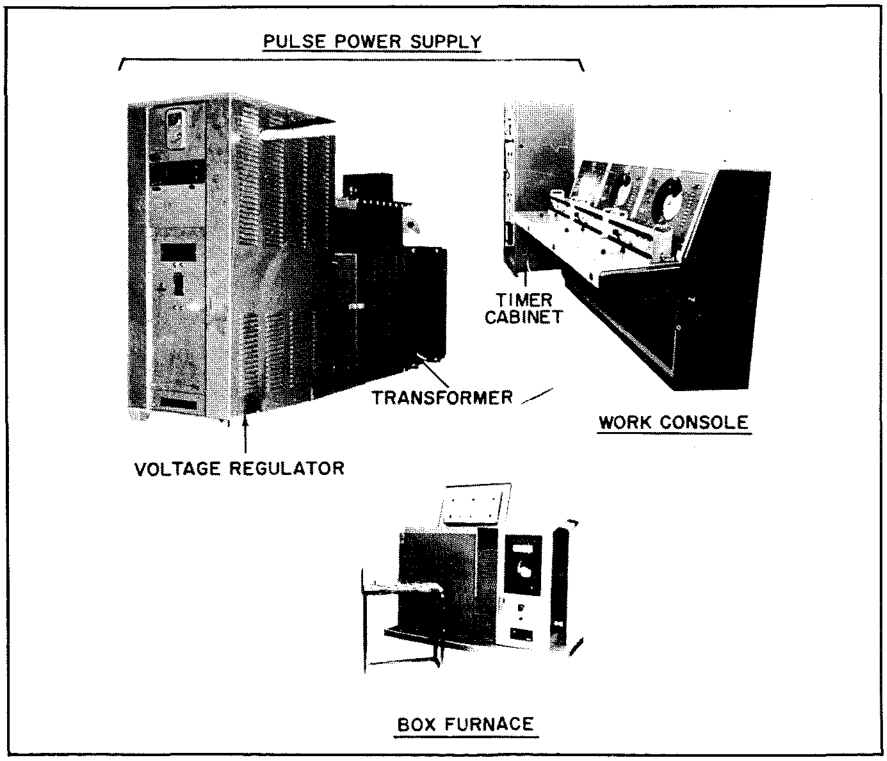
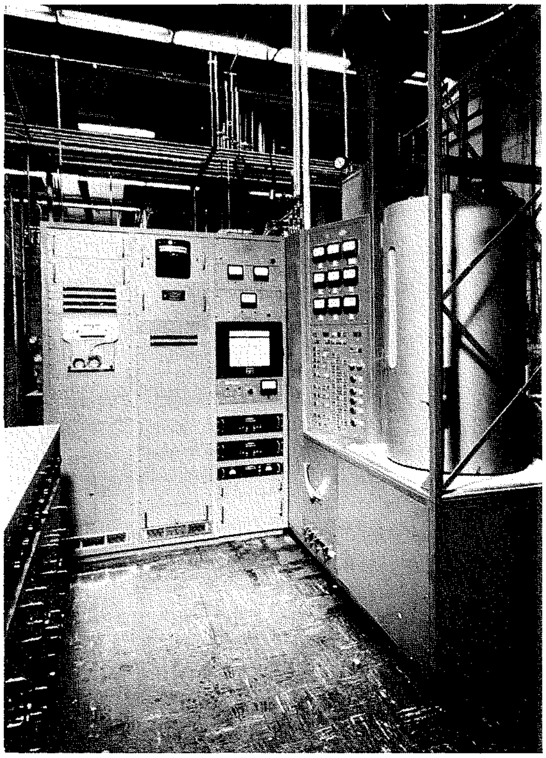


Fig. 6—MTOD-developed Automatic Exhaust Station for traveling-wave tubes.



age setting and subsequent flash within one minute of the time another station, already operational, is scheduled to make a demand of the pulse power supply. Thus, no two stations may demand a voltage setting and flash closer than one minute apart. This requirement is met automatically by the choice of a brush holder arm during the setup of a program. The choice of the brush holder arm also determines the time available for preheat prior to flash. There are eight choices that can be made with time increments of 1¼ minutes starting after a nitrogen purge of 2 minutes and 25 seconds and continuing up to a maximum preheat period of 12 minutes.

The final step in the process is the soak annealing to remove all stress from the helix-bulb assembly. This annealing is accomplished by placing several helix assemblies in low-cost jigs in a box furnace which permits loading of individual assemblies. The furnace is maintained at an elevated temperature until it is fully loaded and then is cycled through an appropriate annealing schedule. The furnace is provided with a controller which is programmed by a cam adjustment. An atmosphere of forming gas is used within the furnace chamber.

Thus, after the selection and insertion of the program components, the operator merely loads the helix-glass-bulb assembly in its position in the jig and presses a button. As the cycle proceeds, indicator lights signal the events of the program as they occur. When the programmed cycle is complete, the controller returns to its starting position and signals *process complete.* Should any malfunction occur, the controller will stop and signal *no go.*

AUTOMATIC EXHAUST PROCESSING

One of the most complex production steps in the manufacture of traveling-wave-tubes is the exhaust processing. The purpose of this processing is to evacuate the tube by simultaneously pumping a vacuum and heating internal surfaces by several different techniques to drive out trapped impurities. The subsequent operation of the cathode and, therefore, the quality and life of the tube is strongly dependent upon this process. The structure of the traveling-wave-tube is dictated by complicated electronic and microwave design criteria and is not compatible with an ideal exhaust system. Special processing requirements are also usually dictated to the extent that common exhaust schedules, or even procedures, are difficult to establish during the production of the small quantities usually required.

Because of the technical complexity of this process, the critical dependence of the product upon the proper techniques and methods, and the high cost of the process in skilled personnel and capital facilities, a program to develop an automatic exhaust system for traveling-wave tubes was initiated at MTOD. The resultant equipment is shown in Fig. 6. In the design of this equipment, modern programming and control techniques were combined with advanced tube-processing concepts to achieve a major improvement in traveling-wave-tube exhaust technology and capabilities.

Special Control Features

The three most important control features incorporated into the automatic exhaust-processing system are punched-tape programming, multiple-element adaptive control, and digital logic block circuitry.

Greater flexibility in processing schedules is achieved through the use of the punched-tape programming. A 24-line block program eliminates the need for a stored memory unit in the equipment. Each block contains all the information required for each step, which includes pressure, time, temperature, and voltage limits, and all the valve position and interlock conditions associated with the process-step address designation.

The main adaptive control of the process-forcing elements is provided by the feedback from the pressure and rate-of-change of pressure measurement and comparator circuits. The comparison of programmed and feedback signals from each of the forcing elements (i.e., voltage, RF, and oven) is made by a series of operational amplifiers which drive either saturable reactor or relay servo-control devices.

The internal logic system employs transistorized *nor* digital logic circuits. This system is excellent for the complicated cross-interlocking necessary for fully automatic operation. The design of this logic system seems to be more complicated than necessary; however, areas of possible simplification have not, as yet, been thoroughly explored.

Advanced Processing Techniques

Several advanced processing techniques are employed in the automatic exhaust-processing system. Among them are such features as:

1) Simultaneous RF induction heating of the electron gun, the collector, and the helix;
2) Maintenance of the entire tube at elevated temperatures in a bakeout oven throughout the exhaust process;
3) Use of pressure and pressure rates-of-change limits to force the exhaust process;
4) Use of programmed cathode breakdown and activation cycles.

The evidence obtained from exhaust programs on two RCA traveling-wave tubes, the types A-1219 and 4036, as well as the subjective evidence gleaned from witnessing the process, indicates that the simultaneous bakeout and processing (compared to existing sequential processing of glass traveling-wave tubes) result in an enormous increase in the rate of out-gassing the tube. The problem of chasing gas from hot spots to cooler areas is virtually eliminated by the new techniques. Similar advantages have been derived from the RF induction-heating system which is part of the station. The simultaneous heating of all elements, including the complete helix, has shown good results. The RF induction-heating system in the station works satisfactorily and, except for the physical awkwardness of the coil-mounting structures and their tendency to oxidize, is well designed for the power ratings used thus far. The feasibility of automatic breakdown and activation of traveling-wave-tube cathodes, including low-noise types, from a punched-tape program under the restraints of adaptive control has been demonstrated.

To date, four types of traveling-wave tubes have been programmed and processed in the automatic exhaust system. Studies of the process data which the machine has generated, together with the resulting changes made to the programming tape to adjust the process, have resulted in substantial exhaust-time reductions with no measurable degradation in tube performance or life. The ultra-low-noise (5.0-db-maximum) 4036 traveling-wave tubes, for example, has been completely processed in 2 hours, compared to the former processing time of 5 hours. The automatic processing system has demonstrated its usefulness in developing optimum exhaust schedules for existing production tube types as well as for new engineering types.

CONCLUSION

The brief descriptions of processes and equipments given in this paper indicate that the application of automatic techniques to the low-volume manufacture of highly engineered and customized products, such as traveling-wave tubes, differs from that of mass-production automation in purpose and approach. Rather than a system of high-speed operational and transfer devices which are primarily designed to replace manual skills and permit large-volume production, this form of automation must attempt to replace and enhance technical skills so that the product yield will be increased and production costs decreased. The emphasis is on such features as process control and data generation.

ELECTRON-BEAM—PLASMA AMPLIFIERS

A Future in Millimeter-Wave Amplification

Electron beam interaction with a plasma is a potential method for amplification of millimeter-wave and submillimeter-wave power. Frequencies as high as 38 Gc have been amplified with an electron-beam—plasma system. The measured gain per unit of length of such a system is high: 13 db/cm at K band. Plasmas which can be utilized in a 300-Gc beam-plasma amplifier have been produced. Efficient coupling of millimeter-wave power to a beam-plasma system remains to be accomplished. Two possible methods of coupling are described: one method uses a strong magnetic field, and the other utilizes a tubular plasma geometry.

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With the recent development of lasers, coherent electromagnetic radiation at higher power levels are available in the red, infrared, and microwave regions of the spectrum. The millimeter and submillimeter wave region of the electromagnetic spectrum remains relatively unconquered by existing technology. Devices for the generation of coherent electromagnetic radiation at frequencies up to 500 Gc are available[1]; however, the cost of such devices is high and the power available is low. Two limitations on existing millimeter wave generators are the extremely close mechanical tolerances that must be maintained in very small structures and the limited power that can be carried by such minute structures. Recent research in the field of plasmas indicates that the interaction of an electron beam with a plasma can be utilized for the amplification of millimeter wave power without the use of very small structures. Problems concerned with the production of the stable, highly ionized, high-density plasmas which are required in electron beam-plasma millimeter-wave amplifiers are solved.[2] The problem of efficiently coupling millimeter wave power to the electron beam-plasma system is not yet solved, but two methods of coupling are under investigation.

THEORY

Excitation of plasma oscillations by a high velocity electron beam traversing a plasma was predicted by Bohm and Gross in 1949.[3] The dispersion relationship for an infinite diameter electron beam in an infinite diameter plasma is:

$$0 = 1 - \frac{\omega_p^2}{\omega^2\left(1 - i\frac{\nu}{\omega}\right)^2} - \frac{\omega_b^2}{(\omega - kv_b)^2} \tag{1}$$

Where: a small signal, characterized by $\exp[i(\omega t - kz)]$ is impressed on the electron beam, ω_p is the plasma frequency of the plasma electrons, ω_b is the plasma frequency of the beam electrons, v_b is the electron beam velocity, and ν is the plasma electron collision frequency. The plasma oscillation frequency ω_p is related to the plasma density n_p by:

$$\omega_p^2 = \frac{n_p e^2}{\varepsilon_o m} \tag{2}$$

Where: e is the electron charge, m the mass of an electron and ε_o is the permitivity of free space. Since the solution of the dispersion equation shows that for $(\omega_p/\omega) \geqslant 1$ the propagation constant k is complex with conjugate roots, growing waves exist in the system. The growing waves are a result of the interaction between the space charge wave on the electron beam and the excited plasma oscillations. The space charge instability which results in power gain is convective, which means that the wave or perturbation on the beam increases with the distance that the beam travels through the plasma. The maximum gain per unit length occurs when $\omega_p/\omega = 1$. The gain mechanism in the system is partially analogous to that in a multicavity klystron. In contrast to a klystron or a traveling-wave tube in which the beam velocity is restricted to within certain values, the beam in a beam-plasma system can travel at any velocity that is greater than the thermal velocity of the plasma electrons. In addition, the high-frequency field in a beam-plasma device does not tend to diminish toward the center of the beam as occurs in a beam-helix or beam-cavity device. The penetration of the entire beam by the plasma-oscillation field results in a higher gain per spacecharge wavelength and permits a larger beam diameter, hence more power output. However, the beam-plasma amplifier has an important limitation which results from the beam electrons colliding with the ions and neutral atoms in the plasma. This mechanism dissipates power in the beam and creates noise in the output.

Beam dissipation by particle collision is the factor which places a limit on the useful plasma density and thus the frequency at which gain will occur. The criterion is that the gain per wavelength caused by the beam-plasma interaction must be greater than the loss per wavelength caused by electron collision; i.e., $Kl > 1$, where: K is the gain constant in nepers/cm and l is the mean free path (in centimeters) of the beam electrons. The gain constant K is fixed by such parameters as the electron-beam diameter, velocity and density, the plasma electron collision frequency, and the operating frequency. The length l is determined by such parameters as the electron-beam velocity, collision cross-section of the gas used for the plasma and total particle density in the plasma.

Consider a large-diameter, 2,000-volt, 1-amp/cm^2 electron beam passing through a cesium plasma. Since the ion density is fixed by the operating frequency, one can calculate the maximum density of neutral particles which can be allowed and still have net gain. Fig. 1 shows the result of such a calculation.[4] Plotted in Fig. 1 are the minimum percent ionization for 10-db gain and 0-db gain as a function of the operating frequencies. The percent ionization is the ratio of ion density to the sum of ion and neutral density. One can see that for the case shown in Fig. 1, the ionization must be greater than 30% for reasonable gain at 300 Gc. This minimum percent ionization may be reduced by increasing the beam velocity and density. However, inhomogeneities in the plasma and the use of smaller beam diameters to prevent high-order modes of the space-charge wave will tend to reduce the gain constant and increase the minimum percent ionization. Thus about 30% ionization will have to be achieved in a usable beam-plasma amplifier operating at 300 Gc.

For a complete picture of the power loss mechanism in the beam-plasma system, the collisions of the plasma electrons must be considered. Even though ν/ω is sufficiently small and the mean free path of the beam electrons is sufficiently large to permit a net gain, the power lost through plasma electron collisions can seriously reduce the efficiency of the amplifier. As the DC energy is transformed to AC energy, electron collisions with the ion and neutrals provide a mechanism for the transfer of coherent energy to incoherent or heat energy in

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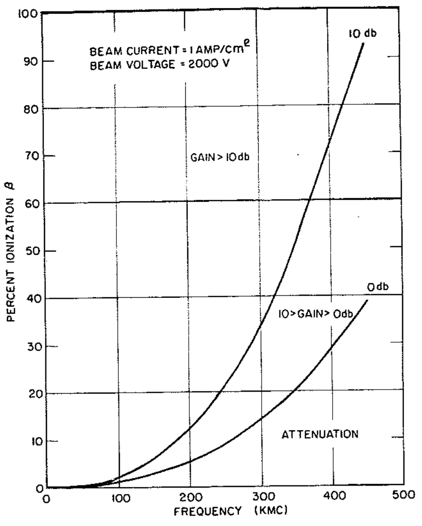


Fig. 1—Minimum percent ionization in a cesium plasma as a function of frequency for 0 db and 10 db gain.

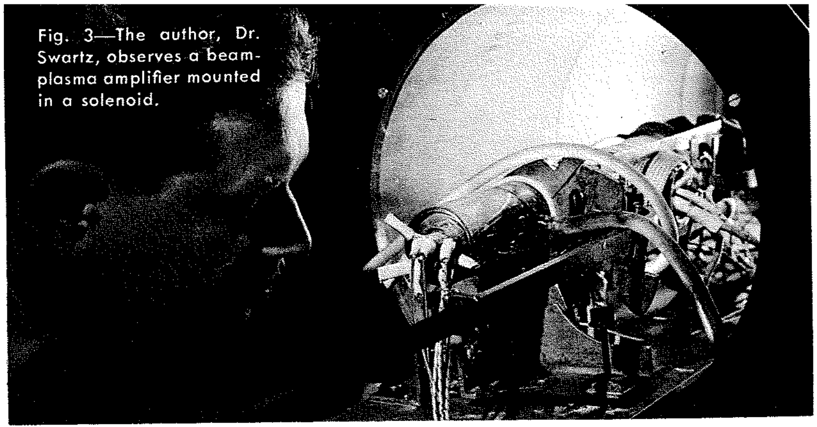


Fig. 3—The author, Dr. Swartz, observes a beam-plasma amplifier mounted in a solenoid.

Dr. GEORGE A. SWARTZ received his BS from MIT in 1952. In 1954 he received his MS and in 1958 his PhD in physics, both from the University of Pennsylvania. His doctoral dissertation was in the general field of solid-state physics. At RCA Laboratories, he has specialized in research relating to microwave phenomena in gaseous plasmas, plasma propulsion, plasma stability, and plasma diffusion. His work in collaboration with Louis Napoli on amplification at 24 Gc by the interaction of an electron beam with a Cs plasma was cited by "Industrial Research" magazine as one of the 100 most important achievements in 1963. Dr. Swartz has received an "RCA Achievement Award" and has published several papers concerned with plasmas, solid-state physics, and instrumentation. He is a member of the American Physical Society and Sigma Xi.

the plasma. Fig. 2 illustrates the flow of coherent and incoherent power in the beam-plasma system. If the plasma electron collisions are considered to be the only power loss mechanism in the plasma, the maximum efficiency η of a beam-plasma amplifier is

$$\eta = \frac{1}{8}\left(\frac{n_b}{n_p}\frac{\omega}{\nu}\right)^2 \qquad (3)$$

Where: n_b is the electron density of the beam.

EXPERIMENTS

In 1958 Boyd, Field, and Gould[5] experimentally verified the Bohm and Gross theory. In their experiment an electron beam was sent through an input helix, a mercury plasma and an output helix. Microwave power at s-band was coupled to the electron beam by the input helix. The density of the mercury plasma was varied continuously. As the density of the plasma reached a critical value the power from the output helix increased sharply. The beam-plasma amplifiers that have been constructed subsequently were of similar design.[6-9] Chernov and Bernashevski have observed 30-db to 40-db power gain at frequencies up to 38 Gc in a mercury plasma.[9] This is the highest reported frequency achieved by such a device.

A 23-Gc beam-plasma amplifier (Figs. 3, 4) was recently tested at RCA Laboratories.[10] The tube contains an electron gun, input helix, discharge electrodes for plasma production, output helix, electron beam collector, and a cesium pump system. Cesium vapor is pumped into the discharge region from a heated liquid supply of the metal under the main tube envelope. The vapor is condensed on the surface of waterjackets in front of the electron gun and allowed to flow back to the main supply. With this system cesium pressures of 0.01 to 0.10 torr are maintained in the discharge region and pressures less than 10^{-5} torr exist in the electron gun region. A cesium plasma is produced by a hot-cathode, hot-anode Penning arc. A 1,400-volt electron beam is sent through an input helix, the cesium plasma and an output helix. The beam-plasma interaction in a 3-cm length of plasma produced a 40-db power gain. Losses in the input helix coupling, output helix coupling and windows cut the net power gain to 8 db. The electronic gain which is plotted as a function of plasma probe current (or plasma density in arbitrary units) for several input power levels is shown in Fig. 5. At the higher input levels, a pronounced minimum is observed in the gain curves. The minimum occurs at a plasma density which produced a maximum gain at low power inputs. The gain minimum is attributed to a decoupling of the space charge wave on the beam and the slow wave on the output helix as higher AC power levels are generated on the beam. The decoupling is caused by a reduction of the beam velocity which is a result of beam energy loss by microwave absorption in the plasma, as described in Fig. 2. The 1-Mc density instabilities present in the plasma were reflected in the detected output signal of the amplifier. An example of the output signal is shown in Fig. 6.

The cesium Penning arc used in the 23-Gc amplifier can produce the highly ionized dense plasma which is required in a millimeter-wave generator.[2] The unusual feature in this arc is the hot cylindrical anode constructed from a high-work-function metal such as tungsten or tantalum. The cesium atoms ionize as they strike the hot metal surface and the ions, thus formed, are continually fed into the arc. This method for supplying ions to the arc eliminates the need to supply them by electron impact with neutral atoms, which is the ion producing mechanism in all other arcs. The electrons are supplied to the arc from hot cathodes at the ends of the anode cylinder and are trapped in the arc by the combination of a negative potential on the cathodes and a strong magnetic field parallel to the anode cylinder axis. With a 7,500-gauss magnetic field, the cesium

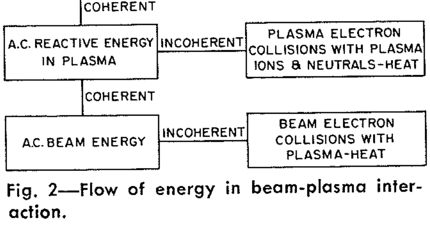


Fig. 2—Flow of energy in beam-plasma interaction.

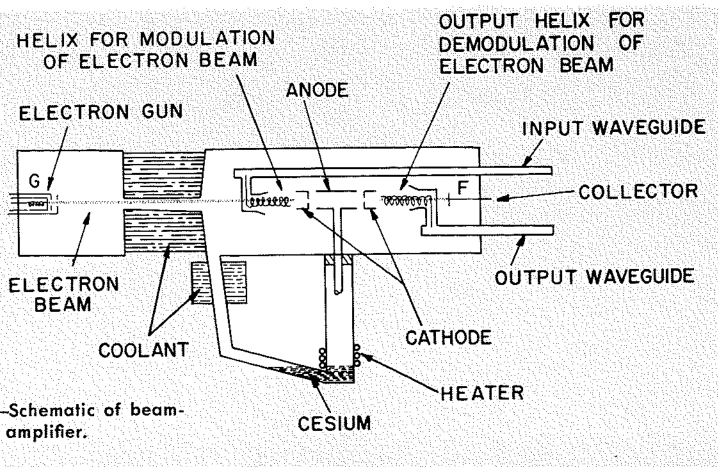


Fig. 4—Schematic of beam-plasma amplifier.

Penning arc produced a plasma density of 10^{15} ions/cm^3, ($\omega_p/2\pi = 300$ Gc) with an ionization of 50%.[2]

Large ion density fluctuations are normally present in a low pressure Penning arc. However, at plasma densities equal to or greater than 10^{15} ions/cm^3, the ion collisions with neutral atoms causes damping of the density fluctuations. Thus, a stable highly ionized dense plasma is available for use in a millimeter wave generator.

FUTURE INVESTIGATION

One substantial problem remains to be solved before a 300-Gc beam-plasma amplifier can be realized. A mechanism for coupling millimeter-wave power to the amplifier *without* resorting to mechanical structures (such as a helix or cavity) must be developed. Two coupling systems have been proposed and investigated. One system requires the presence of a strong axial magnetic field.[6,11] The field causes power gain in the frequency range $\omega_p < \omega < \sqrt{\omega_p^2 + \omega_c^2}$, where ω_c is the electron cyclotron frequency. The plasma waves, which are present in this same frequency range, are characterized by small radial wave vectors or large radial wavelengths which extend to the plasma edge. Since the AC fields are present at the plasma boundary, power may be coupled directly to a waveguide or free space.

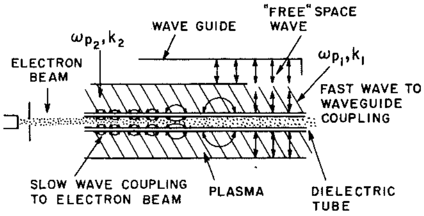
Another system for coupling millimeter-wave power to a beam-plasma amplifier utilizes the backward-wave propagation characteristics of a surface space-charge wave on a tubular plasma.[12] Typical dispersion curves for a tubular plasma and the coupling wave vectors are shown in Fig. 7. A free-space wave at a frequency ω is coupled to the backward-wave mode in the tubular plasma by adjusting the plasma density so that the phase velocities of the two waves are equal. The backward-wave travels to another plasma region where the plasma density increased so that the phase velocity of the backward-wave and the electron beam velocity are equal in direction and magnitude. Power is then coupled between the backward-wave mode and the electron beam. A schematic drawing of a tubular plasma coupling system is shown in Fig. 8.

For a 300-Gc amplifier, which utilizes a 50% ionized cesium plasma, a tube efficiency greater than 10% requires an electron beam density greater than 5,000 amp/cm^2. Electron beam current densities of 800 amp/cm^2 have been produced,[13] and in the next few years, even higher current densities may be developed.

CONCLUSION

Electron beam-plasma amplifiers show promise of providing amplification and generation of millimeter waves. The gain per unit length in such amplifiers is very high because of the close proximity between the electron beam and plasma. However, the power output is noisy because of beam electron collisions with the plasma particles.[9] Plasmas with sufficient density and ionization for use in millimeter-wave amplifiers have been produced. Power coupling to such a high-frequency beam-plasma device is a formidable problem which must be solved. The efficiency of a millimeter-wave beam-plasma amplifier will be low unless electron beam densities greater than 5,000 amp/cm^2 are developed.

Fig. 8—Tubular plasma coupler.



ACKNOWLEDGEMENTS

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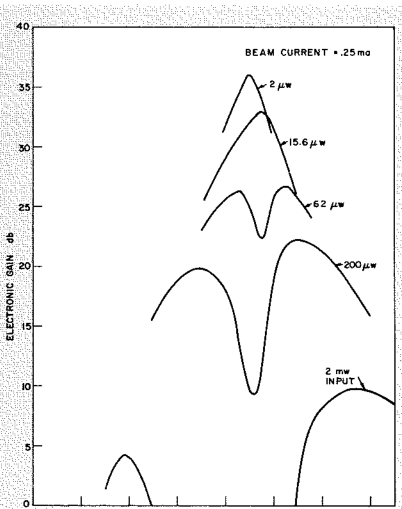
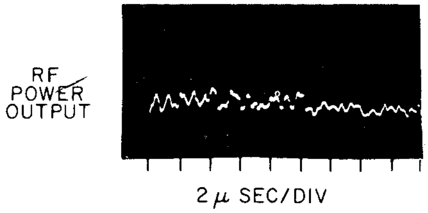


Fig. 5—Electronic gain of a 23-Gc beam-plasma amplifier as a function of probe current in the plasma.

Fig. 6—The 1-Mc fluctuation of 23 Gc power during one cycle of modulation.



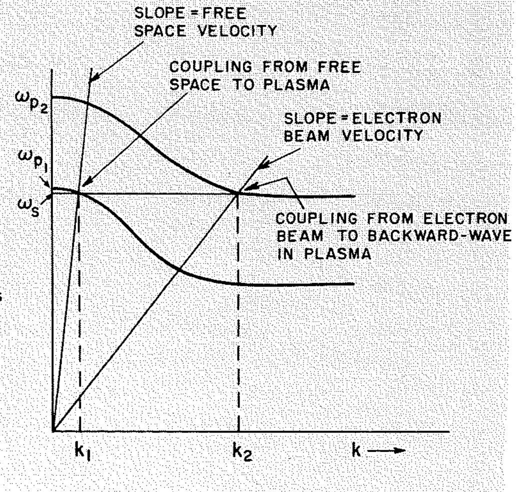


Fig. 7—Backward-wave dispersion curves showing coupling from free space through plasma to electron beam. Coupling is between free space and backward-wave at point 1, where plasma frequency is ω_{p1} and wave vector is k_1. Coupling is between backward-wave and space charge wave on the electron beam at point 2, where plasma frequency is ω_{p2} and wave vector is k_2.

MICROWAVE TUNNEL-DIODE AMPLIFIERS AND OSCILLATORS

The characteristics required of tunnel diodes for efficient use in microwave amplifiers and oscillators are discussed. The development of suitable diodes for specific devices—a C-band tunnel-diode amplifier and an L-band tunnel-diode oscillator—is described. The general design requirements and the problems encountered in the development of these devices are explained. A circuit description and approximate expressions for gain, bandwidth, and noise figure of tunnel-diode amplifiers are given. The effects of variations in the supply voltage, temperature, and load on the frequency stability of tunnel-diode oscillators are considered.

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A tunnel diode is a *p-n* semiconductor junction having a high carrier concentration.[1,2] It exhibits a negative incremental resistance over a range of bias voltages at all frequencies from DC to a maximum value referred to as the resistive cutoff frequency of the diode. At low frequencies, the characteristics of tunnel diodes are not extremely critical and it may be possible to use a given type of diode for a wide range of applications. In the microwave region, however, the characteristics required of the tunnel diode usually approach the state-of-the-art limits, and it is often necessary to develop a tunnel diode specifically for the intended application.

MICROWAVE TUNNEL-DIODE AMPLIFIERS

Tunnel-diode amplifiers have become competitive with other low-noise microwave amplifiers, such as masers and parametric amplifiers. Their noise figures are between 3 and 5 db, which are only slightly higher than those of masers or parametric amplifiers, and their other characteristics (i.e., bandwidth, linearity, temperature stability, power output, and radiation resistance) are equal to or better than those of other types. The most attractive features of the tunnel-diode amplifier are its small size, low DC-power consumption, low cost, and absence of an RF pump.

Tunnel Diodes for Microwave Amplifiers

Critical tunnel-diode characteristics for microwave amplifiers are noise constant, cutoff frequency, and inductance.

The major source of noise is the shot noise contribution of the diode. A measure of the diode noise is the noise constant, I_oR_n, which is determined by the semiconductor material and is a function of the slope of the tunnel diode current-voltage characteristic curve in the negative resistance region. The quantities I_o and R_n are the current and the negative resistance at the bias point.

Of all the available semiconductor materials, only germanium and gallium antimonide are capable of meeting the frequency and noise-voltage requirements of a low-noise microwave amplifier. Although germanium has a slightly higher noise constant than gallium antimonide, its use is advantageous because the gain of germanium diodes is more stable with respect to temperature and power-supply variations. Also, the germanium tunnel diode does not show permanent electrical degradation, which has been observed by Carr[3] in forward-biased gallium antimonide tunnel diodes.

The cutoff frequency of the diode must be two to three times the operating frequency of the amplifier for low-noise amplification. The cutoff frequency is:

$$f_c = \frac{1}{2\pi|R_n|C_j}\sqrt{\frac{|R_n|}{r_s} - 1}$$

Where: R_n is the diode negative resistance, C_j is the junction capacitance, and r_s is the series resistance of the diode. The values of R_n and C_j are primarily dictated by the required amplifier performance and the requirement that the diode be stably biased. Thus, a high cutoff frequency depends mainly on the reduction of the diode series resistance. With the conventional direct-alloyed-junction method of fabricating tunnel diodes, the etching required to obtain the proper peak current produces a very narrow neck directly under the junction area and results in a relatively high series resistance. An RCA-developed process, which uses an epitaxial solution-regrowth technique together with an inert-area masking technique, prevents the neck formation during etching of the tunnel diodes. With this process, diodes having cutoff frequencies of 20 to 30 Gc are easily obtained, and laboratory samples having cutoff frequencies in excess of 40 Gc have been fabricated.

In order to obtain the high degree of stability that is usually required of a microwave tunnel diode amplifier, the diode should be short-circuit stable:

$$L < r_s\,|R_n|C_j$$

Where: L is the series inductance of the diode. Because the inductance of the junction is small, the attainment of a low value for L is primarily a mounting and packaging problem. Fig. 1 shows a low-inductance package in which a very low inductance screen is used to connect the junction to the package.

A Low-Noise, C-Band Amplifier

Either transmission or reflection types of amplifiers can be realized from the use of tunnel diodes in microwave circuits. These types have been generally discussed in the literature in great detail.[4] The most stable and simplest type of tunnel-diode amplifier proved to be the circulator-coupled reflection circuit. Because this type is the most widely used, it is the one that is discussed here.

A block diagram of a circulator-coupled tunnel-diode amplifier is shown in Fig. 2. The circulator, because of its preferential propagation characteristic, is capable of providing isolated input and output ports and of reducing the adverse effects of source and load mismatches on amplifier performance. Low-loss four-port circulators having a 20% bandwidth at C-band and providing 25 to 40 db of isolation are currently available. The amplifier design is based upon the knowledge of the tunnel-diode parameters and the circulator characteristics. An equivalent circuit of a reflection type of tunnel-diode amplifier is shown in Fig. 3. In this circuit, Z_c is the circulator impedance, Z_o is the characteristic impedance of the transmission line, R_B and C_B are the elements of the bias network, and L_o is a tuning inductance. At midband, C_B approaches infinity, and if it is assumed that Z_c is real and equal to Z_o, then the power gain can be approximated by:

$$G_P = \left[\frac{Z_o + |R_n|}{Z_o - |R_n|}\right]^2$$

Where Z_o must be smaller than $|R_n|$ to obey stability requirements. If the gain is large (greater than 15 db), then the voltage gain-bandwidth product can be approximated by:

$$\mathrm{G_vB} = \frac{1}{\pi \mathrm{R}(\mathrm{C_j} + \mathrm{C_c})}$$

And, the noise figure by:

$$F = \frac{1 + 20I_o|R_n|}{\left(1 - \frac{r}{|R_n|}\right)\left(1 - \left[\frac{f}{f_c}\right]^2\right)}$$

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* Messrs. Nelson and Presser are located in Princeton; Messrs. Casterline and Minton are located in Somerville, N.J.

Where: f is the frequency of operation.

A tunnel-diode amplifier was developed for the RCA CW-60 microwave system.[7] The amplifier was of coaxial design and used a four-port circulator and a germanium diode having a cutoff frequency of 22 Gc, a noise factor (I_oR_n) of 65 mv, and a series inductance of 100 ph.

The gain response and the saturation characteristic of this amplifier are shown in Figs. 4 and 5, respectively. The input port-to-output port noise figure was 4.8 db (including a 0.3-db circulator loss). The gain variation over the temperature range from −20°C to +70°C was less than ±0.5 db, and the amplifier was stable with any source or load mismatch. A photograph of the amplifier is shown in Fig. 6.

MICROWAVE TUNNEL-DIODE OSCILLATORS

Although tunnel-diodes are not, as yet, used extensively in microwave oscillators, it is likely that they will be in the future, in applications where small, low-cost signal sources are required without extreme frequency stability.

Tunnel Diodes for Microwave Oscillators

The tunnel diodes used in oscillator applications are usually gallium-arsenide diodes. The voltage and the impedance for a given peak current of a gallium-arsenide diode are each about twice those of a germanium diode. Thus, for a given impedance level, a gallium-arsenide diode has four times the power capability of a germanium diode.

As higher-peak-current diodes are used in order to obtain higher power output from microwave tunnel-diode oscillators, the inductance of the diode becomes increasingly important. If the inductance is high, it not only becomes difficult to obtain the desired operating frequency but it is also difficult to load the diode so that a good portion of the available power will be obtained as output power. Initial evaluation tests of diodes mounted in conventional tunnel-diode packages, such as that shown in Fig. 7, indicated that the diode inductance of 265 ph was too high. The inductance is proportional to the diode height, and efforts to decrease this parameter to an acceptable value resulted in the development of the reduced-height strip-line package shown in Fig. 8. This package consists of a gold-plated copper disk $\frac{1}{32}$ inch thick and $\frac{1}{4}$ inch in diameter. The semiconductor pellet is mounted in the center of the disk, and a stripline washer, made from 0.002-inch-thick, copper-clad teflon, having an outer diameter of 0.125 inch and an inner diameter of 0.030 inch, is soldered around the pellet. A screen provides the contact to the diode. This fabrication

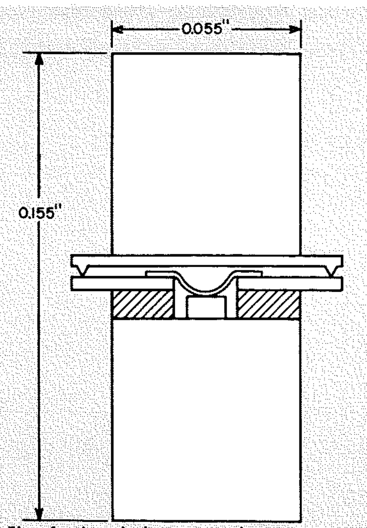


Fig. 1—Low-inductance microwave tunnel-diode package.

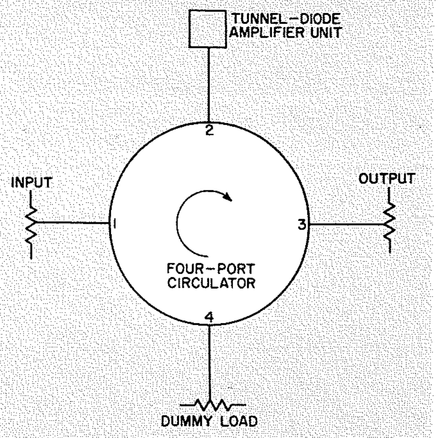


Fig. 2—Tunnel-diode amplifier.

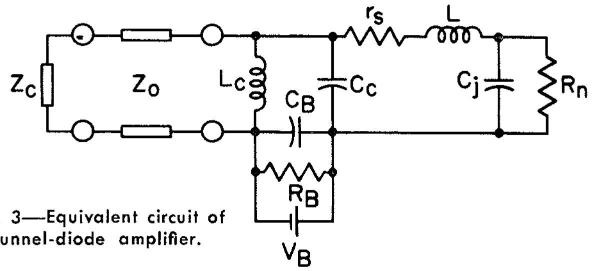


Fig. 3—Equivalent circuit of a tunnel-diode amplifier.

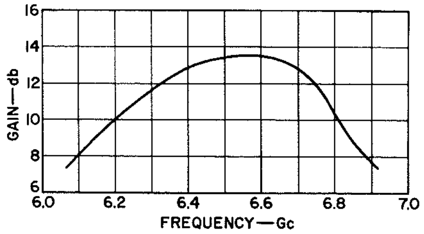


Fig. 4—Gain of the C-band tunnel-diode amplifier vs. frequency.

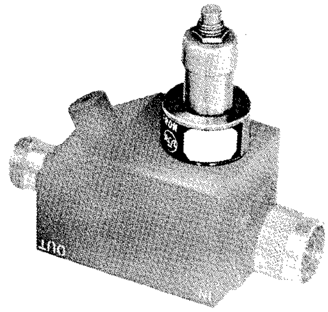


Fig. 6—C-band tunnel-diode amplifier.

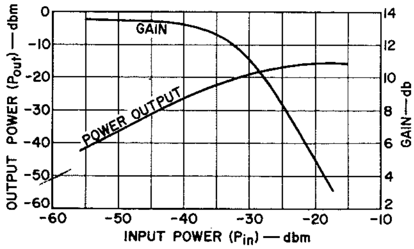


Fig. 5—Saturation characteristic of the C-band tunnel-diode amplifier.

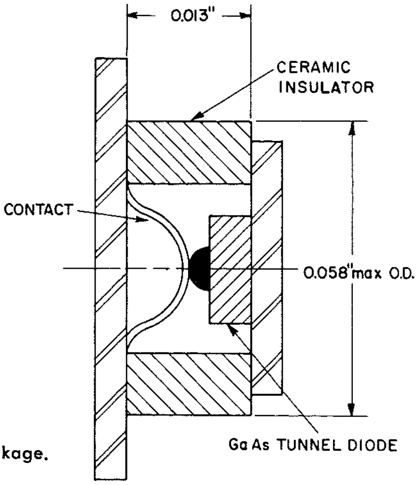


Fig. 7—Conventional tunnel-diode package.

technique has reduced the diode inductances to about 75 ph.

High-peak-current tunnel diodes having high cutoff frequencies must use semiconductor material having a very high carrier concentration and must be fabricated by techniques that result in a low series resistance. The cutoff frequency of a tunnel diode increases as the diode is etched. Therefore higher cutoff frequencies are obtained by making diodes that have peak currents greater than required and etching them to reduce the current to the desired level. The diodes with 600-ma peak current used in the 20-mw L-band oscillator described in the next section were made by etching diodes having initial peak currents of 800 ma to 1,400 ma. Because of these high initial currents, a carrier concentration for the semiconductor material of 8 to 9×10^{19} atoms/cm^3 was required. Attempts to obtain this carrier concentration by diffusion or solution-regrowth techniques were unsuccessful. However, a method of growing doped crystals developed by the RCA Industrial Semiconductor Operations Department provided material having the required carrier concentration. The series resistance of the diode is greatly affected by the length of the neck formed as the diode is etched. This length is minimized by using the smallest possible alloy dots to form the tunneling junction.[5]

The use of these techniques has made possible diodes having the following typical characteristics:

Peak Current	600 ma
Capacitance	30 pf
Series Resistance	0.25 ohm
Cutoff Frequency	8 Gc

Such diodes were employed in the 20-mw-output L-band oscillator.

An L-Band, 20-mw Oscillator

Because the operating voltage of a tunnel-diode oscillator is limited by the negative resistance characteristic of the diode, the attainment of high power output requires the use of high-current diodes. The maximum power output of a tunnel-diode oscillator may be calculated from the curves of effective negative resistance of the diode for various values of bias voltage and RF-voltage swing. For gallium-arsenide tunnel diodes, the generated power output for optimum bias voltage and RF-voltage swing is approximately 90 mw per ampere of peak current. The series-resistance, capacitance, and inductance limitations of diodes currently available reduce the obtainable power to about one half of the generated power at frequencies of 1,000 to 2,000 Mc. Thus, if an oscillator is to supply 20 mw of power output at 1,700 Mc a diode having a peak current of about 600 ma is required.

A gallium-arsenide diode having a peak current of 600 ma has a resistance of about ⅛ ohm. At a frequency of 1,680 Mc, it is convenient to use a distributed-constant resonant circuit. For this particular application, a strip transmission line having a dielectric of 0.002-inch-thick teflon was chosen. This stripline has a very low characteristic impedance (½ ohm) and, thus, simplifies the problem of avoiding low-frequency resonances. A typical circuit, shown in Fig. 9, consists of a half-wavelength open line on either side of the diode. A stabilizing resistor is placed at a voltage node point, a quarter wavelength from the diode, so that little power is absorbed by the resistor. A two-section quarter-wavelength transformer transforms the 50-ohm load to an impedance of about 1 ohm at the diode position. Tuning is accomplished by a variable capacitor which shunts the output line at the junction of the two transformer sections.

Oscillator Characteristics

If the transformation ratios and lengths of the two quarter-wave transformer sections of the oscillator are properly chosen, the variable capacitor can be adjusted to vary the reactance to the diode without any appreciable change in resistance over a limited tuning range (1,660 to 1,700 Mc). Thus, the variation in oscillator power output over the tuning range is minimized. The power output of the oscillator as a function of frequency is shown in Fig. 10.

Variations in the supply voltage of a tunnel-diode oscillator produce changes in both the capacitance of the diode and the harmonic content of the oscillator wave shape and, thus, tend to affect the operating frequency of the oscillator. The capacitance of the tunnel diode increases as the supply voltage increases and the frequency of the oscillator thus tends to decrease. An increase in the supply voltage causes the wave shape of the oscillator to become more sinusoidal as this voltage approaches the valley voltage of the diode. As the supply voltage is increased beyond this point, the wave shape then becomes progressively less sinusoidal. That harmonics reduce the fundamental frequency of a negative-resistance oscillator has been discussed by Van Der Pol.[6] Thus, as the supply voltage approaches the valley voltage of the diode, the reduction in harmonic content causes the fundamental frequency to increase; beyond this point, the frequency begins to decrease.

In low-power oscillators, the effects of variations in diode capacitance and in the harmonic content of the oscillator with changes in the supply voltage may cancel each other, and the oscillator frequency will remain essentially constant over a relatively wide supply-voltage range. However, for microwave oscillators which use high-speed, high-current diodes, the effect of harmonic variations is predominant, and a substantial increase in frequency occurs as the supply voltage is increased.

In the L-band oscillator described here, the frequency may increase as

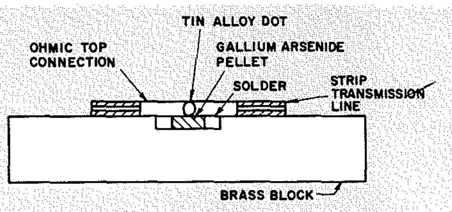


Fig. 8—Stripline tunnel-diode package.

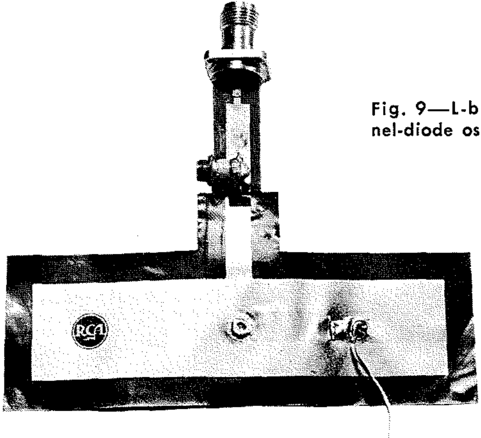


Fig. 9—L-band tunnel-diode oscillator.

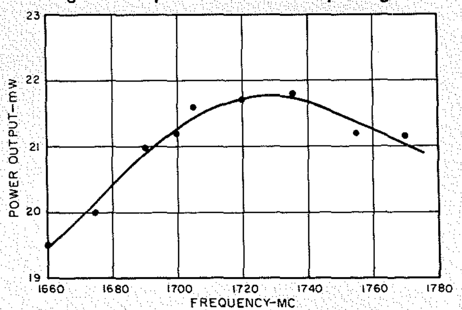


Fig. 10—Power output of the L-band tunnel-diode oscillator vs. frequency.

much as 50 to 100 Mc for a 20% increase in the supply voltage. Because this large frequency shift could not be tolerated in the application of the oscillator, the transistor regulator circuit shown in Fig. 11 was developed. This circuit reduces a 20% change in the supply voltage to a change of less than 1% in the voltage applied to the tunnel-diode oscillator. As a result, the change in frequency in response to supply-voltage variations is reduced to about 4 Mc. The power output and frequency of the regulated tunnel-diode oscillator as functions of the supply voltage are shown in Fig. 12.

CONCLUSIONS

The germanium tunnel-diode amplifier described in this paper has a gain of 13.5 db at a center frequency of 6.6 Gc with a 10% bandwidth and a 4.8-db noise figure. Such amplifiers may be designed for a gain of 20 db, operating at frequencies of 1 to 10 Gc with noise figures of 4 to 5.5 db. Thus, tunnel-diode amplifiers have become competitive with other types of low-noise amplifiers, such as masers and parametric amplifiers. Although the noise figures of tunnel-diode amplifiers are higher than those of masers or parametric amplifiers, the other characteristics of these amplifiers, such as bandwidth, linearity, temperature stability, power output and radiation resistance, equal or exceed those of other types. The most attractive features of the tunnel-diode amplifier are its small size, low DC power consumption, low cost, and absence of a need for an RF pump. Such amplifiers are currently being used as low-noise pre-amplifiers in receivers for relay links and radars and are being considered for use in phased-array radar systems. They have thus found their way into commercial and military systems.

The 20-mw tunnel-diode oscillator, tunable from 1,660 to 1,700 Mc, illustrates that, at this power level, microwave tunnel-diode oscillators have certain disadvantages. Because of the low circuit impedance required, which limits the attainable Q-factor and of the frequency variation caused by harmonic-frequency depression, the frequency stability for changes in supply voltage, load impedance, and ambient temperature leaves much to be desired. A regulator circuit may be used to reduce supply-voltage variations; isolators may be used to limit load impedance changes; and temperature-compensated circuits can be designed. These techniques, however, result in a complicated and expensive device; therefore, tunnel-diode oscillators cannot, at present, successfully compete with such devices as harmonic generators at power levels of 20 mw and above. For lower power units, for example, in the range of 1 to 5 mw, however, the prospects are better. The frequency variations are smaller, and the higher-impedance diodes permit higher-Q circuits which reduce the variations even further. Such oscillators offer advantages of low cost, small size, and tunability for applications where stringent frequency stability is not required. At the present time, oscillators that provide power outputs of 5 mw at frequencies up to C-band, and 1 mw at X-band are possible.

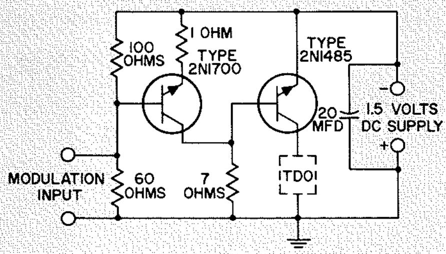


Fig. 11—Transistor supply-voltage regulator.

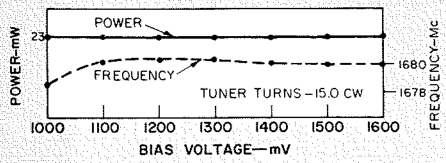
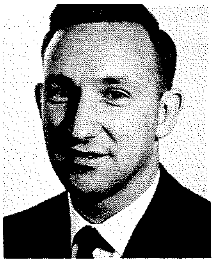


Fig. 12—Power and frequency of the regulated tunnel-diode oscillator vs. supply voltage.

D. E. Nelson

A. Presser

E. Casterline



R. M. Minton

DONALD E. NELSON received his BSEE from the University of Illinois in 1941 and his MS in 1948. From 1941 to 1945 he worked as a development engineer at Westinghouse Electric Corporation. From 1945 to 1949 he was Research Associate at the University of Illinois Tube Laboratory on CW magnetrons and millimeter-wave devices. He joined RCA in 1949 as development engineer and has since that date designed a variety of pulse and CW magnetrons. From 1954 to 1958 he was Engineering Leader with technical direction of all magnetron development in Microwave Tube Engineering at Harrison, New Jersey. From 1958 to 1959 he worked as a system engineer on microwave tubes for advanced defense systems. In 1959 he became a Member of the Technical Staff in the Microwave Tube Applied Research activity at Princeton, N.J., where he has worked on development of tunnel-diode oscillators and amplifiers, transistor oscillator-multipliers, and varactor frequency multipliers. Since 1963, he has been Engineering Leader of the Solid-State Microwave group.

ADOLPH PRESSER received the BEE from the Institute of Technology, Vienna, Austria, in 1950 and the MEE degree from the Polytechnic Institute of Brooklyn, Brooklyn, New York, in 1961, where he is currently studying for his doctorate. From 1950 to 1952 he was a production engineer for the Schrack A.G. in Vienna, Austria, and was employed in industry from 1953 to 1954 and was a development engineer for the Allied Control Co. in New York, New York, from 1954 to 1959. He joined the RCA Electron Tube Division in 1959 and is now with the Microwave Applied Research group at Princeton, New Jersey. He has been engaged in the development of various solid state microwave devices. This work includes the design and development of parametric amplifiers, tunnel diode amplifiers, tunnel diode frequency converters, and tunnel diode oscillators.

EDGAR T. CASTERLINE received his BS in Chemistry from Manhattan College in 1956. He is currently studying for an MS in Physical Chemistry at the Stevens Institute of Technology. From 1956 to 1957, he was employed by the Western Electric Company as an analytical chemist. He entered the military in 1957 and was assigned to Fort Monmouth, New Jersey, where he spent two years engaged in work on thin-film resistors and capacitors and on thermoelectric cooling devices. In 1959, he joined the Micromodule group of the RCA Semiconductor and Materials Division where he worked chiefly on the development of thin-film resistors for micromodules. In 1960, he joined the Computer group, concentrating on development of tunnel diodes. One of these main developments was germanium tunnel rectifiers for the Lightning project, as well as for the commercial tunnel-diode program. In 1963, he was assigned to the Industrial Products group, where he is continuing work on the development of both GaAs and Ge tunnel diodes.

ROBERT M. MINTON received his BSChE from Newark College of Engineering in 1952. Prior to joining RCA in 1959, he worked at the U.S. Army Signal Corps, Ft. Monmouth, doing development work in the fields of reinforced plastics and thin films. At RCA, Mr. Minton worked with the Micromodule group until 1960, when he was assigned to the Computer Department, where he developed high-speed tunnel diodes for Project Lightning. After being transferred to the Industrial Department in 1963, he continued working with tunneling devices and developed a microwave diode for use in the CW-60 microwave amplifier. Mr. Minton is presently responsible for the design, development, and production of all RCA tunnel diodes and tunnel rectifiers.

ACKNOWLEDGMENTS

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MIPIR RADAR RECEIVER—A THREE-CHANNEL REMOTE-TUNED PARAMETRIC AMPLIFIER

The parametric amplifiers described herein have improved radar tracking sensitivity. Single-knob remote tuning allows changes of tracking assignments between beacon and skin, as well as to providing fine tuning to compensate for transmitter or beacon drift. The amplifier has been installed in several operational radars and its performance is being evaluated.

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The functional dependence of radar range upon system parameters is given by the well known range equation:

$$R = \sqrt[4]{\frac{P_t G^2 \lambda^2 \sigma}{(4\pi)^3 S}} \qquad (1)$$

Where: P_t = peak transmitted power, G = antenna gain, λ = transmitted wavelength, σ = effective target area, S = receiving power, and R = radar range. If S is a minimum, then R is a maximum The detectable signal of the receiver is limited by the received noise, thus Eq. 1 illustrates the importance of the receiver noise figure in determining ultimate radar sensitivity, or range. Reducing the receiver noise figure has the same effect on the range as increasing the transmitted power. Increase of transmitted power would involve a tremendous increase in system cost, size, and weight. The range improvement can be obtained by using a parametric amplifier which has low noise characteristics.

The MIPIR radar[1] (*MI*ssile *P*recision *I*nstrumentation *R*adar) is a monopulse system consisting of three parametric receiving channels. Such a system requires minimum gain and phase differences between channels. Achievement of these conditions requires both a stable common pump source and the capability of frequency tracking the three channels.

Other design factors which must be considered are the ability to maintain tracking despite parametric amplifier failure during operation, to withstand the severe shock environment due to the movement of the antenna and to install in a limited amount of space.

PARAMETRIC AMPLIFIER

The regenerative parametric amplifier was chosen as the optimum to meet all of these objectives. The amplifier uses a minimum number of components, which relieves the space problem.

The regenerative parametric amplifier is a one-port device which utilizes a circulator to separate the input and output terminals. Fig. 1 shows the equivalent circuit of such an amplifier. The available power gain of a one-port device is merely the square of the reflection coefficient. At resonance,

$$\text{GAIN} = \left|\frac{G_o - G_{in}}{G_o + G_{in}}\right|^2 \qquad (2)$$

Where: G_o is the source conductance and G_{in} is the input conductance of the amplifier. It has been shown that if only two frequencies, signal and idler (pump frequency minus signal frequency) are brought out, G_{in} is negative for a parametric amplifier. Hence the reflection coefficient is always greater than 1 and the parametric amplifier exhibits power gain. When G_{in} approaches G_o, the amplifier oscillates. Therefore, a regenerative parametric amplifier provides power gain *if and only if* its input admittance has a negative real part.

The value of the negative input conductance is primarily governed by the pump power and the idler frequency. Hence the pump power and pump frequency must be very well regulated in order to achieve stability.

The dependence of amplifier gain as a function of $G_o/|G_{in}|$ is shown in Fig. 2. The curve illustrates that the gain stability is more sensitive to small changes of $G_o/|G_{in}|$ for high values of amplifier gain than for low values of amplifier gain. The variation of amplifier gain depends upon the change of G_o or G_{in}, or both.

The sensitivity of gain variation can be further illustrated by Fig. 3, which assumes that the source impedance has a VSWR of 1.1 from the nominal value of 50 ohms. For example, a nominal gain of 15-db gain could change from 13 db to 17.6 db depending upon the phase of terminal reflection coefficient if the amplifier regeneration remains the same. Therefore, the impedance of the circulator for a circulator-coupled amplifier, which is the case, must be specified to a very close tolerance for broad frequency operation. Thus gain stability of a regenerative parametric amplifier is greatly enhanced at low gain operation.

There is an optimum pump frequency for obtaining gain and minimum noise figure. Based upon the considerations of the available diode types, klystron and diode cutoff frequencies, the optimum pump frequency for C-band operation is in the K_u-band, from 15 to 20 Gc.

C-BAND PARAMETRIC AMPLIFIER

Fig. 4 shows the *c*-band parametric amplifier consisting of waveguide and coaxial tee junction. Pump power in K_u band is fed to the diode through the reduced height waveguide such that no extra post is required for mounting the diode. Along the path of the pump power, a section of waveguide is narrowed so that a high-pass filter is formed which provides minimum attenuation at pump frequency and yet isolates the pump circuit from the idler frequency. Between the high-pass filter and the short at the end of the waveguide, an idler cavity is provided which furnishes the necessary resonant circuit at the difference frequency to allow regeneration at the signal frequency.

The waveguide short serves the dual functions of providing optimum position for pumping the varactor diode and locating the idler cavity. The idler short is located inside the pump short. Thus, the amplifier body is very compact.

The signal is fed coaxially into the diode through several sections of low-pass filter and is broadband matched in

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Final manuscript received August 20, 1964.

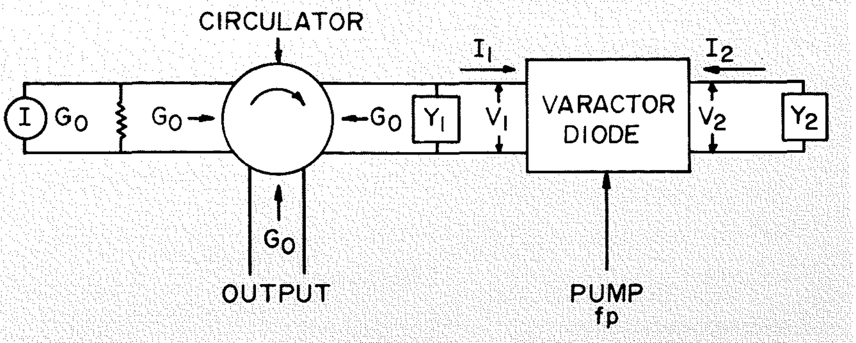


Fig. 1 — Equivalent circuit of regenerative paramp.

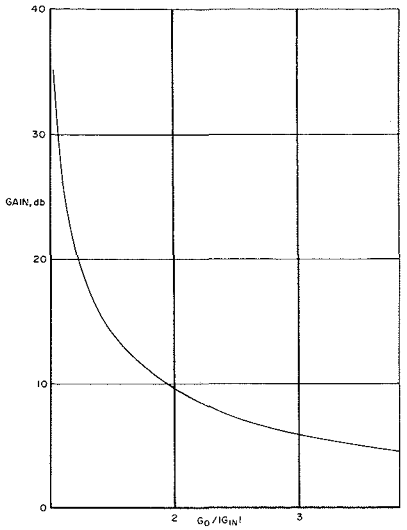


Fig. 2—Power Gain VS $G_0/|G_{in}|$.

the frequency range from 5.4 to 5.9 Gc. Instantaneous response is achieved by adjusting a dielectric vane located between the high-pass filter and the diode. The displacement of this dielectric vane determines the resonance of the idler circuit. Thus, one-knob tuning is obtained without any adjustment of pump power and diode bias while still maintaining relatively flat gain over the full tuning range. This provides an easy means of remote tuning and eliminates conventional mechanically movable contacts, which usually cause erratic results.

To maintain stability of pump power and frequency, a temperature-controlled K_u-band klystron was used in conjunction with a well-regulated power supply. This klystron is used as a common pumping source for the set of three paramps, as shown in Fig. 5. This is essential for a monopulse tracking radar in order to achieve the required differential phase and gain stability.

A switchable circulator was used at the signal input to provide a fail-safe feature. Under the normal operating condition the incoming signal is amplified and flows as indicated by the solid line in Fig. 5. In the event of failure due to diode burn-out, pump power, etc., a positive DC voltage could be applied to the circulator so that the signal is directed to the mixers and the parametric amplifiers are by-passed. The radar system would then remain operative at reduced sensitivity.

MECHANICAL ASSEMBLY

Since there are two frequencies involved in a monopulse tracking radar operation, skin and beacon, mechanical means are provided to preset the idler cavity tuning for these two frequencies. The mechanical assembly consists of the electromechanical devices which provide proper positioning of the idler tuning element, as shown in Fig. 6. The tuning element is mechanically linked to a solenoid armature by means of shouldered shaft. If the armature were permitted to travel freely, its position when the solenoid was energized and unenergized would represent the extremes of travel of the tuning element in the idler cavity. The maximum insertion depth of the element, which is obtained when the solenoid is not energized, corresponds to the parametric amplifier tuned to the highest signal frequency. Conversely, when the solenoid is energized, the armature is pulled in against spring tension, and the parametric amplifier is tuned to its lowest signal frequency.

Since the tuning range is 5.4 to 5.9 Gc, the depth-of-tuning element insertion must be so adjusted that the energized and unenergized states of the solenoid represent two separate frequencies within this range. When properly preset, these two frequencies will correspond to the skin and beacon frequencies which are employed in the radar system.

The limits of tuning-element insertion corresponding to skin and beacon frequencies are fixed by the position of the arms of two bell cranks. The bell cranks stop the shaft movement at the precise tuning-element insertion depth corresponding to the preset frequency desired. The position of the bell crank is fixed by the micrometer movable spindle, mechanically connected to the flexible cables from a fine-tuning drive motor. The location of the bell crank is governed by

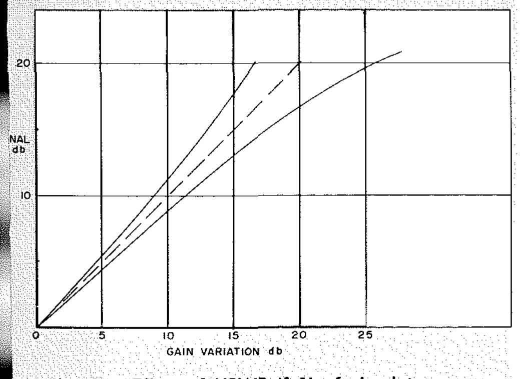


Fig. 3 — Effect of VSWR (1.1) of circulator on paramp gain.

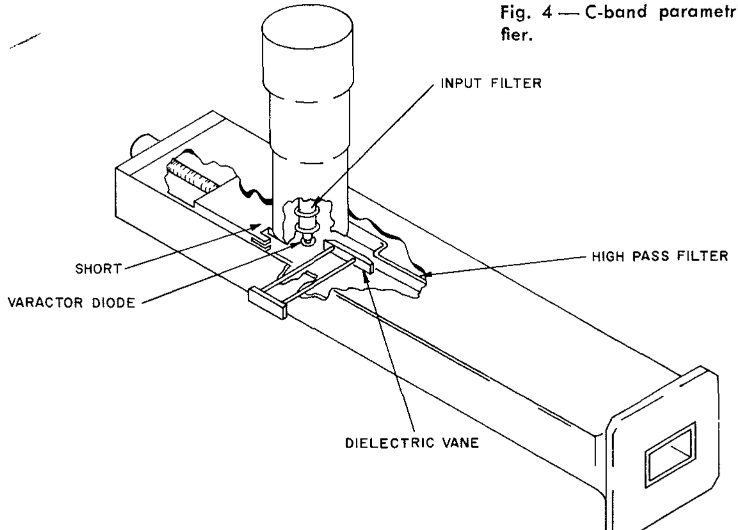
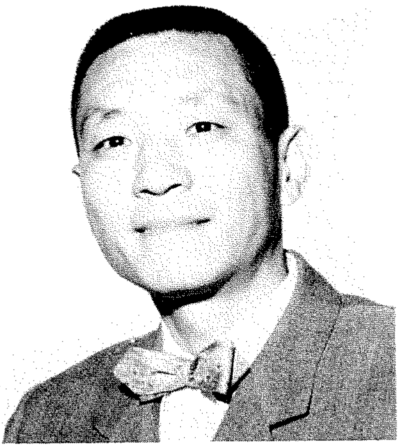


Fig. 4 — C-band parametric amplifier.



H. B. YIN graduated from the University of Shanghai, Shanghai, China in 1943 and received his MSEE from the Syracuse University in 1950. He joined RCA in 1951 with the Broadcast Equipment Section where he designed and developed UHF filterplexers and was responsible for the impedance measurements of UHF antennas. In 1954 he transferred to the Advanced Development Section of the Home Instruments Division where he worked with RF and IF circuits including mixers and oscillators, using vacuum tubes, transistors and tunnel diodes. Specifically, he was responsible for the development of magnetic-tuned TV tuner and transistorized tuner; he also studied the interference problem in a TV receiver. In September 1960, Mr. Yin transferred to the Missile & Surface Radar Division where he has been concerned with the application of solid state devices to various search and monopulse tracking radars. He was in charge of design and development of parametric amplifiers for the DYNASOAR project and MIPIR radar, and travelling wave maser at S-band using rutile as the active material and meander line as the slow wave structure. He has been giving technical consultation to the development of microwave diode switches, tunnel diode amplifiers, and various groups concerning the use of solid state devices. Presently, he is engaging in development of the receiver module for phased arrays.

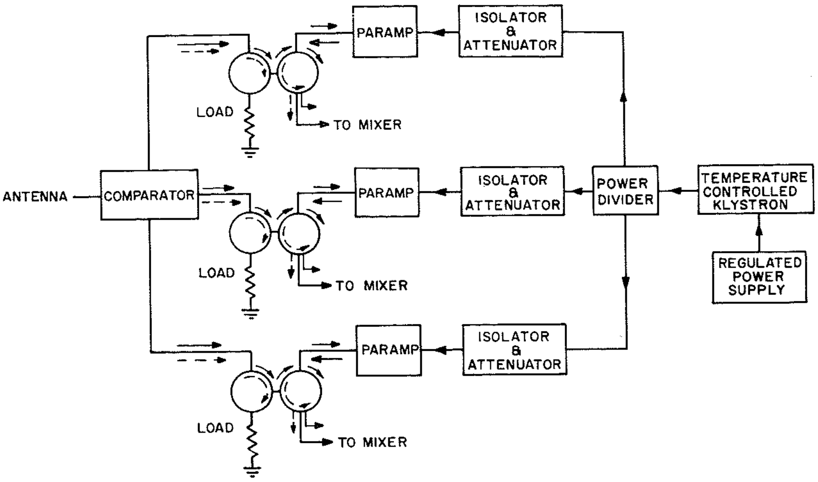


Fig. 5 — Three channel parametric amplifier system for MIPIR monopulse radar.

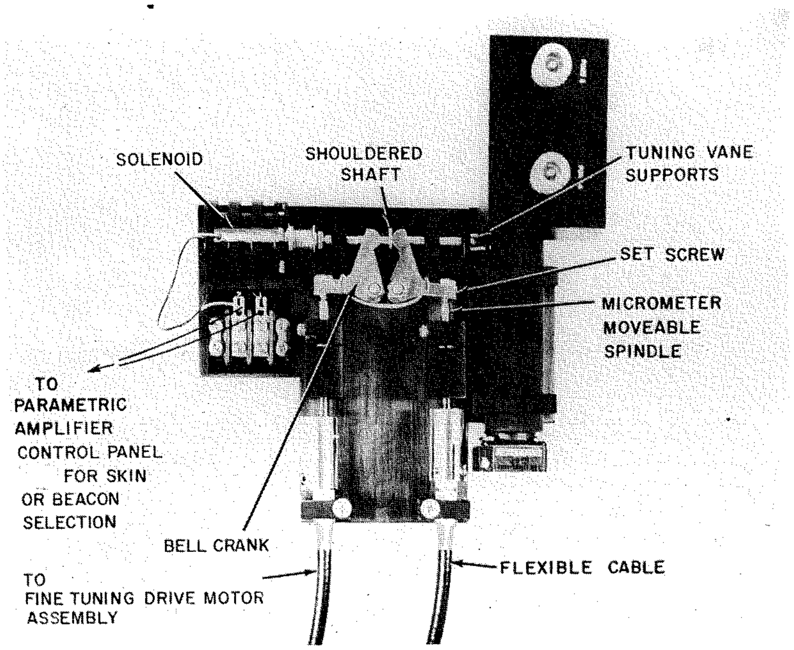


Fig. 6 — Tuning element positioning mechanism.

the position of the micrometer movable spindle. A change in the fixed limits of travel of the shouldered shaft connected to the tuning element determines a new insertion depth of the tuning element in the idler cavity and a new center frequency for parametric amplifier skin or beacon response.

Once these frequencies are preset, the electromechanical mechanism also provides remote switching capability on electrical command from the console. Fine tuning of the resonant cavity above and below the preset skin and beacon center frequency is also available to the operator at the console.

The pretuning establishes for skin and beacon frequencies a fixed relative position which must correspond to the true relative locations of skin-return center frequency and beacon-return center frequency. Normally beacon frequency will be higher than skin frequency and control command from the console will call for the higher frequency preset condition of the amplifier for beacon tracking.

Conditions may exist for specific missions, however, in which the center-frequency relative positions are transposed; skin frequency is then above beacon center frequency. When this condition occurs, the command signal from the console must provide for switching to the lower tuned frequency for amplification of beacon return, and to the higher frequency for skin return.

Fig. 7 shows the parametric amplifier system assembly, which contains the three parametric amplifiers; the drive motors, gearing and mechanical linkage to permit remote tuning of the amplifiers; the pump klystron assembly; waveguide connecting the pump to the amplifiers, coaxial cable providing connections into and out of the circulators; and necessary power and control cabling and terminal boards. Every component is solidly bracketed and every tuning adjustment is supplied with a firm locking mechanism because the system is installed in the RF head of MIPIR radar and is subject to severe shock.

PERFORMANCE

The gain of the amplifiers in the operating frequency band is shown in Fig. 8.

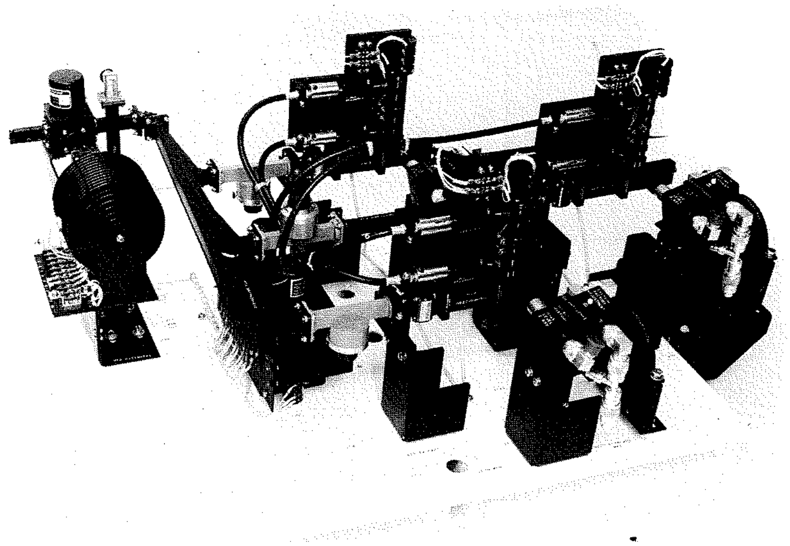


Fig. 7 — Parametric amplifier system assembly.

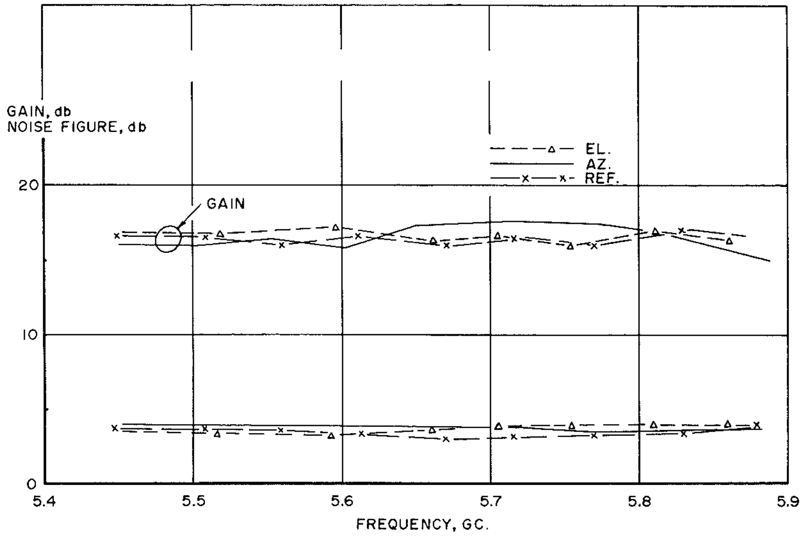


Fig. 8—Gain and noise figures of MIPIR paramps.

The gain is essentially flat and the noise figure is less than 4 db. This includes the insertion loss of the circulator and degradation of the mixer. The frequency tracking among three channels is within ±1.0 Mc. Table I summarizes the performance characteristics of the amplifiers.

The diode used is of the GaAs type, MS4106 (Micro-State), and has a cut-off frequency of 200 Gc at −2 volts. The temperature of the klystron was regulated to ±4°F by the use of a finned heat sink in which a thermoswitch and a heated were imbedded. A frequency stability of ±500 kc with an amplitude stability of ±0.1 db was achieved.

With careful design of the electromechanical system, three amplifiers are frequency tracked within ±1 Mc by a common motor driven idler tuning through flexible cables.

The system noise temperature improvement of the radar with the parametric amplifiers installed can be demonstrated. The system noise temperature can be stated as:

$$T_S = T_A + (L - 1)\,T_P + L\,T_R \qquad (3)$$

Where: T_A = antenna noise temperature, L = insertion loss between the antenna and the receiver, T_P = physical temperature of the losses, and T_R = receiver noise temperature.

Substituting typical values of $T_A = 30°K$ for antenna elevation angle greater than 5°, $L = 1.2$ db, $T_P = 290°$, $T_R = 438°K$, then $T_S = 700°K$. This can be compared with the radar system without parametric amplifiers where the noise figure of mixer paramplifier = 8 db and $L = 1.2$ db, and $T_S = 2{,}150°K$. *The improvement in system noise temperature is about 3:1.*

CONCLUSION

Improvement of radar tracking sensitivity with a set of parametric amplifiers has been achieved. A unique feature of a single-knob-tuned amplifier system has been developed. This system is mounted on a separate chassis which can be installed into the MIPIR (AN/FPQ-6) radar with ease. In case of failure, such as diode burn-out, pump, etc., the amplifier can be bypassed and the radar will remain operational at reduced sensitivity. The remote-tuning capability is provided to readily allow changes of tracking assignments between beacon and skin as well as to provide fine tuning to compensate for transmitter or beacon drift. All controls are accessible to the operator at the console. The amplifier assembly has been installed in several radars and tests are in process.

Although early indications show excellent performance of the paramps themselves, system operation requires external fixed phase compensation between channels. When switching of the paramps in and out is required during tracking, care must be taken that both modes are properly compensated.

ACKNOWLEDGEMENT

The author would like to thank W. Grump and J. Naglee for their assistance in performing the experiments. Special acknowledgement is given to G. VerWys who did perform most of the initial work on this project.

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TABLE 1—Parametric Amplifier Performance

Frequency Range (Tunable)	5.4 to 5.9 Gc
Pump Frequency	16 Gc
Noise Figure	less than 4 db
Bandwidth	larger than 15 Mc
Gain	16 db (nominal)
Diff. G Stability	0.5 db long term
Diff. Phase Stability	±3° long term

FOUR-DIGIT DIFFERENTIAL PCM ANALOG-DIGITAL CONVERTER FOR VOICE APPLICATION

The four-digit, log-compressed, differential PCM system with low-frequency pre-emphasis-de-emphasis gives near "hi-fi" speech quality. Its novel method of frame synchronization on 'flats" (pauses) does not require an additional sync bit. The overall audio harmonic distortion for the whole conversion (encoder and decoder) process is an average of 5% (26 db) in the frequency band from 700 to 800 cps. This overall audio harmonic distortion decreases to approximately 2% (34 db) at lower frequencies like 70 cps to 100 cps. The latter feature makes the differential PCM system very suitable for driving a vocoder.

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The best known method of coding a signal into binary form is pulse code modulation (PCM), which quantizes the absolute magnitude of the continuous analog signal. The binary stream of the PCM can be sent from one repeater (shaper) to another, as long as the sum of all noise sources does not exceed half of the pulse amplitude. The binary coding also makes it possible to encrypt the digital stream for secrecy or privacy.

There is, however, a price tag on PCM. For good speech quality, at least a 6-bit PCM system is necessary. Such a conversion requires expensive circuitry, adds quantizing noise, and increases the bandwidth for transmission. These considerations turned our attention to a related conversion system, *delta modulation.*

A special type of delta modulation is the single-digit delta system. Its hardware- and bandwidth-saving features, however, are counteracted by its signal-to-noise ratio, which is generally poor. These considerations lead to a hybrid system, *differential* PCM, or as described by van de Weg[1], *n-digit delta modulation.* It retains the basic feature of a single digital delta system: it overloads whenever the slope of the analog signal gets too large. This feature makes it very suitable for a speech spectrum which drops 6 db per octave above 800 cps. The increase from ±1 digit to $\pm n$ digits also improves the signal-to-noise ratio. This, in turn, decreases the savings in hardware. A 4-digit differential PCM system is approximately equivalent to a 6-digit straightforward PCM system. As will be shown later, its typical features are the integrated approximation signal F (step waveform) and the subtraction of F from the continuous analog signal S in an analog feedback loop as part of the encoder. The transmitted differential, log-compressed PCM stream is converted (expanded), continuously added up in an identical integrator, and the approximation signal (step waveform) is smoothed in a bandpass filter.

The author participated in the design and development of such a system and conducted the evaluation of an engineering model.

DELTA MODULATION

The delta modulation technique, invented in Europe, uses a one-digit code only. This single-digit delta is a special case of multiple-digit delta modulation. The latter is also called *differential* PCM. (The title *differential* PCM stems from quantization of a difference signal.) It can be seen in Fig. 1 that the audio signal S is subtracted from the approximation signal F, with F containing all previous integrated Q-values. The differential PCM system is therefore a quantization in time and amplitude.

In order to assure good quality for weak speech signals, the quantization is done on a logarmithic scale. Two actions take place during each sampling time at the encoder.

1) *Quantizer Loop Action*: The comparator compares the difference sample against ground and thus determines the most significant bit, the sign bit 2^3. During this time the analog feedback loop of the encoder is open. Three digit-at-a-time trials determine then the quantized value at point Q which defines best the analog difference sample at point A (Fig. 1).
2) *Analog Feedback Loop*: The input to the operational type integrator of the encoder is now closed, and the input to the comparator, point B, is now open. The final quantized value derived under above is added to the previous values in the integrator thus forming an approximation signal F. The subtractor amplifies the difference between F and S and the sample-and-hold circuit presents a sample of the difference signal to point A of the comparator. Since the integrator output will ride on a DC equilibrium it is necessary to eliminate DC through an RC network of sufficient time constant to pass low-frequency loop-correction frequencies.

In order to understand a differential PCM system fully, it is necessary to take a short look at some delta-modulation-technique expressions — such as *dynamic range, digital-analog overload, overload distortion, quantization distortion,* etc.

Dynamic Range

Slope of Sine Wave. A delta system is quantization in time and amplitude, that is, a slope system. Considering Fig. 2, we can calculate the magnitude of a difference sample (increment between two sampling periods. The derivative of a sine wave $V(t) = A \sin 2\pi ft$ is:

$$\frac{dV(t)}{dt} = A\ 2\pi f \cos 2\pi ft. \qquad (1)$$

At the zero crossing, at $t = 0$ we get the maximum possible slope S:

$$S = \frac{dV(t)}{dt} = A \cdot 2\pi f \frac{\text{Volt}}{\text{sec}} \qquad (2)$$

Eq. 2 shows that S is linearly proportional to the sine wave peak value A and

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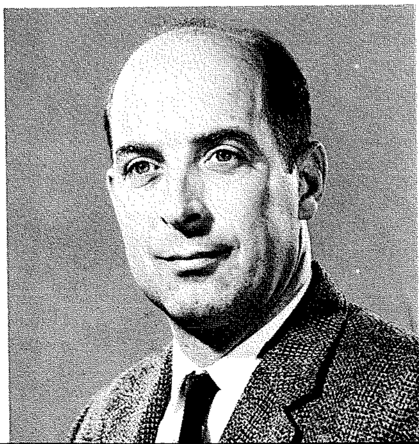
EUGENE KING received a Hauptdiploma in Electrical Engineering (MSEE) from Technische Hochschule (technical university) of Munich, Bavaria, West Germany in 1950. From 1950 to 1953 he was a design and development engineer for AM and FM home, portable, and car radios. In 1953 he joined RCA Victor, Ltd. Division in Montreal as electrical engineer responsible for design and development of the HF, frequency-linear permeability tuning unit of communications receiver CR-188. In 1957 he transferred to the Surface Communications Division. (now Communications Systems Division) working on the circuit and subsystem design on transistorized PPM Time Division, Multiplex Terminal AN/PCC-27 equipment. In 1958 he became Project Engineer for the TRADER V-1 program, responsible for technical direction and coordination of design and development of preliminary development models; and in 1960 he became Project Leader for TRADER V-2 responsible for technical and administrative control of the program. From 1961 to 1963 he worked on classified proposals involving cryptographic analog techniques and as Project Engineer for the four digit, differential PCM, Voice A/D Converter Unicom. His current assignments include design of special circuits for VOCOM and an AR & D project involving new voice PCM concepts using integrated circuitry.



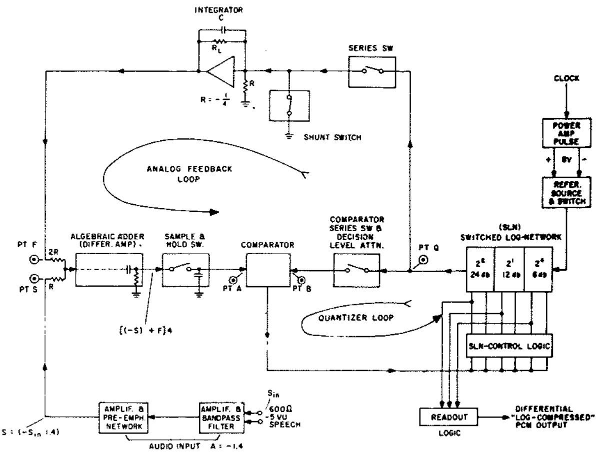


Fig. 1—Four-digit differential PCM (digit-at-a-time) encoder.

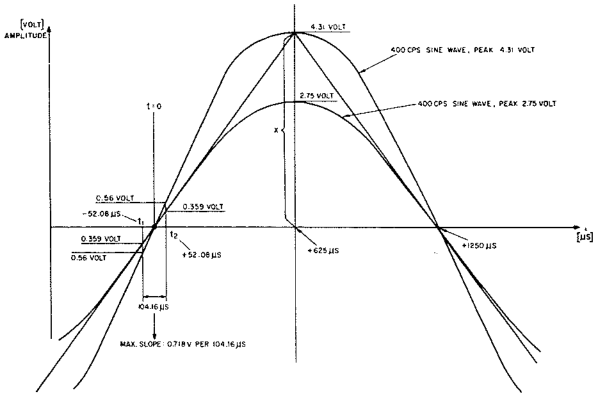


Fig. 2—Digital and analog overload, voice analog-digital converter.

frequency. A sine wave of $f = 400$ cps and a 2.75-volt peak represents, therefore, a maximum possible slope of 0.718 volt per sampling time.

Slope of a Single or Multiple Digit Delta System: A single-digit delta system gives us only two choices, one quantized level σ (volt) up, and one quantized level σ (volt) down at each sampling time: $2^n \cdot \sigma = 2\sigma$ (single-digit delta, $n = 1$ digit). A multiple digit delta system of $n = 4$ digits gives 16 choices at each sampling time:

$$\frac{(2^n - 1)\,\sigma}{T_s} = (2^n - 1)\,\sigma f_s\,\frac{\text{Volt}}{\text{sec}} \qquad (3)$$

with $f_s = \frac{1}{T_s}$, and f_s = sampling rate

Eqs. 2 and 3 have to be equal, because the slope of a sine wave can never exceed the slope capability of a delta system, or overloading and hence distortion will occur:

$$A\,2\pi f = (2^n - 1)\,\sigma f_s \qquad (4)$$

Digital and Analog Overload

Digital Overload: Now we can specify the dynamic range at which this overload distortion should occur. The point to choose is the gravity point of an average speech spectrum which is approximately at $f = 400$ cps.

A system with a chosen highest quantized value of ±8 volt must hit this overload point at a maximum sine wave input of $f = 400$ cps and a 2.75-volt peak which corresponds to a maximum possible increment of 0.718 volt per sampling time. This in turn determines the gain of the individual loop circuits and the audio input circuitry.

$$\left(\frac{\textit{Highest quantized value}}{\textit{Max. possible increment}}\right)$$

$$= \textit{total gain from audio input to comparator pt. A}$$

Analog Overload: A sine wave of $f = 400$ cps and 2.75-volt peak produces only at zero crossing digital overload (level ±8).

In order to insure that encoder and decoder circuits will not saturate (hang up), their dynamic ranges have to be calculated for a succession of maximum increments. Looking at Fig. 2 we can see that extending the tangent to the maximum slope results in a peak value of 4.31 volts, which is a triangle waveform of $f = 400$ cps. A symmetrical triangle waveform will therefore result in an alternating Q-value sequence of 12 times +8 volt followed by 12 times −8 volt. Its buildup at the integrators determines the maximum dynamic ranges necessary for the converter circuits.

Quantization introduces an unwanted property, the increase of quantization distortion with decreasing input signal amplitude. In order to minimize this affect at low amplitudes, it is necessary to introduce logarithmic quantization. Logarithmic quantization is logarithmic compression followed by linear quantization (transmitter) and linear quantization followed by logarithmic expansion (receiver).

Comparing the curves of Fig. 3, we see that for an amplitude swing of 5% we get $k = 20\%$ for logarithmic quantization. In order to get approximately the same distortion factor, $k = 20\%$ for linear quantization, we can only allow an amplitude swing of approximately 25%. This means our dynamic range improved by a factor of approximately 5.

Looking more closely at Fig. 3c, we also find that quantization on a logarithmic scale gives a definite improve-

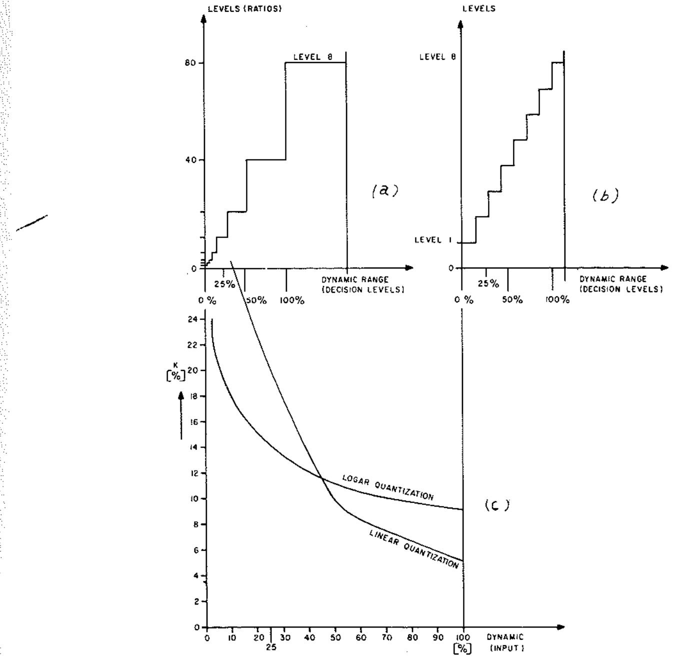


Fig. 3—Quantization-distortion for linear and logarithmic quantization.

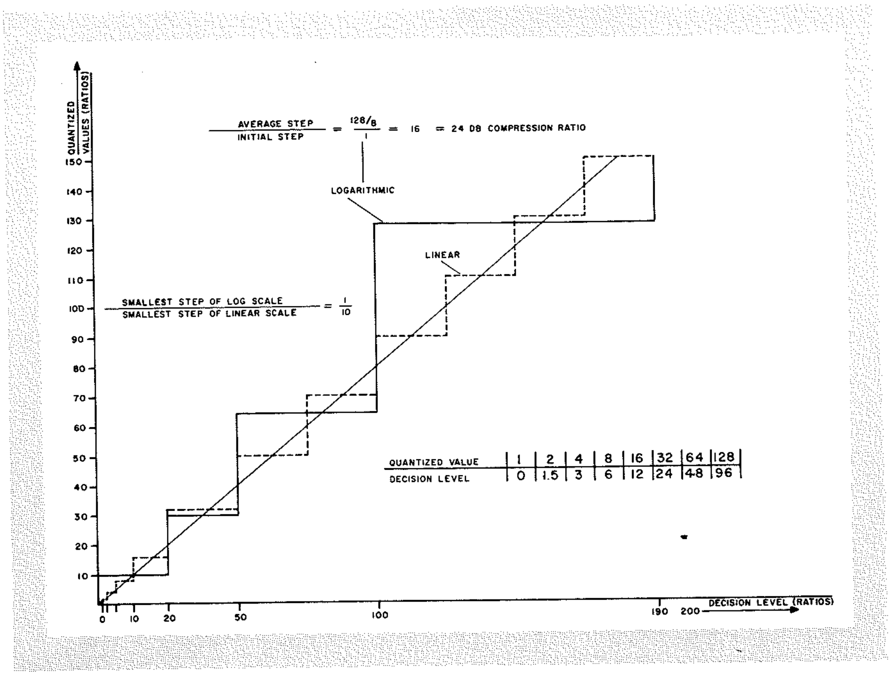
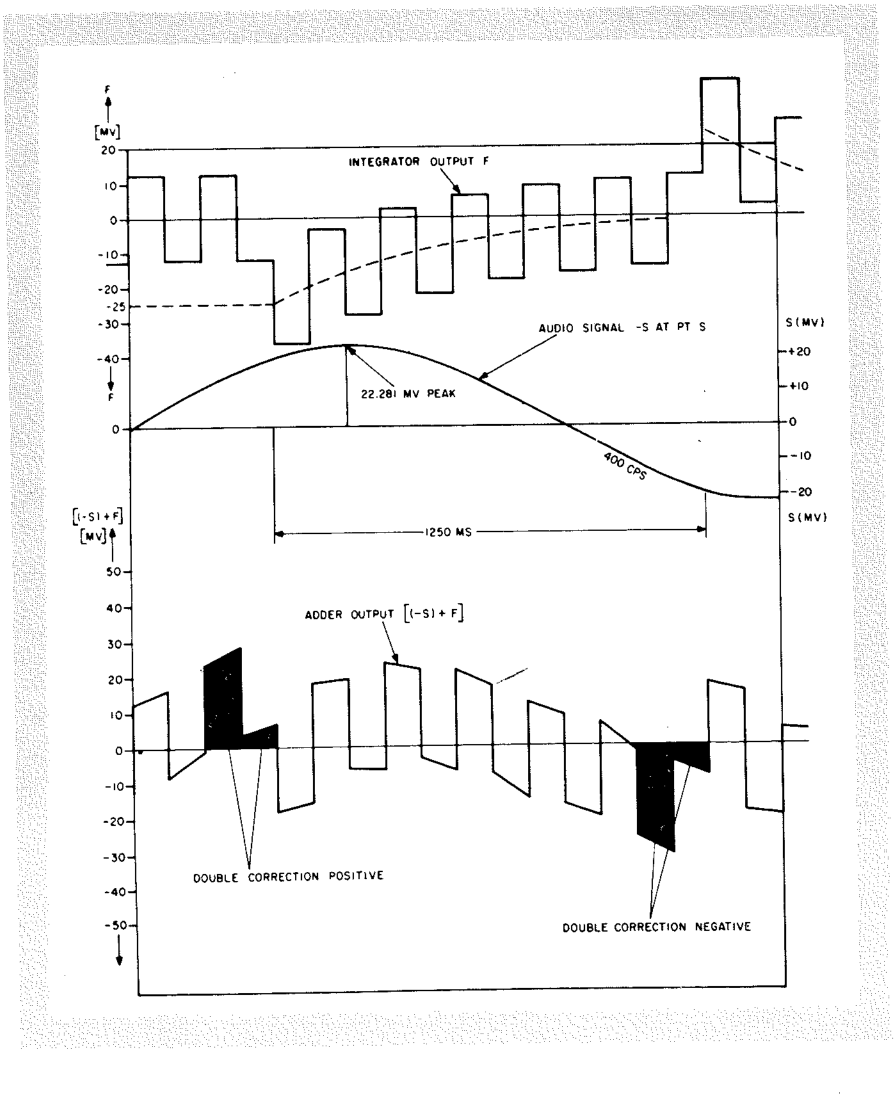


Fig. 4—Quantized values versus decision levels (24-db compression).

Fig. 5—Audio signal behavior.



ment of 4 to 5 db in signal to quantizing noise ratio as compared to linear quantization.

For example, at a dynamic swing of 20%, logarithmic quantization (with 20-db compression ratio) results in a distortion factor of $k = 15\%$, whereas linear quantization results in a distortion factor of $k = 23\%$. This is a signal-to-quantizing noise improvement of 1.54:1 or 4 db in favor of logarithmic quantization.

At a dynamic swing of 15% the improvement is approximately 4.5 db. Above a dynamic range of approximately 40% the logarithmic quantization distortion is somewhat higher than by linear quantization, but of low significance.

A typical logarithmic quantizer scale is shown in Fig. 4. Two different "staircases" can be seen, one for linear quantization without compression and one for logarithmic quantization with a compression ratio of 24 db. The vertical risers of the staircase are called decision levels. The steps of the staircase are called quantized value.

Audio Signal Behavior

Fig. 5 shows small audio signals causing still Q-level $= 100$ mv oscillation but with alternating ± double correction (no excitation of Q-level ± 2). The integrator output F has an exponential droop from ±25 mv to its normal equilibrium for zero-input oscillation.

Error Signal

Fig. 6 shows a typical error signal for repetitive sine wave of 400 cps and 2.5-volt peak input into 600-ohm encoder input.

Overall Audio Harmonic Distortion of a Differential PCM Encoder-Decoder

The main distortion contribution stems from the quantization process. A vital part is played by the matching of the decoder-expansion against the encoder-compression curve. Another source of distortion is the audio input circuitry of the encoder and the audio output circuitry of the decoder.

Fig. 7 shows the overall audio harmonic distortion versus frequency at a constant input of 1/10 full sinusoidal load (20 db = 1/10 referenced to digital overload at 400 cps). The curves of Fig. 7 are actually measured figures of an engineering breadboard model. The curves follow approximately a 6-db/octave slope which is similar to the slope of a speech spectrum.

The approximately true quantization distortion of the encoder-decoder can be obtained by subtracting the distortion of the audio input-output circuitries, and audiogenerator.

The 25-db reference is equal to 5.6% distortion.

Decoder

The differential log-compressed PCM stream does not carry any extra frame sync bit. With zero signal input or nearly no incremental change of the signal amplitude during sampling times, the encoder feedback loop oscillates in a very stable manner of alternating quantized level ±1 which gives in the decoder an *out-of-band* frequency of 4,800 cps.

Frame synchronization on "preponderance of flats" in speech is a recent Bell Telephone Laboratories invention. The binary code associated with each quantized value has been chosen so that in an out-of-sync case of one, two, or three bits, the voltage on the decoder frame-sync RC-integrator will build up fast enough in order to advance the read in counter by one bit within approximately 20 to 300 msec.

This framing method has been proven with speech in the differential PCM analog-digital UNICOM.

The switch-log-network of the decoder is identical and matched within approximately 5% to the encoder network. The decoder integrator is similar to the encoder integrator except its charging and leakage time constant and gain. The framing circuitry consists of two Schmitt triggers, one shunt-relay, one RC integrator, associated logic elements, and audio output circuitry.

SUMMARY

The four-digit, log-compressed, differential PCM system with low-frequency preemphasis-de-emphasis gives near "hi-fi" speech quality. Its novel method of frame synchronization on "flats" (pauses) does not require an additional sync bit. The overall audio harmonic distortion for the whole conversion (encoder *and* decoder) process is an average of 5% (26 db) in the frequency band from 700 to 800 cps. This overall audio harmonic distortion decreases to approximately 2% (34 db) at lower frequencies like f = 70 cps to 100 cps. The latter feature makes the differential PCM system very suitable for driving a vocoder.

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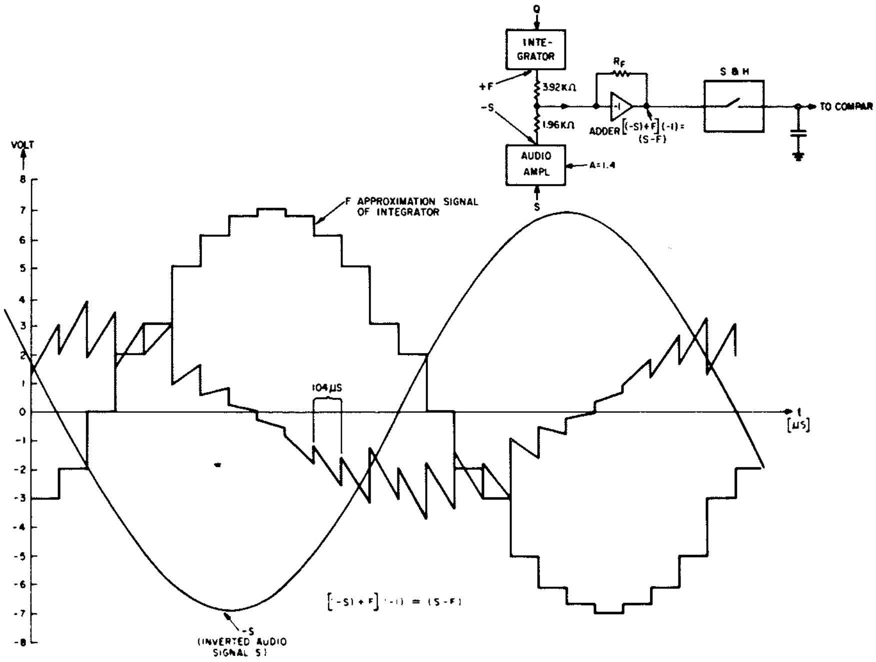
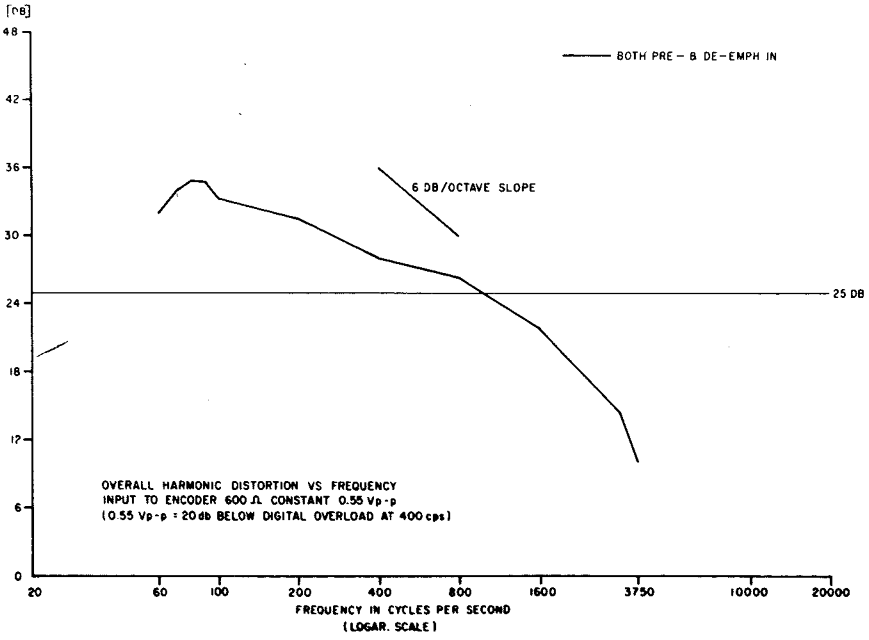


Fig. 6—Typical error signal for repetitive sine wave of 400 cps and 2.5-volt peak input into 600-ohm encoder input.

Fig. 7—Overall harmonic distortion vs. frequence input to encoder at constant input of 1/10 full sinusoidal load. (20 db = 1/10 referenced to digital overload at 400 cps).



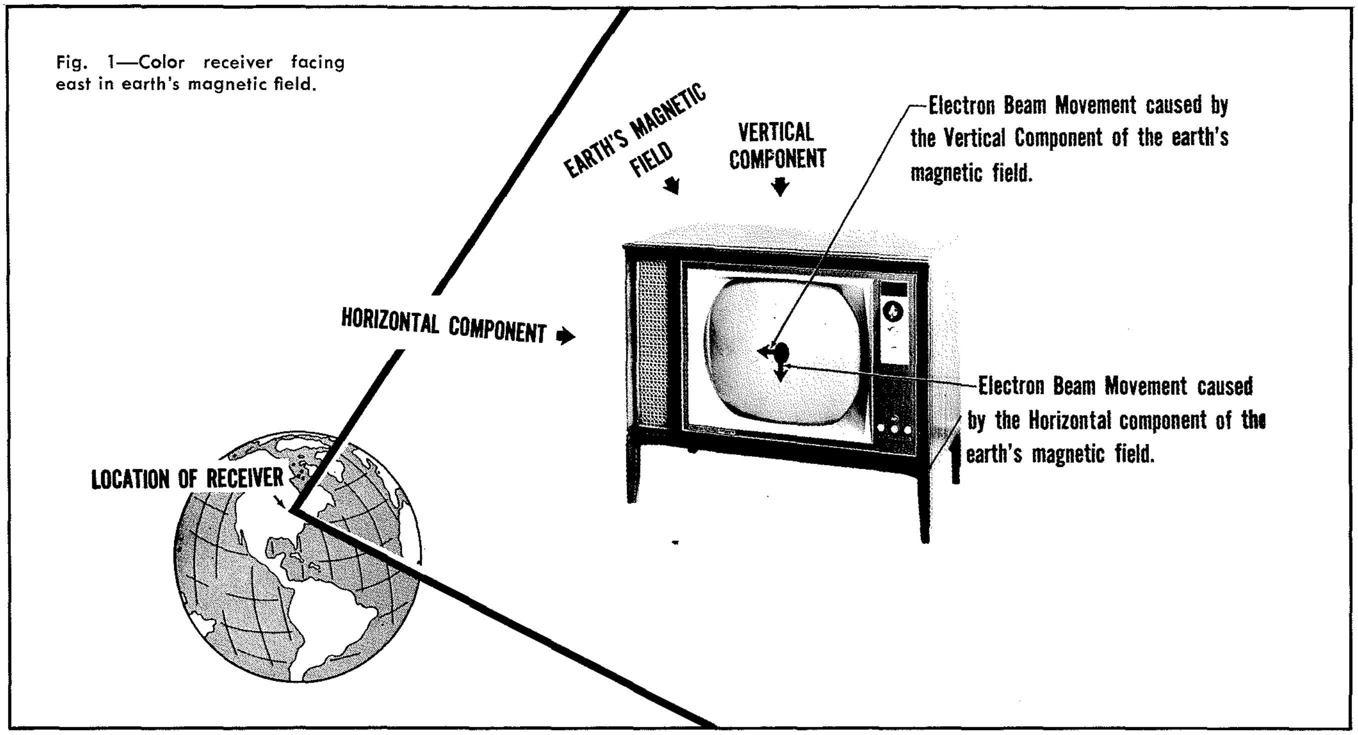


Fig. 1—Color receiver facing east in earth's magnetic field.

AUTOMATIC DEGAUSSING FOR COLOR TV RECEIVERS

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A characteristic of the three-gun shadowmask color-television tube is its high sensitivity to magnetic flux fields; even comparatively small magnetic fields produced by the earth can cause visible color errors. So that the color picture tube can be properly aligned and oriented in any direction in the earth's magnetic field, some form of magnetic shield is needed. Before a practical magnetic shield can perform its shielding function properly, the shield must be degaussed in the presence of the specific magnetic field to be rejected. Degaussing enables the metal shield to "forget" its previous magnetic orientation and realign magnetically, thus counteracting any new magnetic field condition. Degaussing affects the metal in the picture tube's shadowmask-frame assembly in the same manner. Previously, the color television receiver was manually degaussed by driving a solenoid-wound coil with the 120-volt AC line voltage. The coil would be moved around the front of the tube and then withdrawn slowly. This operation was required every time the position of the color receiver was changed with respect to the earth's magnetic fields. Automatic degaussing is designed to give the color television receiver freedom of movement in the earth's magnetic field without using manual degaussing.

To illustrate the magnetic flux problem, consider the color television receiver in the earth's magnetic field (Fig. 1). Any component of the earth's field which crosses the electron beam path perpendicularly, as the electrons travel from the tube electron guns to the phosphors on the picture tube's screen, will cause additional undesired deflection of the beams; such a condition can cause the electrons to illuminate phosphors of the improper color. The magnitude and direction of the added deflection will be dependent upon both the geographical area and the orientation of the receiver with respect to the earth's magnetic field in that geographical area.

The earth's magnetic field can be represented as magnetic lines of force between the south and north poles directed toward the north pole. This magnetic field can be considered to have vertical and horizontal components at any given point on the earth's surface. If a color receiver is moved from the north pole to the south pole it would experience a change in the magnitude of the vertical component of the earth's field from a maximum value (directed toward the earth at the north pole) to a maximum value directly away from the earth at the south pole. The point of zero vertical flux would occur at the equator. The magnitude of the horizontal component of the earth's field would change from a zero value at the two poles to a maximum value at the equator.

If a color receiver is placed in a given geographical area, the vertical component of the earth's field will give a constant added deflection causing the electron beams in the picture tube to be displaced horizontally from a zero magnetic field condition. Rotating the receiver to face any particular direction will not cause a change in beam deflection due to this vertical magnetic field component. However, the horizontal component of the earth's magnetic field will cause a change in the beam deflection when the color receiver is rotated. The rotation will vary the angle of intersection between the horizontal magnetic field component and the electron beam path in the picture tube.

As in any commercial receiver product design, it is necessary to make a "cost-versus-desired-performance" decision; it was decided that automatic de-

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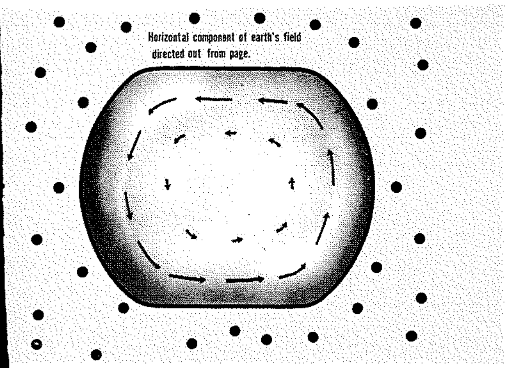

Fig. 2a—Relative electron beam movement caused by the horizontal component of earth's magnetic field when receiver is facing north. (Horizontal component of earth's field directed out from page.)

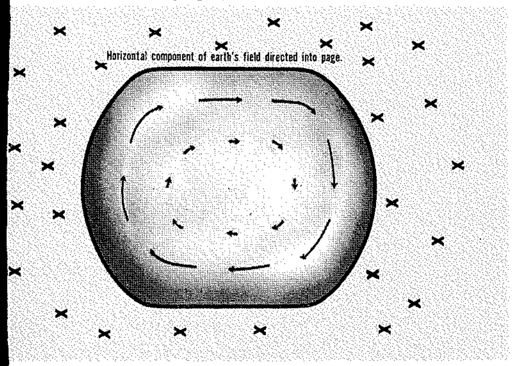

Fig. 2b—Relative electron beam movement caused by the horizontal component of earth's magnetic field when receiver is facing south. (Horizontal component of earth's field directed into page.)

gaussing must enable the color receiver to have freedom of movement in any geographical area where the variations in the earth's magnetic field are not greater than the variations encountered in the continental United States. The color receiver can be aligned to correct any color errors caused by the earth's magnetic field; geographical limitations only affect the amount of variations in the earth's magnetic field that the receiver can withstand without realignment. This same geographical limitation is true for previous color receivers using manual degaussing. Limiting the geographical area greatly reduces the beam movement caused by variations of the vertical component of the earth's field. The major cause of the beam movement then becomes the orientation of the set with respect to the horizontal component of the earth's field.

When the color receiver faces north or south, the horizontal component of the earth's field affects the electron beam due to the radial component of the 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 tube with no beam movement at the center of the tube where the electron beams are parallel with the horizontal components of the earth's magnetic field (Figs. 2a and 2b). Due to built-in tolerances in the tube these tangential movements in the RCA 21FBP22, 70° color tube caused by a north or south orientation are not great enough to cause serious visible color errors; such color errors must be considered, but only a limited amount of correction is necessary.

When the color receiver faces east or west, the horizontal component of the earth's field acts over the horizontal component of the distance that the electron beams travel (from the electron guns to the phosphors on the tube's screen. With the receiver facing east, the beams are deflected vertically downward as compared to a zero field condition. To face the receiver west, reverses the direction of travel of the electron beam in the earth's field, thus causing the electron beams to be deflected vertically upwards as compared to a zero field condition (Figs. 3a and 3b). Such vertical beam movements are great enough to cause serious visible color errors.

The above conditions place the following design requirements on the shield and degaussing system for the 21FBP22, 70° color picture tube:

Shield

1) After degaussing, the shield must correct for the horizontal movement of the electron beams caused by the vertical component of the earth's field.
2) After degaussing, the shield must correct for vertical and tangential movement of the electron beams caused by the horizontal component of the earth's field.

Automatic Degaussing

1) After initial factory alignment, automatic degaussing must magnetically realign the shield and shadowmask-frame assembly of the tube when the receiver is oriented in any direction with respect to the horizontal component of the earth's magnetic field.

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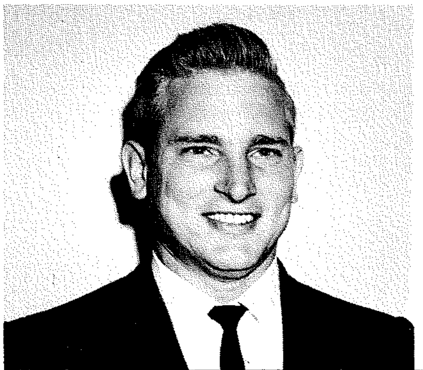


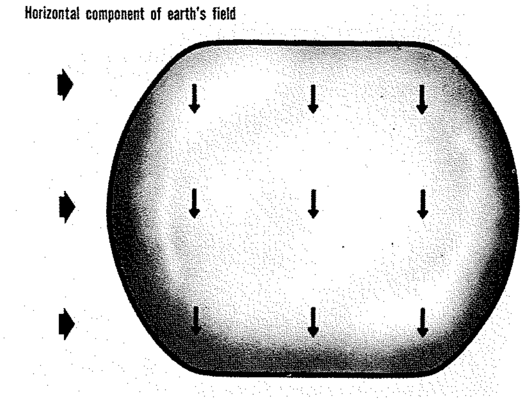


Fig. 3a—Relative electron beam movement caused by the horizontal component of earth's magnetic field when receiver is facing east.

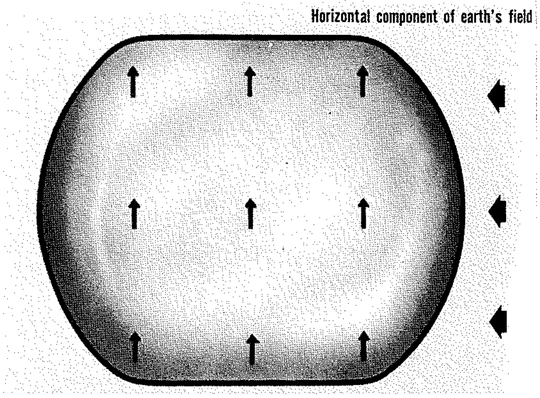


Fig. 3b—Relative electron beam movement caused by the horizontal component of earth's magnetic field when receiver is facing west.

2) Automatic degaussing must magnetically realign the shield and shadowmask-frame assembly of the tube to compensate for the changes in the magnitude of the vertical component of the earth's field that occur throughout the continental United States. The main magnetic alignment of the shield and shadowmask-frame assembly for the vertical component of the earth's field can be done at the time of initial factory alignment. Automatic degaussing need only correct for vertical magnetic field variations from the initial aligned condition.

MAGNETIC SHIELD DESIGN

When a ferromagnetic material is placed in a magnetic field, the resultant field consists of the original field and the field created by the orientation of the magnetic domains in the ferromagnetic material. To perform as a magnetic shield, the magnetic domains in the metal must be orientated by the external field so that a field is always created in opposition to the external field within the area to be shielded. The induced opposing field must be redirected by the external field when the direction of the external field is changed.

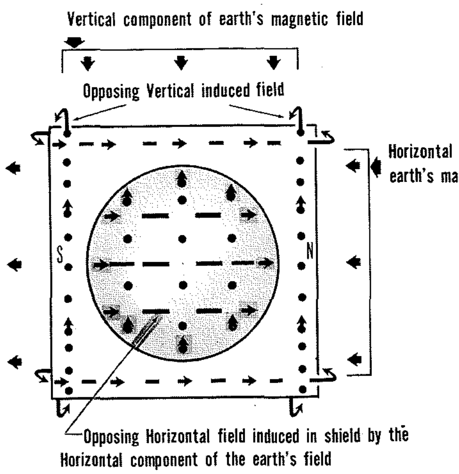


Fig. 4a—Induced field, receiver facing west.

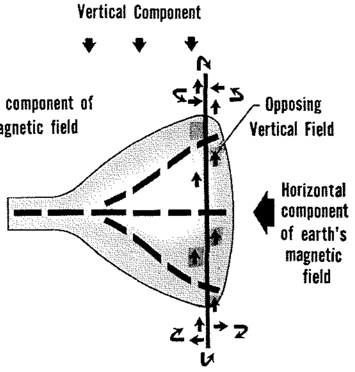


Fig. 4b—Induced field, receiver facing south.

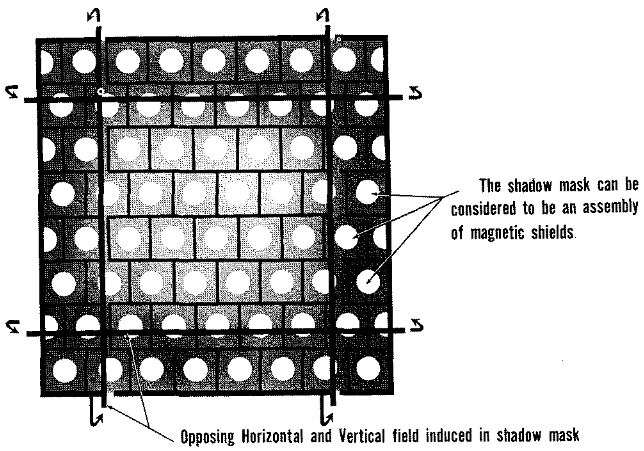


Fig. 4c—Induced field, section of shadow-mask of picture tube, receiver oriented west.

Since the greatest degree of shielding from the horizontal component of the earth's field is required when the color receiver is facing east or west and only a small degree of shielding from this component is required when the receiver is facing north or south, the shield for the 21-inch 70° tube can be of fairly simple construction; a planar steel shield mounted concentric with the tube to the rear of the tube's shadowmask-frame assembly gives the needed shielding. Figs. 4a and 4b show the opposing fields induced in the shield by the earth's magnetic field for the set facing west and south. When the receiver faces east instead of west, or north instead of south, the opposing flux field pattern induced in the shield by the earth's magnetic field would be the same except that the horizontal component would be reversed.

The metal used for magnetic shielding should have magnetic domain boundaries which can be moved easily by a small magnetizing force, namely the earth's magnetic field. This property enables the shield to create an opposing flux field that magnetically realigns whenever the external magnetic conditions are changed. Such metals are called "magnetically soft." Practical considerations indicate that soft steel should be used in the shield; this metal has a limited ability to realign magnetically. The shield needs to be degaussed before the metal can "forget" its previous orientation and enable the external field to realign the magnetic domains in the metal to counteract a new magnetic field condition. Therefore, everytime the color receiver is moved geographically or rotated in a given geographic area the magnetic shield of the picture tube has to be degaussed.

Degaussing also enables the earth's field to induce an opposing field in the shadowmask-frame assembly of the picture tube. The shadowmask can be considered an assembly of many small magnetic shields (Fig. 4c). The earth's field induces an opposing field in the shadowmask that aids the field induced in the magnetic shield shown in Figs. 4a and 4b.

DEGAUSSING FLUX PATTERN

Experimental data indicate that degaussing a shield by a flux pattern which is parallel to an external field enables the external field to create the strongest opposing magnetic field in the shield. The color picture tube needs the greatest amount of shielding from the horizontal component of the earth's field when the color receiver is facing east or west. Therefore, when the color receiver is orientated east or west, the degaussing flux pattern should be directed parallel to the horizontal component of the earth's field. This flux pattern is achieved by the combined effect of four coils mounted on the shield and polarized as shown in Fig. 5. The horizontal degaussing flux pattern in Fig. 5 also enables the earth's magnetic field to induce in the shield and shadowmask of the picture tube the limited opposing horizontal flux field needed when the set is orientated north or south.

To give the color picture tube the greatest shielding from the vertical component of the earth's field a vertical degaussing flux pattern should be used. Therefore, at the time of initial factory alignment, the color receiver should be degaussed by a horizontal and vertical degaussing flux pattern. However, after the initial factory alignment, a horizontal degaussing flux pattern can give the needed correction for the changes in the vertical component of the earth's field that occur in the continental United States.

DEGAUSSING CURRENT

To drive the four coils and perform the degaussing operation an AC current which decays to zero is required. To equal the performance of manual degaussing with the above described shield and coil system, the driving current needs a first-cycle-peak amplitude of approximately 4 amperes. This driving current is developed by using a thermistor-varistor switch in the secondary of the receiver's power transformer (Fig. 6). The characteristics of the varistor and thermistor which provide the switching action are shown in Figs. 7a and 7b.

When the receiver is turned on after being off for a few minutes the thermistor is at ambient temperature with a re-

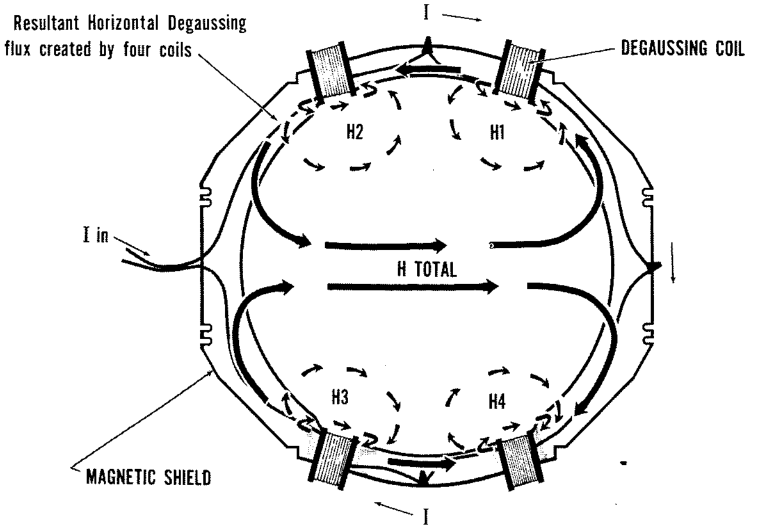
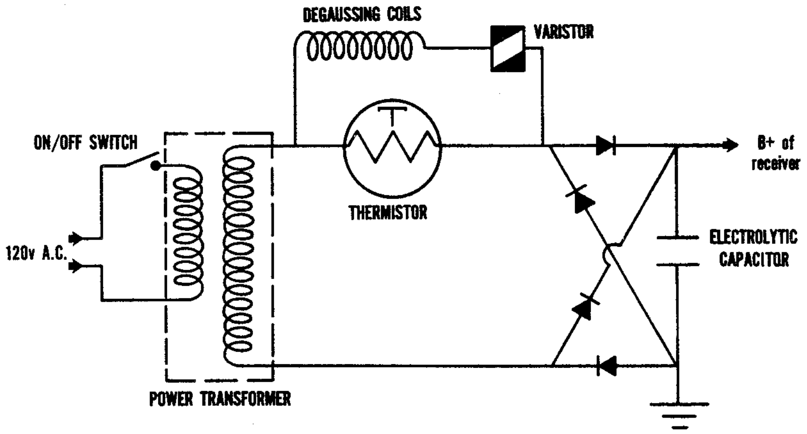


Fig. 5—Degaussing flux pattern.

Fig. 6—Power supply.



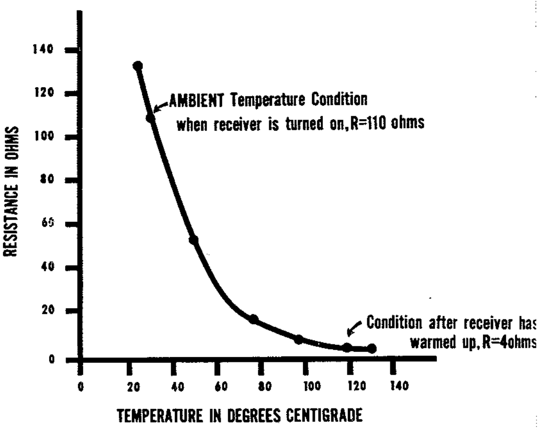


Fig. 7a—Negative resistance versus temperature characteristic of themistor.

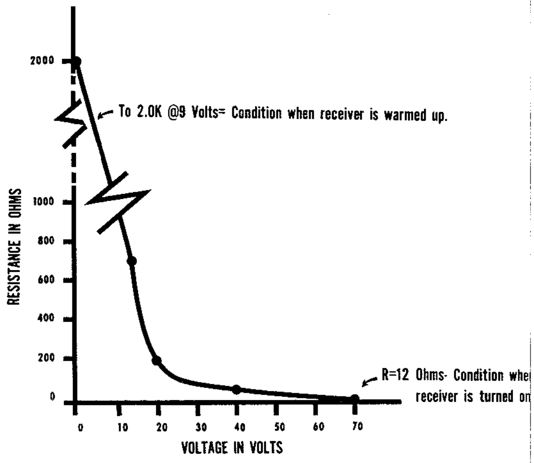


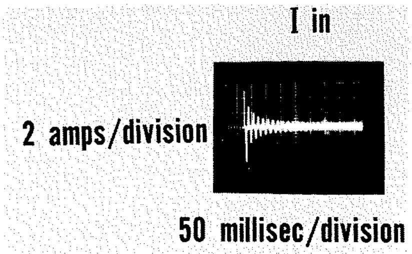
Fig. 7b—Negative resistance versus voltage characteristic of varistor.

sistance value of 110 ohms. The charging current of the electrolytic capacitor creates a 60-volt peak potential across the varistor, giving it a resistance value of 12 ohms. Thus, at the instant the set is turned on, the 110-ohm thermistor is paralleled by the 12-ohm varistor plus the 9-ohm degaussing coils causing the majority of the initial capacitor-charging current to drive the degaussing coils. The remainder of the capacitor-charging current heats the thermistor, decreasing its resistance. The current in the degaussing coils decays to zero due to the charging of the capacitor and the heating of the thermistor which decreases the voltage across the varistor, increasing its resistance. By the time the receiver has warmed up to produce a picture (approximately 35 seconds) the resistance of the thermistor has decreased to 4 ohms while the resistance of the varistor has increased to 2 kilohms. The current through the degaussing coils is effectively cut off, causing all of the receiver's normal operating current to pass through the thermistor. The waveform of the current through the degaussing coils is shown in Fig. 8. Since degaussing is completed within the warm-up time of the receiver, the viewer will see no visible effects caused by the degaussing operation.

SUMMARY

Automatic degaussing of the color receiver gives these three definite advantages: 1) The local dealer is not required to manually degauss the color television receiver before displaying it to his customers. 2) A service call is not required when the color television receiver is relocated in the owner's home. 3) The effect on the color television receiver of stray magnetic fields in the owner's home can be eliminated.

Fig. 8—Current through degaussing coils.



ACKNOWLEDGEMENTS

Credit for the fundamental design of the system belongs to P. G. McCabe, J. Stark, Jr., and J. K. Kratz. The author also wishes to acknowledge the valuable assistance received from G. L. Roth and E. B. Boyer.

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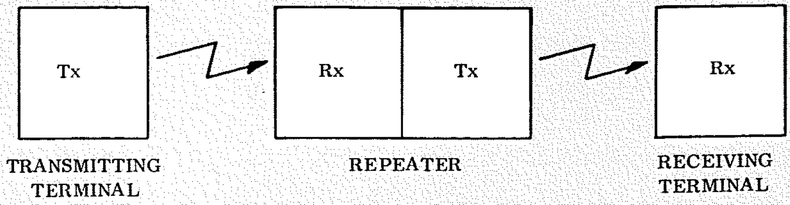


Fig. 1—Microwave system.

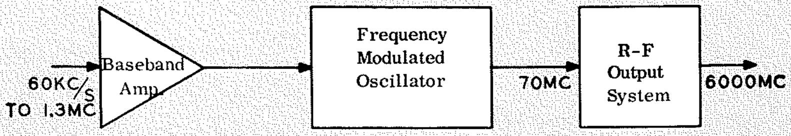
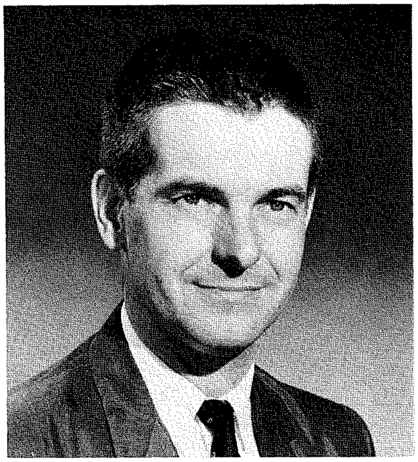


Fig. 2—Transmitting terminal.

E. J. FORBES received his BSEE from the University of Manitoba, Winnipeg, in May 1950. He then joined the Canadian General Electric Co. in Toronto where he worked on the design of various microwave equipments. In 1953, he joined RCA and became engaged in advanced development of microwave communication circuits, 6,000-Mc traveling — wave — tube application in particular. During 1954-56 he was engaged in the study of tropospheric scatter propagation at UHF and SHF. In this program he was responsible for the instrumentation and operation of test links as well as the reduction and primary analysis of the experimental data. Mr. Forbes has been most recently responsible for the design and development of the CW-60 radio equipment with its switching, fault reporting, and service channel facilities. In 1961 he worked on the MM-600-6 equipment. His work included the 6,000 Mc transmitter and determination of specifications for branching networks, waveguide components, and special power supplies.



DESIGN OF CW-60 SOLID-STATE MICROWAVE RELAY EQUIPMENT

In recent years, a new generation of microwave communication equipment has been developed to serve industrial users and common carriers. The industrial users — railroads, airlines, pipelines, and utilities — depend upon microwave equipment for economical and efficient communications in conducting their businesses. The common carriers provide communication services for profit. However, they have common needs — reliability, maintainability, and longevity. To fulfill these needs, RCA has pioneered with the CW-60, a solid-state microwave communication product. This 240-voice-channel, 6,000-Mc unit meets the industry standards for medium capacity, long haul equipment. An additional feature is 120-voice-channel insert capability at through-repeaters.

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THE RCA CW-60 solid state microwave relay equipment was displayed publicly for the first time two years ago at the Pipe Line Industry Electrical Association Conference in Galveston, Texas. In the past two years, the Broadcast and Communications Products Division has refined the equipment, added some interesting auxiliary units, and has successfully installed it in the field.

THE RADIO SYSTEM

Fig. 1 shows a simple microwave communications system; a transmitting and receiving terminal connected by a repeater station. The equipment is of the IF heterodyne type. This is not a new type and a number of technical papers describe it in varying depth.[1,2,3] The IF heterodyne equipment does not demodulate to traffic frequency at repeaters where the station gain is principally at an intermediate frequency. The heterodyne system avoids the recurrent distortions of the remodulation type equipment. This makes it more suitable for long haul application.

The Broadcast Division chose the intermediate frequency (70-Mc) for the CW-60 to gain international acceptance and to be consistent with its other product lines. Fig. 2 illustrates a transmitting terminal consisting of an IF oscillator, frequency-modulated by the sending voice multiplex. The traffic, or "baseband," spectrum, occupied by a 240-voice-channel signal, extends uniformly from about 60 kc to 1.3 Mc. The FM signal is shifted to the microwave channel by up-conversion in a mixer. The skeleton receiving terminal in Fig. 3 is simply a superheterodyne FM receiver, and interconnecting it with the transmitting terminal provides a two-way heterodyne repeater (Fig. 4). Antenna duplexing plus a low-noise tunnel-diode preamplifier are added to enhance the available fade margin.

When the CW-60 was planned, the tunnel-diode amplifier (TDA) was relatively new to the industry. It was selected only after serious consideration of other alternates. Complexity and expense ruled out the equally effective parametric amplifier.

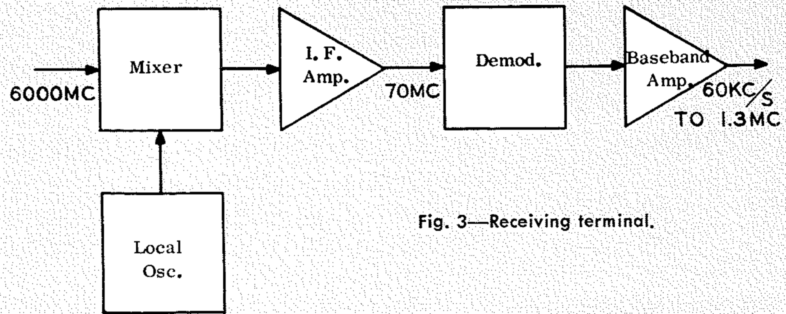

Fig. 3—Receiving terminal.

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A single amplifier is used to drive the two receivers of a hot standby repeater. The effective loss in the equipment noise figure seen at the antenna terminals resulting from the power splitting of the input signal to two receivers is small because of the moderate gain and small noise figure of the TDA. However, because the TDA is a common element to both the main and standby receivers and was new at the time, the diode and its associated circuitry were designed with great care. The construction was planned so that failure of the diode module does not damage the fade margin over the associated path by more than about 20 db. Thus, if the diode fails, communication is maintained. In addition, there is an alarm to indicate such a failure.

An important economic factor in this system is the effective repetitive use of certain units in terminal and repeater arrangements as shown by a comparison of Figs. 4 and 5. The fact that the building blocks of the terminal are common to the repeater is an important plus not only in the area of low initial cost resulting from volume production, but also in the areas of spares and maintenance practice.

The key note of the CW-60 is reliability of service. It is designed to be operated from a 48-volt battery kept charged at a floating rate to reduce the dependence on a continuous commercial mains supply. To provide an adequate length of equipment service on battery power after a mains failure, power consumption (typically 500 watts for a hot standby repeater implemented with alarm and order wire) had to be minimized to keep battery costs reasonable.

The question arises, *"If we are power-conscious and the equipment circuits are as reliable as advertised — why hot standby?"* The answer is that the past decade of the microwave relay business has made the user *demand* standby service. In the early days this was due to the short life expectancy of the klystron and triode. Of late, with the advent of increased system loading, (240 voice channels being commonplace) and with added complexity of maintenance routines, the hot standby feature has an additional purpose—a more economical use of operating personnel.

SWITCHOVER FACILITY

One of the major features of the CW-60 equipment is the solid-state design. No relay or thermionic device is used. Even incandescent indicators are few. Also the number of circuit transfer points has been minimized to decrease the likelihood of a transmission break caused by a circuit failure. In Figs. 4 and 5 only the transmitter output and receiver baseband output are switched. The switching function has been provided through solid state computer logic consisting of an arrangement of comparators, *or* and *and* gates, flip-flops, and diode-switch circuits. The terminal sensing on the transmit side through a unique mixing circuit in the up-converter and FM demodulator detects both continuity pilot and transmitter output variation. Deterioration of either pilot or RF output causes transfer to the standby transmitter. The receive side consists principally of a pair of pilot receivers operating a baseband switch which functions on the deterioration of the pilot.

The repeater switching is divided into two basic areas—the transmitter and the receiver. Because there is no switching at any IF point, repeater performance can be sensed completely by transmitter output detectors. Repeater switching is thus derived from such monitors. Each standby receiver assembly has a common active element—the TDA. A comparator circuit has been included to sense its status. Because this circuit cannot differentiate between short deep propagation fades and a TDA failure, radio alarms are delayed by about half a minute. The system is weighted to the main side, so that common circuit failures or sequential main-standby failures result in the main path being chosen. Standby IF oscillators maintain system continuity if there is a complete receiver failure. The transmitter switch is an electromagnetically polarized assembly of threeport microwave ferrite circulators. This center-stable device and its transistor driving circuit is monitored and alarmed. Without that, an open magnet coil could disable the system and could be located only through a station-by-station tour.

TABLE I—Station Alarms

Terminal	*Repeater*
Rx1 or Rx2 squelch	Tx1 or Tx2 (E-W)
Rx1 pilot	Rx1 or Rx2 squelch (E-W)
Rx2 pilot	loss of received signal (E-W)
loss of received signal	Tx1 or Tx2 (W-E)
Tx1 output or pilot	Rx1 or Rx2 squelch (W-E)
Tx2 output or pilot	loss of received signal (W-E)
24-volt regulator	24-volt regulator
IF gain, manual	IF gain, manual
station control, manual	station control, manual
RF switch driver	RF switch driver

Table I lists the points that have alarm circuits. Because manually overriding an automatic function disables the standby equipment, an alarm is provided to indicate any manual override.

ALARM AND ORDER WIRE EQUIPMENT

Complementing the CW-60 is the solid state CVA-22 Order Wire and Alarm shown in the lower door of Fig. 6. A

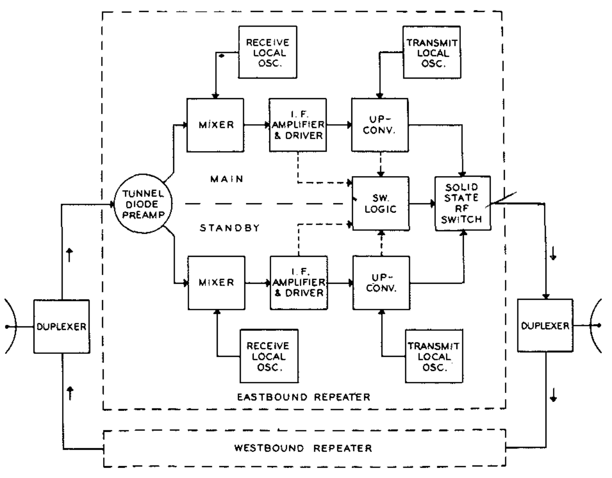


Fig. 4—Two-way standby heterodyne repeater.

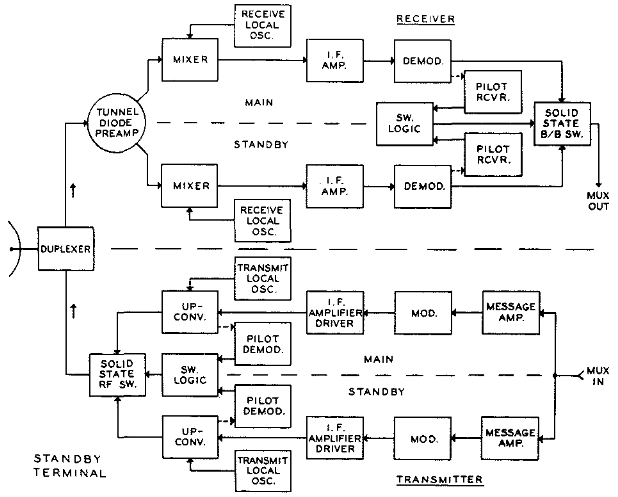


Fig. 5—Standby heterodyne terminal.

handset and loud speaker, which may be muted, provide party-line connections to all stations of a section. A button-operated signalling oscillator can release the mute at all stations during loud hailing calls. The connection to the radio is made via the FM crystal oscillator in the *transmit* local oscillator and the local demodulator. External four-wire connections can handle standard switchboard levels. Hybriding options have been designed for junction stations and E&M signalling facilities are available for extension to dispatcher or operation control points.

The talking facility occupies the physical band below 4 kc while the alarm system is contained within the 4- to 8-kc band. The alarm equipment at each reporting point, such as a repeater station, monitors 14 points and reports the status of those points to a supervisory location. Each supervisory location has the capability of monitoring up to 18 reporting stations and special provisions can double that capacity. Status and decoding modules are shown exposed in the shelves of Fig. 7.

The alarm reporting equipment produces a train of digital pulses. This train, called a message, frequency-shift-keys (FSK) an RCA CT-42 Tone Transmitter. The FSK tone is then transmitted over the CW-60 radio to the supervisory point. There it is demodulated by a complementary CT-42 Receiver and the received message and the pulse train are passed to the decoding and display equipment. From this display an operator visually notes the status of each station and the monitored points therein —whether they are *normal* or *off-normal.* Audible alarm is given when any major function becomes *off-normal.*

The supervisory location consists of status and decoding card shelves with decorative but functional lamp display doors (Fig. 8). The status shelf functions as an indicator of the general condition of the system by station, while the decoder shelf provides detailed information about a particular station selected by a switch.

Each status shelf and door has a capacity of six stations and displays *status* and *tone failure* for each. The shelf houses six tuned FSK tone receivers corresponding to the stations monitored. The output of these receivers is the pulse train or message. This is processed by the status module and operates the *status* and *tone* lamps. The output of any receiver can be connected by a switch to the decoder shelf.

The decoder door has 14 alarm lamps, two combination lamp/switches, *data* and *acknowledge,* plus a six-position status shelf selector. Three modules—decoder, shift register, and alarm driver —are provided in the shelf.

PACKAGING

Microwave relay equipment users have generally adopted 8-foot ceiling heights. However, the number of radio racks for a tube equipment in a hot standby repeater station ranged from three to six depending upon auxiliary features. The floor area occupied in an office building by such an array presented a serious space burden on the user.

In developing the CW-60 equipment, the Broadcast Division anticipated a progressive reduction in space requirements, as competitive suppliers joined the solid state trend, and foresaw an eventual need for the attractive cabinet. The CW-60 contains a hot standby repeater with order wire, alarm-reporting, and switching circuits in a single 7½-foot cabinet. This calls for dense packaging which presents a heat dissipation problem. The solution, simple and effective, is a natural chimney flow formed at the front and rear by the unit panel profile. In this way, blowers, which consume power and can become maintenance nuisances, are not necessary. The CW-60 enclosed in simple huts, will perform in environments ranging from hot deserts to the frigid mountain tops.

Fig. 6 shows a hot standby repeater as well as close-ups of several constructional features. These include slide-in printed circuits (also shown in Fig. 9),

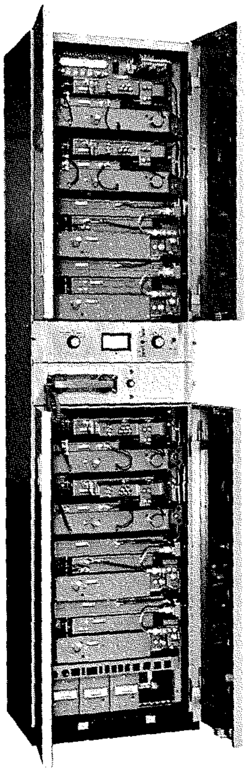
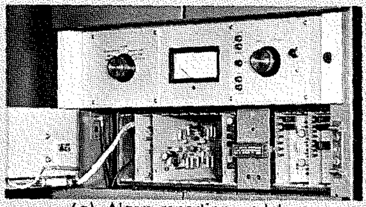
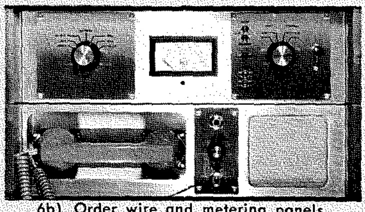


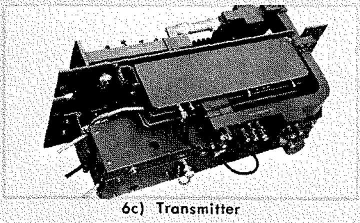
Fig. 6—Standby repeater rack.



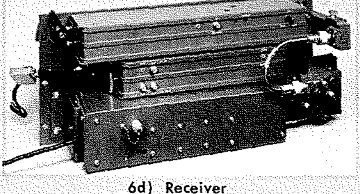
6a) Alarm reporting modules



6b) Order wire and metering panels



6c) Transmitter



6d) Receiver

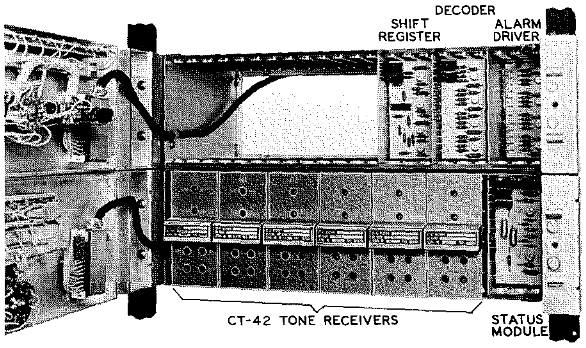


Fig. 7—Supervisory shelves. The three modules at top right are (l. to r.) shift register, decoder, and alarm driver. The six similar modules in lower center are the CT-42 tone receivers. To their immediate right is the status module.

dip brazed shielded assemblies (containing printed circuits), aluminum castings for heat radiators, and waveguide and cavity components.

The CW-60 cabinet is compatible in appearance and profile with the CV-600 voice multiplex product line.[4] Economical and decorative double doors provide access to the units for maintenance. Normally, routine operational and status checks are made with the controls and indicators on the central door in the inset of Fig. 6. Similarly, the order wire circuit is within easy reach of the maintainer.

PERFORMANCE

Table II presents the important electrical parameters of the equipment. The system performance specification is competitive with that of other equipment with higher power which is either all tube or hybrid klystron—and—solid-state remodulation equipment.

The typical overload curve of the CW-60 system is shown in Fig. 10 illustrating the optimization of system bandwidth, modem linearity, and group delay equalization. The baseband idle noise distribution for a typical receiver is presented in Fig. 11. The uniformity of this characteristic is indicative of the minimizing of noise contributions of the crystal-controlled receive and transmit local oscillators. These oscillators have been provided with sufficient bandwidth plus a modulator to facilitate 120-voice channel insert capability at a heterodyne repeater.

THE FUTURE

The choice several years ago of the heterodyne configuration for RCA's high capacity tube equipment permitted the evolution of a 6,000-Mc version from the 2,000-Mc microwave equipment MM-600. These microwave units became the trunks of major transcontinental systems (Alaska, Western Union, Trans Canada, CENTO). The CW-60 has a similar flexibility, so that it can be extended to other bands beyond the common carrier and industrial and for uses other than telephone multiplex transmission.[5] The use of building block techniques produces a versatile product line.

CONCLUSION

The CW-60 is a medium capacity, total solid state, microwave relay system developed for both short and long haul service. It has been designed to provide flexibility from the aspect of protection and station configuration. The CW-60 product line is supported by a solid state voice multiplex line and an alarm reporting equipment. It is presently being designed for other operating bands and uses to broaden the CW-60 market base.

TABLE II—CW-60 Specifications

Carrier Frequency Range	
Common Carrier	5,925—6,425 Mc
Operational Band	6,575—6,875 Mc
Baseband	
Frequency range	60—1052 kc
Input-output level	—45/—20 dbm per channel
Impedance	75 ohms
Service Chan	
Orderwire	0.3 to 4 kc
Alarm tones	4 to 8 kc
Input-output level	—16/+7 dbm
Impedance	600 ohms
Alarm Functions	14 at each of 18 stations (CVA-22)
Standby.Switching Time	1.5 msec
System Capacity	240 voice channels
Insert Capacity	120 voice channels (@ heterodyne repeaters)
1000 Mile Performance	52 db (FIA) signal-to-noise ratio
Transmitter	
Power output	0.1 watt nom.
Freq. stability	±0.0015%
Emission designator	10,000 F9
Receiver	
Noise figure	7.5 db
Intermediate frequency	70 Mc
Bandwidth	20 Mc

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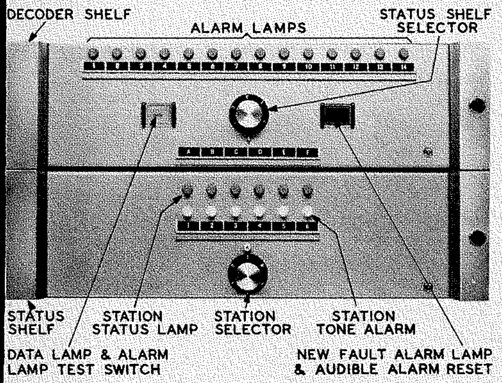


Fig. 8—Supervisory display.

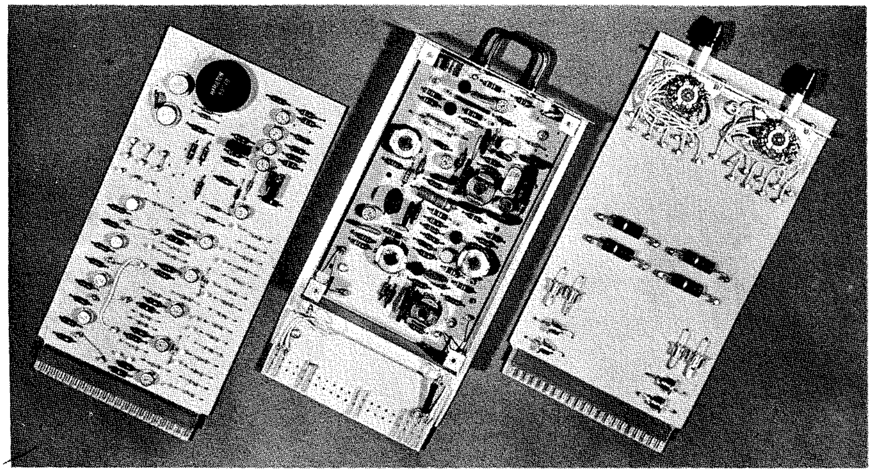


Fig. 9—Switchover modules: left, alarm extend board; center, pilot receiver; right, manual control.

Fig. 10—Loading characteristic

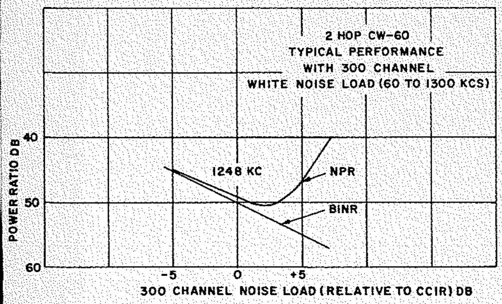
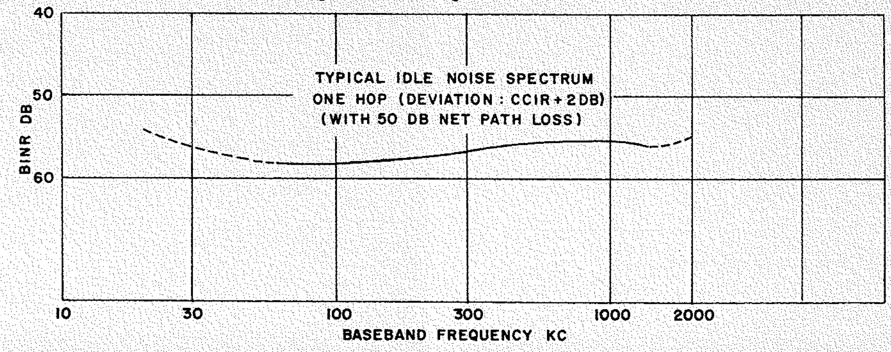


Fig. 11—Quieting characteristic.



SOLID-STATE CV-600 FREQUENCY-DIVISION MULTIPLEX FOR 600 VOICE CHANNELS

The design of RCA's CV-600 Frequency Division Multiplex equipment provides the natural complement to the new solid state CW-60 Microwave equipment. This paper discusses the scope and importance of multiplex in communications and describes the CV-600 and its salient engineering problems and features. The CV-600 design provides multiplexing to a common baseband of up to 600 voice-frequency channels for transmission on microwave, radio, or line links. It provides all standard signalling options for out-of-band working, and includes complete carrier generation, supervisory, standby and testing facilities for the smallest system to the largest and most sophisticated installation.

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MAKING a profit in the field of communications is a multisided challenge like any business activity. It is not sufficient to design good equipment if the market is not there, if the equipment cannot provide the facilities the customer wants at the time of sale, and if the price isn't compatible with the features supplied.

For several years, RCA has recognized that there is a large and constantly expanding communication market requiring frequency-division multiplex equipment.[1,2] Such equipment would enhance RCA's long established capability in microwave communications both for the private user such as the railroad, public utility or pipeline operator, and for the ever expanding international market.[3,4]

Carrier multiplex originated in the United States after World War I as a solution to the growth of telephone traffic. Simply stated it enables many telephone circuits suitably transposed in frequency to different parts of a common baseband to be transmitted over one line or radio link. Because there is a desire to have a large number of circuits on any one link, bandwidth economy is vital. The most economical arrangement would transmit only a single sideband and then only of sufficient bandwidth to include those frequencies which contribute significantly to the intelligibility and character of the speech. As a result, almost all high density carrier multiplex designs use single sideband channels with the transmitted audio restricted to 300 cps to 3.4 kc. Channels are spaced at 4-kc intervals to allow sufficient gap between the sidebands for separation by filtering at the receiving end. The carriers are usually suppressed well below the speech energy in order to reduce the loading requirements on the system.

Carrier multiplex of this type is used throughout the world on all types of cable and microwave trunk circuits. These circuits contain a great variety of different equipments. It would be reasonable to assume that the interface problems between different systems on long circuits would present considerable difficulty, yet it is possible for you to speak over great distances to any part of the civilized world.

The key to a satisfactory national and international telephone circuit is standardization of performance and methods of operation including levels, frequencies, impedances, stability, etc. In the U.S.A. the commercial telephone networks have largely established their own common performance and operating standards. International communication standards have been established by agreement among the nations acting through the CCITT (International Telegraph and Telephone Consultative Committee) as part of the International Telecommunication Union.

CCITT 'standards' issued in the form of recommendations are designed to eliminate interface problems, establish overall performance standards and to assist the orderly growth of communications. They are deliberately constrained in their scope so as not to restrict the growth of new techniques or equipment designs. U.S. commercial and international CCITT standards are alike although equipment types and operating practices may vary. However, such interface problems as do exist between this country and overseas circuits have been solved without great difficulty by the use of adaptor or convertor equipment developed for this purpose.

To satisfy the widest possible market, RCA's Carrier Multiplex was designed to meet all CCITT requirements and to have a performance level suitable for long distance toll quality circuits. All active circuits throughout the equipment use solid state devices giving the well known advantages of low power consumption, small size and high reliability.

Performance required for the private microwave business is generally similar or identical to CCITT standards. However, there is a significant difference in the economic justification reflected in the price and sophistication of the equipment for the two market areas. The end result is a need for lower cost equipment on the lower density short haul systems generally found in the private communication business.

SYSTEM DESIGN

The simplest design for a multiplex carrier system would involve modulating each channel with a different carrier, filtering out each single sideband and combining all sidebands into one common baseband. However, for a system involving hundreds of channels, this would involve hundreds of individual

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Final manuscript received Sept. 15, 1964

FRANK L. CAMERON graduated with a B.A. (Cantab) in Natural Sciences from Cambridge University, England in 1940. After serving in the Electrical and Mechanical Engineering branch of the Armed Forces until 1946, he joined Central Rediffusion Services London where he worked on the development of a new coaxial Television Distribution System for use in Montreal, Canada. In 1952 he joined the British Telecommunication Research, Ltd. in charge of their wideband line carrier development and later as Senior Project Engineer took charge of all 4 mc and 12 mc wideband carrier projects. From 1959 he was responsible for the development of a high speed automatic transistor tester using digital control circuitry and later carried out a design study of internal data transmission problems for factory control. Mr. Cameron joined RCA's Microwave Department at Camden in 1961 where he was responsible for building a Multiplex Design Group. He is presently Group Leader in charge of the CV-600 Carrier Multiplex development. He is a member of the IEEE and the British IRE and has been granted several patents.

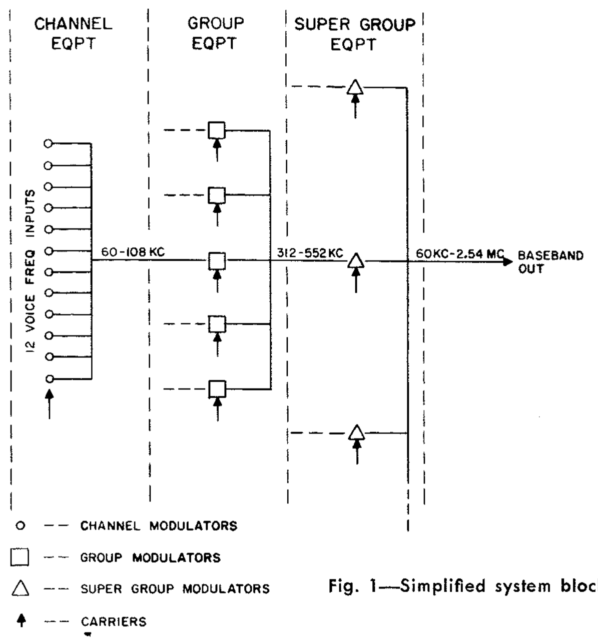


Fig. 1—Simplified system block diagram.

carrier supplies and hundreds of different filter network designs.

The cost of the large design effort needed and complexity of the resulting equipment would make this solution impractical. If however a limited number of channels are blocked up in this way and two or more of these resulting blocks are remodulated so as to shift them to different positions in the baseband spectrum, the resulting saving in carrier supply equipment and network design is considerable. The same process can be repeated in further 'higher' modulation stages to achieve simplification in the overall circuit complexity even for a large number of channels (Fig. 1). Historically, the first design to use these principles was the Bell System A-type channel bank designed for 12 channels, with each sideband spaced at 4-kc intervals in the frequency band 60 to 108 kc.[5,6] The scheme has been adopted internationally and forms the basis of the standard CCITT modulation plan. Three modulation 'stages' are used for a 600-channel system involving three separate blocks of equipment.

1) Channeling equipment to produce a 12-channel group from 60 to 108 kc.
2) Grouping equipment to produce from five such groups a standard 60-channel super group from 312 kc to 552 kc.
3) Super-grouping equipment to translate the standard 60-channel super group to any one of the different positions in the 600-channel spectrum (60 kc to 2.54 Mc).

The RCA channeling equipment uses a two-step modulation process within the channeling equipment itself to derive the standard 60-to-108-kc, 12-channel group—but this is incidental to the standard principles outlined above though important to the channel bank in further simplifying its design.

The RCA multiplex modulation plan to comply with CCITT is shown in Fig. 2. The 0.3-to-3.4-kc voice channel is modulated with a 28-kc carrier to give a sideband from 28.3 to 31.4 kc. Twelve of these sidebands are further modulated, with one each of the carriers from 92 kc to 136 kc to form the 60-to-108-kc, 12-channel group. This constitutes the first basic modulation "stage." The following group-modulation and super-group-modulation stages are also shown in Fig. 2. The basic super group referred to above is also called super group No. 2. This super group is passed through direct without further modulation, as shown in Fig. 2. It will be noted that there are only 27 carriers in the channeling, grouping, and super-grouping equipment for a complete 600-channel terminal. It will also be seen that all except the first modulation step involves using only the lower sideband of the modulation process thus eliminating most, if not all, of the higher-order and unwanted intermodulation products produced by harmonics of the carrier frequencies.

Corresponding to this plan, Fig. 3 shows a simplified block diagram of the channel, group, and super-group equipments. Each equipment consists essentially of modulating circuitry for the frequency translation, filters for selecting the required output, and a means of combining the various outputs in a noninterfering way.

The process so far described involves the stacking up of many channels into a common baseband for outward transmission. All speech channels, however, are two-way and this involves the reverse process at the receiving end of the system. The equipments described above

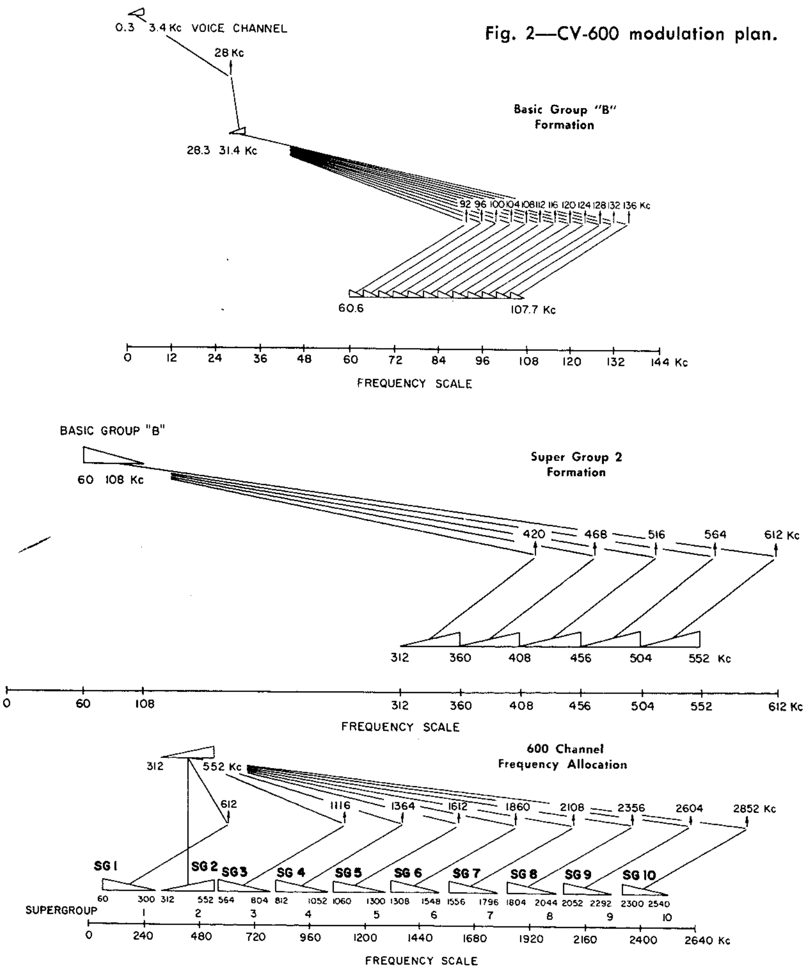


Fig. 2—CV-600 modulation plan.

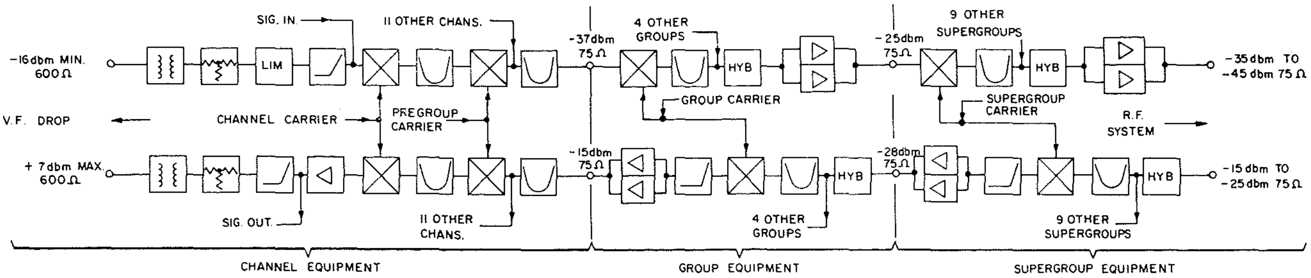


Fig. 3—CV-600 system.

therefore contain both *MOD*ulators and *DEM*odulators (generally called *modems*) to give the required duplex two-way working. From the customer's point of view, the overall channel circuit response from audio to audio is all important. Fig. 4 shows a typical channel amplitude and group-delay response for CV-600 equipment as compared with CCITT limits normally allowed for a back-to-back terminal (i.e. 1/3 of CCITT limits for an international circuit).

CHANNEL EQUIPMENT

The channel modem shown in Fig. 5 forms the heart of the multiplex system since there is one such unit for each channel end and two for each two-way circuit. Several problems make this unit far and away *the most difficult* single unit to design.

The problem of major technical complexity is the design of the channel bandpass filter and its associated low-pass filter (used to bolster the channel-bandpass-filter response and eliminate speech interference at the signalling frequency). An effective design must use the minimum number of components for low cost and yet give considerably lower sideband rejection and a high response stability under wide temperature variations.

The RCA channel filter response and structure based on an insertion-loss type of design is shown in Fig. 6. The asymmetry of response with frequency made it impossible to design by the use of standard tabulated functions and resulted in the decision to develop a potential analog plotter[7] from which it was then possible to derive the required pole and zero values of the rational polynomial. In addition, the need to design approximately 32 other complex filters for the full multiplex project required a powerful all-round network design capability. At the start of the project a range of computer programs was created to enable the designers to synthesize the networks, analyze the performance of the result for cross checking purposes and realize the element values.[8] Networks were built only after the full theoretical calculations had proven their performance characteristics. Over 35 different structures were examined and synthesized before a suitably stable channel bandpass filter was obtained to meet the required specification.

The modulators in this unit are balanced transistor modulators working in a common base circuit configuration. This form presents a constant impedance allowing it to be integrated into the design of the filter with accurate and predictable results. The two-step modulation process used results in a universal channel modem identical for each channel of the 12-channel 60-to-108-kc group, compared with a one-step modulation process where each of the 12-channel filters would be different and occupy the 12-channel positions in the 60-to-108-kc band. In the RCA universal channel modem the pre-group frequency alone determines the channel position giving complete module interchangeability, greater maintenance flexibility and allowing a substantial reduction in the number of spare modules required.

The channel modem cost has a dominating influence on the overall cost of a complete system. In consequence the design of this unit must be technically optimum and have the lowest possible cost consistent with the required performance.

At the output of the modems, the channels are connected to the channel combining unit in a low-impedance circuit designed to minimize interference between the channels. Serviceability of the equipment is important. It is therefore essential for the user to be able to remove channel modems for this or other reasons *without* upsetting the levels or response of other channels in the same Group and this has been achieved in the combining circuit by designing for a high output impedance in the modem.

Fig. 4—Channel amplitude and envelope delay response (typical).

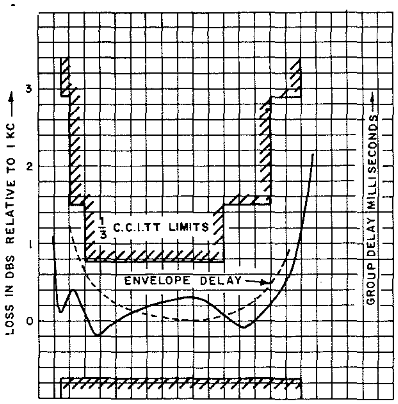
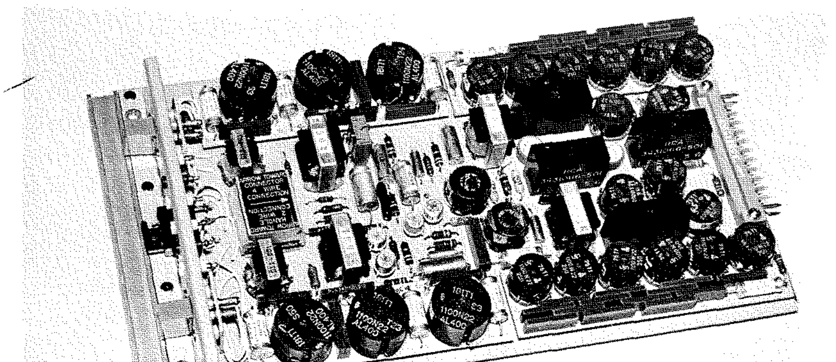


Fig. 5—Channel modem.



The combining unit also includes a "clean-up" 60-to-108-kc filter.

The complete assembly mounted in a single shelf is called the 12-channel bank.

GROUP AND SUPER-GROUP EQUIPMENT

The grouping equipment having the function and form shown clearly in Figs. 3 and 7 is composed of five group-modem units and one group-combining unit in a single equipment shelf. Each group modem translates a 60-to-108-kc, 12-channel group to one of the five positions in the super group No. 2 spectrum (312 to 552 kc) and filters out the desired output. The translated groups in this case are combined with wideband hybrid transformers into the common 60-channel output.

The general design of super-group equipment is similar to the group equipment though requiring higher frequency and wider-band filters and modulators.

Throughout the various translation stages negative feedback amplifiers are used to compensate for circuit losses. Those amplifiers affecting 12 or more channels (i.e. within the group or super group) use a standard redundant design (see Fig. 8) with parallel emitter to emitter feedback with the outputs connected through wideband hybrids. By this approach isolation of each amplifier half is achieved. The design is proof against open or short circuit faults on all the active components. The separate three stage transistorized amplifiers with their associated circuitry may be unplugged and removed individually for servicing without affecting the overall transmission level.

SIGNALLING EQUIPMENT

Each telephone circuit, to be of any value, must also transmit signalling information to inform the distant operator of incoming calls or operate automatic switching equipment for circuit routing purposes. Unfortunately, many forms of telephone signalling are in practical use today. They divide, however, into two broad categories: *inband* and *out-of-band* systems each with advantages and disadvantages for different applications. (These systems use tones within or outside the audio transmission band of 300 cps to 3.4 kc.)

The RCA signalling equipment (a version of which is shown in block schematic form in Fig. 8) uses an out-of-band signalling tone at 3,825 cps which is injected into the audio circuit within the channel modem unit in the *transmit* direction. Signalling tones at 3,825 cps received from the distant terminal are extracted from the high-level audio *receive* output. Signalling information may originate at a local subscriber's dial telephone or from a switchboard in the form of DC pulses transmitted at a rate up to 14 pulses/sec. These arrive at the carrier terminal on a *transMit* wire conventionally called the *M* lead. In the *recEive* direction, dial pulses are sent back to the exchange on the *E* lead. This *E & M* signalling is the most common system in use. The RCA *E & M* signalling unit is a high-quality unit designed to pass dial pulse information using a tone at −20 dbmo with mark-space ratios varying from 47.5 to 67.5%. The signalling level used must be low enough so as not to overload the system, since supervisory signals may be transmitted back while the circuit is in use. If faults occur on a system degrading levels by more than 10 db, the resulting signalling tones may cause incorrect circuit switching—even though the audio signal degradation may not be noticed by the user. To avoid such problems, the RCA signalling unit is made to operate down to 10 db below its nominal working level at which point it ceases to transmit further information giving instant warning of circuit failure.

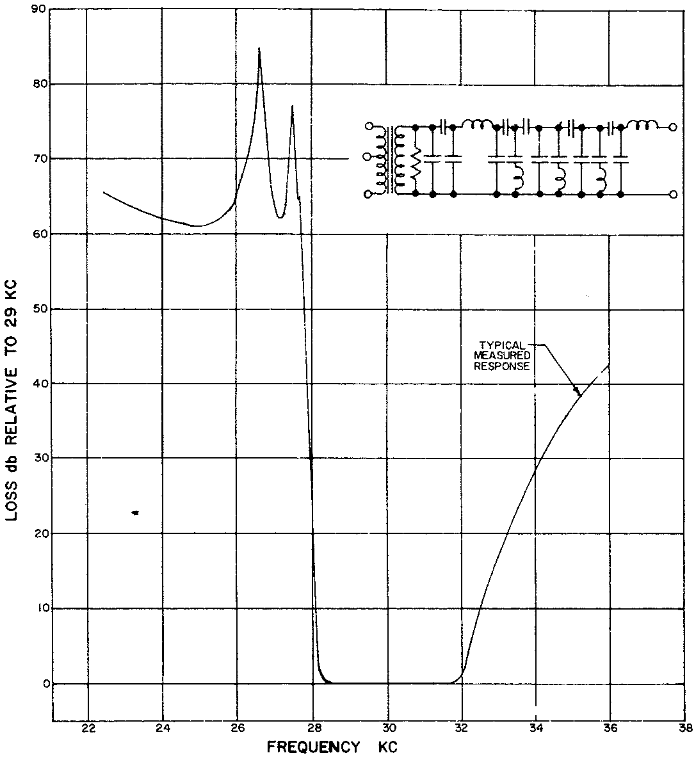
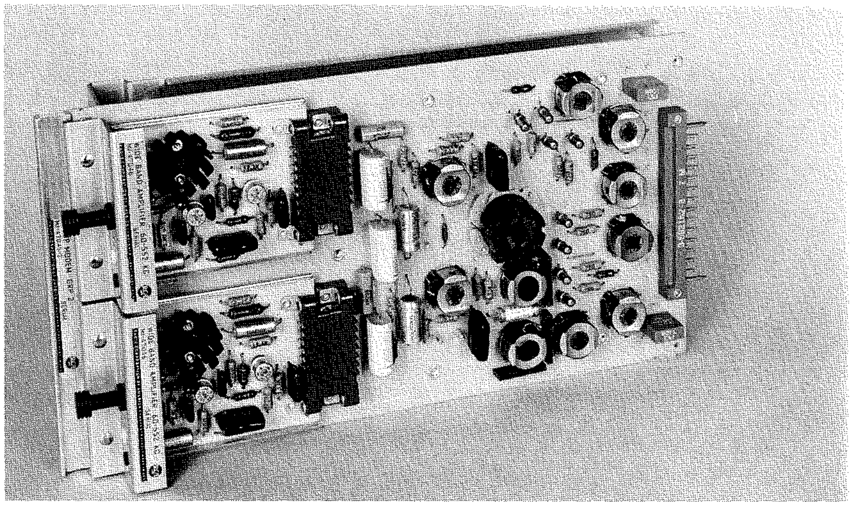


Fig. 6—Channel B.P.F. response

Fig. 7—Group modem.



Not all telephone circuits originate

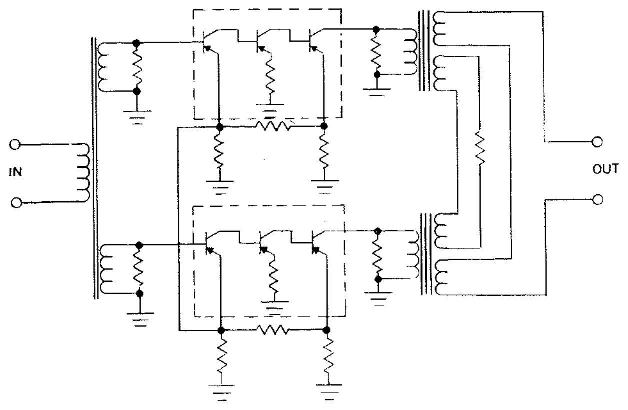


Fig. 8—Basic redundant amplifier design.

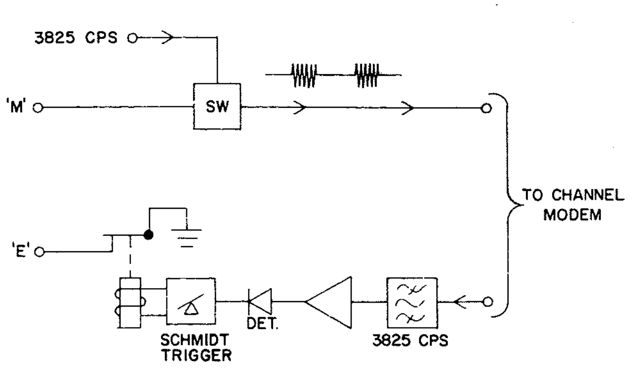


Fig. 9—E & M signaling.

dialing tones. Purely *on-off* signalling which will not be present whilst speaking can be transmitted at a higher signal tone level with a corresponding simplification in circuit complexity. The RCA design provides a total of four different standard signalling units (the *E & M* unit is shown in Fig. 10) which combine in various ways the dial pulse and *on-off* type of control for transmit and receive directions. These options are:

1) *E & M signalling*
2) AC *ringdown* (uses 20-cps ringing tone to the telephone line)
3) DC *loop subscriber* (extends a DC loop from the subscriber's end of circuit)
4) DC *loop switchboard* (closes a DC loop to the switchboard end of circuit)

CARRIER GENERATION EQUIPMENT

The production of carriers for the various channel, group, and super-group modulators of a high-channel-capacity system requires generation equipment with a high order of frequency stability.

A channel modulated to a high frequency sideband and demodulated back to audio will retain a "pitch error" equivalent to the frequency difference between the two carriers. The CCITT recommendation for maximum pitch error at any audio frequency is 2 cps overall on any long-distance circuit. By definition, this allowance would include the error of three transmitting and three receiving multiplex terminals which can add or subtract in a random fashion. Stability of this order is only obtainable with special temperature controlled crystal oscillators using high stability crystals. Unfortunately, as is well known, crystal "frequency" drifts with time. Stability drifts as low as 1.5 parts in 10^7 per month would still cause nearly 0.5-cps error in a month at the highest super-group carrier on a 600-channel system.

The principle followed in the RCA equipment as well as in most multiplex designs is to concentrate on a single high-stability master frequency-source from which all other carriers may then be derived.

Fig. 11 shows a simplified block diagram of the carrier generation equipment. The RCA Multiplex uses a single master oscillator at 1,024 kc utilizing an AT cut crystal enclosed in a proportional controlled temperature oven maintained at 70°C to a fraction of a degree. This frequency is then divided down with binary dividers to a fundamental 4 kc which is amplified and use to drive a saturated core harmonic generator.[9] The output 4-μsec pulse with an amplitude of approximately 90 volts is rich in harmonics, and the required channel and pregroup carriers are filtered off directly. Group and super-group carriers are derived by extracting 12 kc and 124 kc, and using these to drive further harmonic generators to produce the higher carrier frequencies.

Failure of carriers may involve many channels and cause serious interruption of traffic. Therefore, the carrier equipment can include as shown automatic standby or redundant equipment for all active circuitry. A standby master-oscillator and harmonic-generator unit can be provided, as is Fig. 11. Failure of the 4-kc pulse at the output of the main harmonic-generator unit will initiate an automatic switch over to the standby units by means of a sealed mercury relay. Identical standby facilities are provided for the 12-kc-group and 124-kc-super-group amplifier and harmonic-generator units. All amplifiers shown are similar to those described earlier for the transmission equipment. In addition the output of carrier amplifiers is monitored continuously with circuits designed to sense the presence of carrier and initiate alarms on failure.

Although carrier generation equipment of this type satisfies all requirements for speech transmission on long distance circuits, the advent of high-speed data transmission has introduced the need for even higher stability and lower pitch error. Such improvement can best be obtained by synchronizing the master frequency-source by means of a transmitted pilot.[10] The RCA synchronizing system uses a standard 308-kc pilot which is filtered out with a selective crystal filter, amplified, and divided down to 4 kc. At the same time, a free-running crystal oscillator at 128 kc feeds into binary dividers giving a resulting output also at 4 kc. The two 4-kc signals are injected into a phase-comparator circuit whose output is used to control a varicap in the crystal oscillator circuit.

The synchronizing units as presently designed can be directly substituted for the standard master-oscillator unit giving zero pitch error for those systems where data transmission requirements demand it.

DATA

The need for and growth of data communications has led in the last decade to much investigation into the potentiality and limitations for data transmission of the existing telephone networks.[2,11,12,13]

The Carrier Multiplex equipment is an important part of this network and must play its part in satisfying the special needs in this area. Data information can be subdivided into:

1) *Slow-speed data* — teletype, for example up to 100 words/min equivalent to 37.5-cps keying rate.

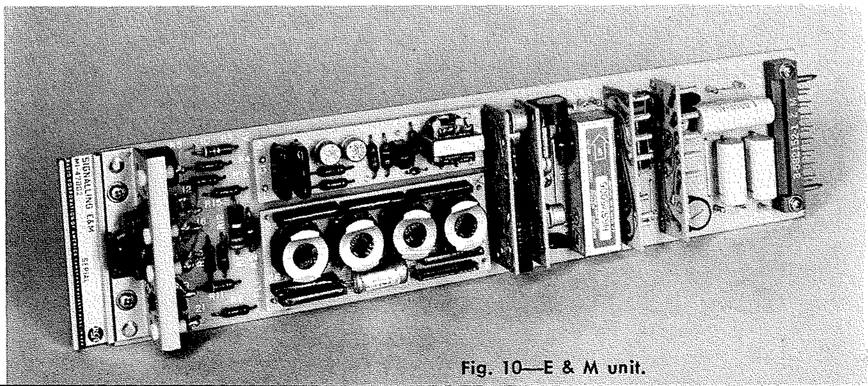


Fig. 10—E & M unit.

2) *Medium-speed data*—up to maximum of 1,500 baud.
3) *High-speed data*—greater than 1,500 baud.

Apart from supplying a satisfactory amplitude frequency response the Multiplex performance is critical in two main respects: *pitch error* and *envelope delay.* The CV-600 pitch error performance of less than 2 cps (CCITT requirement) overall will satisfy the majority of message forms; however, high-speed synchronous data systems particularly of the form using clock sampling techniques demand zero pitch error. Substituting the two synchronizing units in place of the master oscillator is the only action then required.

The channel-envelope delay response given in Fig. 4 is also satisfactory for and does not limit the transmission of slow- and medium-speed data on the normal telephone network for acceptable error rates.[11,14] The use of high-speed data would, however, require group-delay equalization of the voice-frequency channel. Since all CV-600 modems are the same, a standard group-delay equalizer can be added in the applications where this is required.

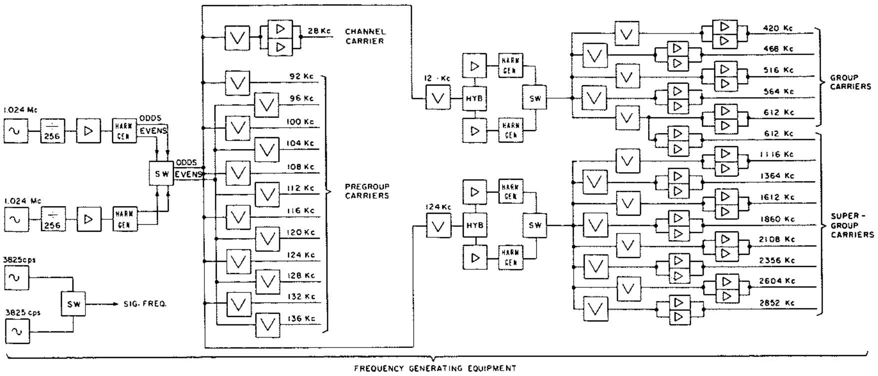


Fig. 11—CV-600 carrier generation.

PACKAGING

All multiplex transmission, signalling and carrier supply units were designed on printed-circuit boards as plug-in modules. Although various designs of mounting shelves for plug-in cards already existed within RCA, no one of them was entirely satisfactory.

The shelf requirements could be summarized as: 1) low cost, 2) high strength, 3) flexibility for mounting position of modules, 4) open construction for maximum air circulation, and 5) separate mounting plate for the rear connectors. The design of a suitable card nest which meets all these requirements is illustrated in Fig. 12. Plastic card guides were designed to fix with spring metal clips into holes in the top and bottom plates. Two hole spacings have been provided: 13-way and 24-way, with room for a third row of holes of arbitrary spacing if ever required. Card shelves of any height are possible with this approach using the standard bottom and top plates with varying size of side plate. So far, only 2- and 4-unit-high shelves have been used.

CONCLUSIONS

The design of RCA's CV-600 Frequency Division Multiplex equipment provides the natural complement to the new solid state CW-60 Microwave. This paper has shown the scope and importance of multiplex in communications and to provide a general description of the CV-600, with the salient problems and features. The CV-600 design will provide multiplexing to a common baseband of up to 600 voice-frequency channels for transmission on microwave, radio, or line links. It provides all standard signalling options for out-of-band working and includes complete carrier generation, supervisory, standby and testing facilities for the smallest system to the largest and most sophisticated installation.

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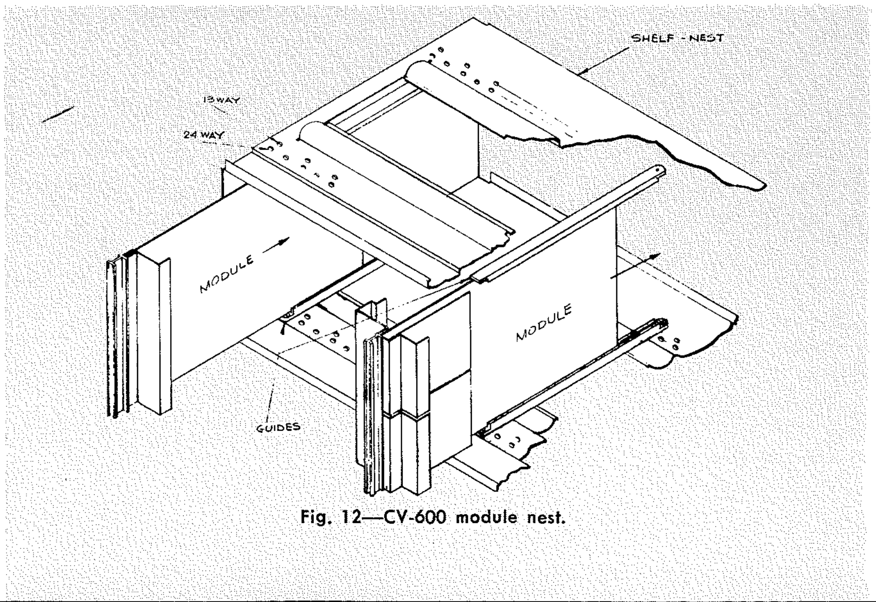

Fig. 12—CV-600 module nest.

FUTURE MICROWAVE COMMUNICATIONS REPEATERS FOR SPACE

Improved booster performance coupled with an ever advancing technology has created a need for improved performance in space communications equipment. This paper describes several microwave communications repeaters which use various configurations of solid-state devices, conventional heterodyning, and traveling wave tubes. Limitations in power and bandwidth are discussed. It is shown that peripheral problems relating to modulation techniques, spacecraft stabilization, spacecraft power, and user access have a critical bearing on the design, performance, and reliability of the equipment.

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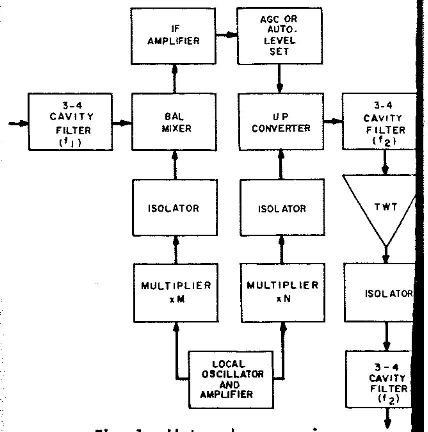

Fig. 1—Heterodyne receiver.

Active communications satellites are required to receive and retransmit communications signals without introducing significant distortion. The repeaters that perform this function must make efficient use of the available power to provide sufficient traffic capacity. They must also provide sufficient reliability to satisfy the system performance and economic requirements. In general, the spacecraft configuration largely depends upon the overall technical and economical objectives, and various trade-offs thereof. For example, a passive communications system using an ECHO-type balloon places the complexity burden on the ground station. Active medium-altitude communications satellite systems like RELAY or TELSTAR require a relatively complex satellite with less of a burden on the ground station. Active stationary satellites are generally even more complex because of attitude the "station-keeping" requirements. On the other hand, the ground station is less complicated because antenna steering is minimal. It should be kept in mind, therefore, that the satellite equipment is intimately related to an overall systems concept.

Generally, the efficiency of these equipments depend basically on the efficiency of the final amplifier. Traveling wave tubes (TWT), triodes, and tetrodes are commonly used at medium power levels and at the higher frequency ranges, generally above 500 Mc; transistors are used at low and medium power levels and at lower frequencies. The techniques used are similar to ground-based equipment; however, the thermal and voltage-breakdown problems are somewhat different. Because of the primary power limitations, output efficiency is a key parameter of space systems. The performance is affected not only by the basic efficiency of the output stage and other space electronic systems, but on the modulation scheme selected.

Frequency and phase modulation make effective use of the average transmitter power available and can provide high processing gains (FM improvements). The extra frequency spectrum required has not been a serious disadvantage as yet. Amplitude modulation systems have been used in space systems but not in systems requiring high capacity. Amplitude modulation is relatively inefficient for space use because the relatively high average power and peak power required is difficult to supply from solar-cell arrays. Traveling wave tubes, for example, have a large bandwidth capability and are relatively efficient transmitters of FM or PM signals with excellent phase linearity. However, the traveling wave tube is generally an inefficient AM amplifier due to the amplitude nonlinearities at saturation (the optimum operating point). The following discussion describes some configurations, existing hardware implementation, traffic capacity, reliability, and other significant parameters of the repeaters.

TWO REPEATER CONFIGURATIONS

Two particular repeater configurations are discussed here in detail: 1) the intermediate-frequency repeater, and 2) the reflexed TWT repeater. Both involve traveling wave tubes and solid-state devices, but with interesting differences.

Operation of Intermediate-Frequency Repeater

The first configuration (Fig. 1) is a conventional heterodyne receiver (preselector, mixer, and IF amplifier). The output of the IF amplifier drives an "up" converter to convert the IF frequency to a microwave frequency. The IF amplifier provides the required gain. A TWT usually serves as the final output stage. There are many variations of this basic configuration. For example, separate local oscillators can be used, instead of the one shown in the figure, in order to ease the frequency allocation plan (a single local oscillator restricts the frequency choices available). The up converter can be operated linearly or at saturation (limiting). Also, the low-power pump may be circulated through the TWT before application to the up converter, in order to improve receiver efficiency. In the case of the low-power pump, lower conversion losses may be obtained. The interstage network in the IF amplifier may be one of the following:

1) RC type; to minimize group delay* and enable FM-AM-PM conversion (which causes crosstalk if more than one carrier is present in the IF); (The group delay is defined as $d\phi(\omega)/d\omega$ where $\phi(\omega)$ is the transfer phase function. It can be shown that terms of $d\phi/d\omega$, which are nonconstant with the frequency, can cause signal distortion of phase modulated waves)
2) synchronously tuned, or
3) stagger tuned.

Bandwidths of the order of 100 Mc are readily achievable as state-of-the-art.

Operation of Reflexed TWT Repeater

Fig. 2 shows the reflexed TWT configuration. To show significant features of the design, the signal paths and levels are given at the input frequency f_1 and the output frequency f_2. Both frequencies are assumed to be in the microwave band. In this configuration, the input signal is amplified in the first TWT, translated to frequency f_2 in a low-level mixer, and recirculated through the first TWT in order to establish the operating point of the final TWT. The output TWT functions as the final amplifier stage at frequency f_2.

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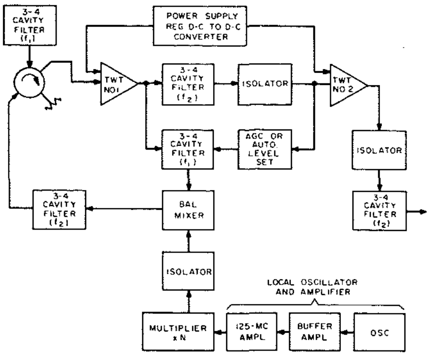

Fig. 2—Reflexed TWT configuration.

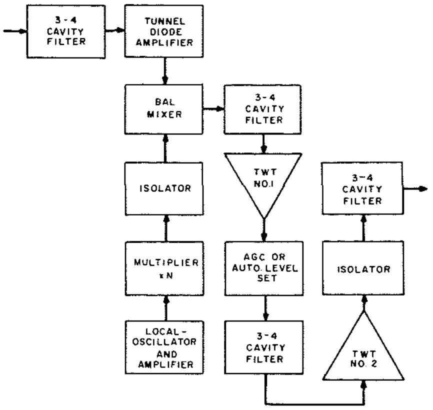

Fig. 3—Alternate TWT configuration with tunnel-diode amplifier.

TABLE I—Comparative Reliabilities of IF and Reflexed TWT Configurations

Type of Component	*I.F. Repeater*		*Reflexed TWT*	
	Components (2 Repeaters)	*Failures* per 10^8 hrs.*	*Components (2 Repeaters)*	*Failures* per 10^8 hrs.*
Capacitor (Tantalum)	29	102	2	7
Capacitor (Glass, ceramic)	263	263	56	56
Diodes	40	400	26	260
Transistor	38	932	12	288
Resistor	260	260	46	46
Coil	78	390	10	50
Cavity	40	40	46	46
Crystal	2	20	2	20
Transformer	4	100	—	—
Isolator	8	80	8	80
TWT	2	400	4	800
Total Failures		*2987*		*1653*

* Failure rates for components from reference 2 except for diodes and isolators which are from reference 3.

The major difficulties with reflex operation are the stability of the reflex loop and the effects of reflex operation on communications performance. It can be shown that the reflex loop can be made absolutely stable at RF frequencies, but that a feedback loop does exist at "modulation" frequencies if the TWT gain is not constant with input drive. Selecting the operating point of the first TWT at 10 db below its saturated power output generally ensures operating-point stability with little penalty in overall efficiency.

With an input TWT having a low-level gain of 40 db, only nominal requirements are imposed on the microwave filters to ensure stability; these requirements may readily be realized by using lightweight three- or four-cavity filters. The reflexed TWT is essentially transparent to communications traffic since the level of the intermodulation products (second-order and third-order products), the crosstalk caused by FM-AM-PM conversion, and the small-signal suppression will be determined by the final TWT, which operates closer to saturation. However, because the TWT gain is a function of operating point, the range of allowable operating points must be limited to avoid envelope oscillation.[1]

The total repeater gain is about 80 db at frequency f_2. To ensure stability, the loss from the output of the final TWT to the input of the first (reflexed) TWT must be in excess of 80 db (at frequency f_2). The specification for the output filters must be determined by this requirement; specifically, the "feedback" signal at frequency f_2 must be restricted to a gain much less than one, to avoid the group delay effects of "multipath." The minimum required loss varies with the traffic requirements; however, the following example is derived. With a single feedback path, and a feedback factor k, the resulting wave V_R may be described by:

$$V_R = V_I\,[1 + k \exp(2\omega l/v)]$$

Where: V_I is the initial wave, ω is the radian frequency, l is the length of the feedback path, and v is the propagation velocity in the feedback path.

The resulting phase is:

$$\tan\phi \approx \phi = k \sin\frac{2\omega l}{v},$$

The group delay is:

$$\frac{d\phi}{d\omega} = \frac{2kl}{v}\cos\frac{2\omega l}{v}$$

When:

$$\frac{d\phi}{d\omega} = \frac{2kN}{f}\cos\frac{2\omega l}{v}$$

Where: N is the ratio of length l to wavelength λ at radian frequency ω, and f is $\omega/2\pi$.

The group delay has the form of ripple, with a peak-to-peak value of $4kN/f$ and a periodicity of approximately $f'/2N$ where f' is the frequency deviation from f. For a feedback path having an electrical length of 100 wavelengths, the periodicity is 20 Mc and the peak-to-peak ripple is 3.16 nsec (for k corresponding to a feedback reduction of 30 db). Thus, the feedback path attenuation must be 30 db.

Comparison of Reliability

In order to determine the reliability, a part estimate was made based upon preliminary designs, and more or less conventional part failure rates were obtained for typical space equipment. A factor was included for moderate part preconditioning. The results are included in Table I. The reflexed TWT configuration is more reliable than is the standard IF configuration according to the results.

Comparing the two sets of data, the ratio of mean-time-between-failures (MTBF) is about two to one in favor of the reflexed TWT repeater (33,500 hours for the IF configuration as compared to 60,500 hours for the reflexed TWT). Use of extensive parts preconditioning and other controls would decrease this ratio, providing an estimated MTBF of 78,000 hours for the IF configuration and 94,300 hours for the reflexed TWT repeator. The reliability calculations assume that the failure rate of the output TWT was identical to that of the reflexed TWT, i.e., both tubes are of like construction. This assumption was made using the long-life criteria established for medium-power output tubes. In addition, noise figures of less than 6 db have been achieved with TWT's using beam-shaping techniques; however, the basic tube reliability and life associated with such techniques have not been demonstrated and may be poor because of the high cathode loading. When limited to low-current densities and techniques related to long-life tubes, the unsaturated noise figure rises to between 15 and 20 db (13 db may be achievable, unsaturated).

The relatively high noise figure is a disadvantage of the TWT repeater; but, depending on the traffic requirements, ground station performance, etc., it may not be limiting because a trade-off exists between satellite noise figure and ground-station transmitter power.

An alternate repeater configuration would precede the input TWT with a tunnel-diode amplifier, as shown in Fig. 3. The tunnel-diode amplifier is inherently capable of a failure rate identical to that of the transistor. The disadvantage of this configuration is that a third TWT must be added to provide comparable gain with the other configurations. This added TWT not only reduces the reliability but presents all the system gain at a single microwave frequency. Shielding for proper performance may become difficult. A preliminary investigation for space use suggests that tunnel diodes are primarily experimental devices at present.

Traffic Potential

The traffic growth potential of the IF configuration is limited by the bandwidth that may be realized in the IF amplifier and up converter. At this time, the realizable bandwidth of the IF amplifier is limited to approximately 200 Mc; the bandwidth of the up converter is more restrictive and depends upon the conversion efficiency. For example, the saturated up converter on the RELAY satellite provided a 15-Mc bandwidth with a 7-db conversion loss. In contrast, the bandwidth of the TWT repeater may be expanded to several hundred megacycles.

Feasibility

The IF configuration employs techniques and hardware similar to those that have been space proven on the RELAY and TELSTAR satellites. For the reflexed TWT configuration, the input TWT must be capable of adequate performance at frequencies f_1 and f_2. This may not be difficult since traveling wave tubes are available with octave bandwidths. Other requirements relating to noise figure, crosstalk, AM linearity, gain, output power, and efficiency must be considered with regard to particular requirements.

A weight estimate for an IF repeater is given in Table II and a power estimate in Table III. A weight estimate for a reflexed TWT repeater is given in Table IV and a power estimate in Table V.

User Access

One final consideration is user access. Thus far the repeater discussion has considered only single access for frequency modulated signals, i.e., one carrier per repeater. Access to more than one user (and therefore more than one carrier), while retaining the advantages of FM, can be accomplished in a multiple FM carrier system if the repeater amplitude response can be made sufficiently linear. That is, two carriers at frequencies f_1 and f_2 passing through a repeater whose amplitude response is:

$$AV_{out} = A_1V_{in} + A_2V_{in}^2 + A_3V_{in}^3 + \ldots$$

will result in products like $2f_1 - f_2$, $2f_2 - f_1$, and other products which will fall outside the band located around frequency f_1 and around frequency f_2 but in the band of other signals in the vicinity. If the repeater including the TWT is made sufficiently linear to make these effects tolerable, the system capacity is generally reduced because of the decrease in TWT power. Other multiple-access systems involving PCM and TDM have been devised and are prolific in the literature. One other interesting method is the SSB-FM system in which a multiplicity of senders transmit SSB. The received spectrum in the satellite may be converted to FM by a remodulation process that does not require the demodulation of the signals. All users receive the single-carrier FM traffic.

TABLE II—Weight Estimate for IF Repeater

Description	*Weight (Ounce)*
TWT	40.0
Power Supply	56.0
Level Set (AGC)	3.0
Cavity Filters (3 @ 3 oz.)	9.0
Mixer	2.0
IF Amplifier	13.0
Up-converter	1.7
Circulator	8.0
Isolators (3 @ 3.3 oz.)	9.9
Multiplex Chain	16.0
Osc. and Amplifier	5.0
Total	*163.6 (10.2 lbs)*

TABLE III—Power Estimate for IF Repeater

Description	*Power (watts)*	
IF Amplifier	1 to 5	
L.O.	1 to 5	(depending on linearity of up converter)
TWT Power Supply	33	*based on:* 10-watts output 38% efficiency 80% P.S. efficiency
Total	*35 to 43 watts*	

TABLE IV—Weight Estimate for Reflexed TWT Repeater

Description	*Weight (ounce)*
TWT No. 2	40.0
TWT No. 1	24.0
Power Supply (Combined)	64.0
Cavity Filters (5 @ 3 oz.)	15.0
Mixer	2.0
Isolators (3 @ 3.30 oz.)	9.9
Circulator	8.0
Multiplex Chain	8.0
Osc. and Amplifier	5.0
Level Set (AGC)	3.0
Total	*178.9 (11.2 lbs)*

TABLE V—Power Estimate for Reflexed TWT Repeater

Description	*Power (watts)*	
TWT Power Supply (for TWT No. 2)	33	*based on:* 10-watt output 38% efficiency 80% P.S. efficiency
(for TWT No. 1)	2	1.5 watts for TWT (0.5-watt filter + 5 to 10% efficiency)
L.O.	≈ 0.300	

CONCLUSION

In summary, the equipment requirements for space communications are characterized by bandwidth, effective radiated power (which is a combination of transmitter output and antenna gain), linearity (amplitude and phase), and severe reliability requirements. Power and weight restrictions are imposed by the launch vehicle. The designer's task is to optimize for the particular mission.

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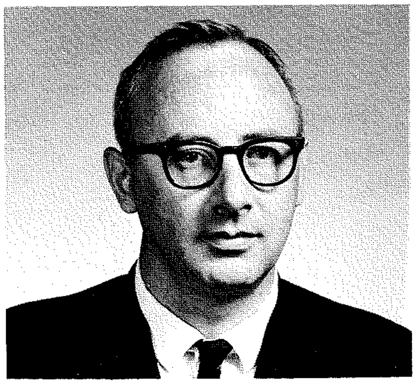

MICROWAVE BREAKDOWN IN SPACE COMMUNICATIONS

A Case Study of RANGER VI and VII

This paper discribes the successful effort that eliminated the voltage breakdowns from the passive components of the RF coaxial lines of the RANGER TV subsystem[1] and includes a consideration of the general RF voltage breakdown problem. Initial tests of the TV and communications subsystem of the RANGER VI spacecraft revealed the presence of RF output power dropouts. The random dropouts occurred in simulated space environments consisting primarily of test vacuum levels of 10^{-4} torr or less. Visual examination of suspected components revealed that RF voltage breakdowns were the cause of the power dropouts. Voltage breakdowns were eliminated from the cavity-tube power amplifiers by pressurizing. The same technique was successfully used on the 50-ohm dummy load component. However, pressurizing did not provide a desirable solution of the elimination of breakdowns from microwave coaxial components, such as the radio frequency connectors or the four-port hybrid. Voltage breakdowns were eliminated from the microwave coaxial components by careful incorporation of special designs which are detailed in this paper.

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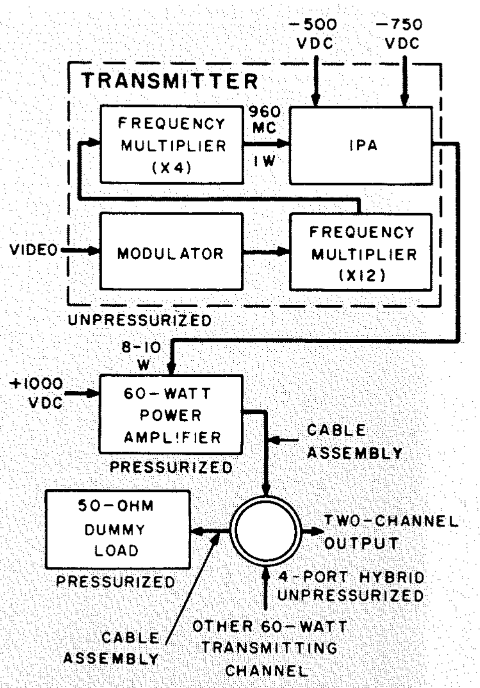

Fig. 1—RANGER communication equipment.

THE planned mission profile of the RANGER spacecraft provides for television transmission[8] through 235,000 miles of space with a path loss exceeding 200 db. Moreover, to ensure the reception of high quality TV pictures at the ground station, a receiver carrier-to-noise ratio of 12 db was established as a system constraint. Spacecraft RF circuit losses also added to the anticipated system requirements so that the required RF power output from the RANGER spacecraft TV transmitter was increased to an unprecedented 60 watts.

In an effort to provide partial redundancy, in addition to increased transmission capacity for the six RANGER cameras, a second 60-watt transmitter was included. Simultaneous transmission from both transmitters (with 1-Mc spacing between carriers imposed by the ground station) is obtained by connecting the transmitter outputs to a single, high-gain antenna through a four-port hybrid (Fig. 1). The second output port of the four-port hybrid is terminated with a high power, 50-ohm dummy load. Therefore, the 60-watt power level is contained in the following components in the TV and communications subsystem: two transmitter 60-watt power amplifiers, the 50-ohm dummy load, four-port hybrid and interconnecting connector cable assemblies.

TYPES OF RF VOLTAGE BREAKDOWN IN SPACECRAFT

The operating environment of a spacecraft containing high-power RF equipment imposes a consideration of two types of RF voltage breakdowns: 1) *an ionization, partial-vacuum breakdown,* and 2) *a multipactor, or hard-vacuum, breakdown.*[1] (Although a multipactor breakdown was never identified as such in this program, the existence of such a breakdown possibility was recognized and investigated.) Fig. 2 is a pictorial representation of the two general breakdown areas under consideration; the ionization breakdown occurring at the higher pressure of several torr, and the multipactor breakdown occurring at the lower pressures.

Both types of breakdowns depend upon the formation of free charge carriers within the breakdown gap. The ionization breakdown, as the name implies, depends upon the ionization of the intervening gaseous medium as its source of charge carriers. The multipactor breakdown requires a certain combination of electric field intensity, frequency, pressure, and boundary conditions to net an increase in free charges. This process depends on the secondary emission characteristics of the gap boundaries as the source of free charges.

The RF dropouts experienced during the initial system tests of the RANGER VI spacecraft occurred only in test chamber vacuums of 10^{-4} torr or less. Moreover, RF dropouts occurred during simulated lunar missions which allowed a vacuum soak of 66 hours before application of power. Therefore, it was reasoned, the power dropouts were multipactor discharges occurring in the lower pressure regions indicated in Fig. 2.

At the time of the initial observation of RF power dropouts it was difficult to determine with certainty where the breakdowns were occurring in the subsystem. To remove the ambiguity of multiple breakdowns occurring in a random connection of several components, retests were planned on an individual component level. This necessitated the removal of the 60-watt line components from the spacecraft. Since prior testing of components, such as the four-port hybrid, had uncovered no deficiencies, a deliberate testing program focused on RF breakdowns was planned.

An examination of available literature indicated two important characteristics of the multipactor breakdown. First, the multipactor breakdown is a random phenomenon dependent on the velocity and position statistics of free electrons to initiate the breakdown process. To eliminate or reduce the effects of this stochastic dependence of the breakdown, another electron producing mechanism was introduced. The second electron producing phenomenon was of the form of external gamma radiation. The gamma radiation incident on the gap boundaries resulted in the scattering of

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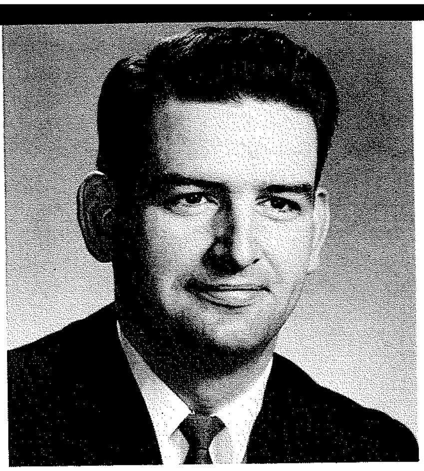


T. D. BREEDEN attended George Washington University and was awarded the degree of BSEE from the University of Maryland in June, 1960. He is currently attending the Drexel Institute of Technology and shall receive the MSEE degree in June 1965. During the summer of 1959, he was a Project Engineer Trainee in the Search and Height-Finding Radar Section of the Department of Navy, Bureau of Ships, Washington, D. C. Mr. Breeden joined RCA in June, 1960, and was responsible for computer design efforts in Moorestown and Camden on the specialized engineering training program. After his permanent assignment in the Microwave Group at AED, Mr. Breeden worked on spacecraft traveling-wave tube amplifiers. More recently he directed a successful program which eliminated RF voltage breakdowns from the RANGER communications subsystem. His work on RANGER led to an AR & D program investigating methods of controlling voltage breakdowns in spacecraft subsystems.

a large number of free electrons into the breakdown gap.

Although the effect of physical variations is not considered, it has been reported that 1 millicurie of a gamma emitter such as cobalt-60 resulted in 10^6 more free electrons in the gap than without the presence of the radiation. However, for present purposes it is sufficient to state that the presence of the gamma radiation increases the probability of the occurrence of a random voltage breakdown without altering the threshold breakdown power level. (A more detailed description of the use of the cobalt-60 isotope to initiate RF voltage breakdowns can be found in Ref. 3 in the *Bibliography.*)

The elimination of the random characteristic of a multipactor breakdown is an important practical requirement. Observations have revealed that the time between application of power and a breakdown is dependent on this characteristic. Indeed, the presence of the isotope has been known to reduce the waiting period to minutes instead of a possible interval of hours.[3]

A second important consideration of the multipactor breakdown is its dependence on power level. This second breakdown characteristic indicated a need for investigating the breakdown susceptibilities of the RANGER V communication components at power levels other than the system level of 60 watts. Two reasons for this are: 1) the elimination of breakdown power levels between the *off* condition and the system level to prevent turn on problems, and 2) the removal of marginal breakdown areas dangerously near 60 watts.

Ref. 4 (*Bibliography*) presents considerable evidence supporting the closed loop characteristic of a multipactor breakdown as shown in Fig. 3.

Since frequency and dimensions were physical constants of the components, the power level was varied in the RF breakdown testing program. The effect of the other factors affecting voltage breakdown characteristics have not been treated here but are available in Refs. 5, 6, 7, and 8 (*Bibliography*).

A theoretical multipactor discharge is purely reactive with the electron current in the breakdown gap lagging the applied voltage by 90°. However, the practical results and implications of a space charge in which a swarm of electrons are under the influence of the applied electric field alters the ideal situation. The net effect of the space charge changes the breakdown so that it appears as a complex impedance with a resistive and reactive component. The resistive portion of the breakdown implies power dissipation and localized heating.

BREAKDOWN INVESTIGATION

The breakdown investigation included the heating aspect of the multipactor discharge just mentioned by placing thermocouples in areas suspected of supporting any RF breakdowns (see Fig. 4). Power meters were employed in the input RF line to detect the change in characteristic impedance of the 50-ohm lines caused by a breakdown. The RF power meters employed in the output lines detected changes in RF power passing through the device under test.

The RF source consisted of a CW magnetron and a regulated power supply. The control circuitry of the power supply enabled the operator to set the magnetron at any RF power level throughout a continuous range between 40 and 200 watts. The circulator provided over 20 db isolation between the magnetron and the breakdown so that excessive power levels would not be reflected back to the magnetron. Both input and output RF power levels were continuously recorded so that correlation between these two parameters was always available.

The time resolution of the RF power meters and the strip recorder was limited to several milliseconds because of the mechanical nature of these devices. Therefore, an oscilloscope was connected to the output of a crystal detector and was visually monitored to detect the presence of shorter duration breakdowns. The oscilloscope increased the time resolution of the test configuration to 0.1 μsec. This test configuration possessed the inaccuracies prevalent in any microwave measurement; but, ability to detect changes in power level was present and constituted the salient feature of the measurement. The resolution of the RF power level was 0.1 db per scale division for both the recorder and the oscilloscope.

The physical arrangement of this test configuration is illustrated in Fig. 5. The lead shielding was necessary to reduce the 1-Mev gamma radiation of the cobalt-60 to safe levels in the working areas.

The testing program, utilizing the equipment set up of Fig. 4 and 5 included the following general procedures. The test specimen was placed in the bell jar, and the test set-up was calibrated at room ambient. The cobalt-60 was positioned, and a vacuum of 10^{-4} torr or less was established. The test specimen was allowed to "soak" in this environment for 4 hours. At the end of this interval, input power was applied and was varied through 12 discrete steps (as illustrated in Fig. 6) searching for multipactor discharges sensitive to power level.

The power level incident on the test specimen ranged between a low value of 40% under the system level and high of 100% overload. The power level interval between zero and 40% under system level was "explored" by successful equipment turn on.

The ascending power profile (Fig. 6) was interlaced with the descending power sequence to realize the 12 different power levels. The oscilloscope was carefully monitored for RF dropouts during each of the changes in power level. The close observation during power level transistions was necessary to ensure the absence of narrow power level, multipactor breakdowns.

TEST RESULTS AND APPROACHES TO RE-DESIGN

This test series revealed extensive deficiencies in two general areas: 1) RF connectors used on the RANGER flight cables, and 2) the feedthrough connectors used on the bell jar walls as part of the test apparatus. Although the malfunctions occurred during these tests, the breakdowns could not be positively identified

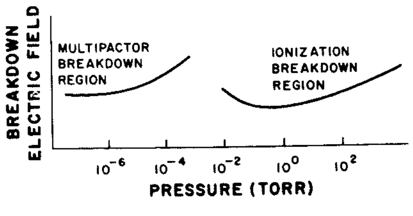


Fig. 2—Multipactor and ionization breakdown.

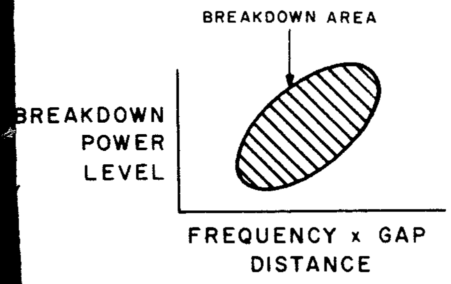


Fig. 3—Multipactor breakdown closed loop characteristic.

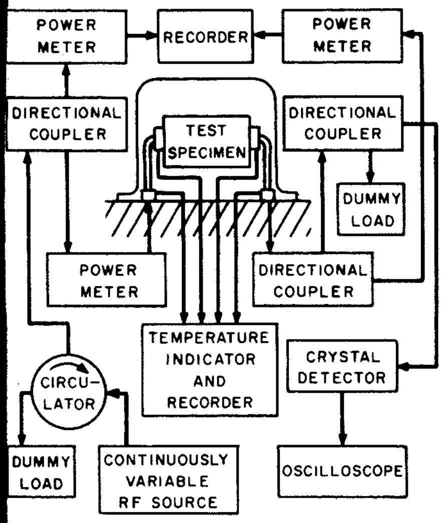


Fig. 4—Breakdown test configuration.

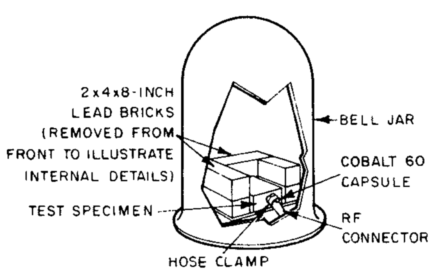


Fig. 5—Physical arrangement of test specimen and cobalt 60 capsule.

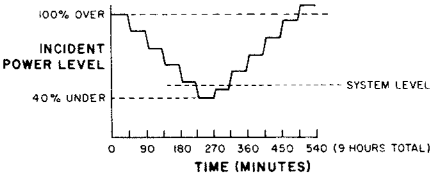


Fig. 6—Multipactor breakdown incident power profile.

as multipactor discharges because of the inability to establish the true pressure level in the breakdown gap.

The flight RF connectors were of two manufactured types of the TNC series. As shown in Fig. 7, the two types of connectors possessed dielectric interfaces and voids which offered a "straight line" breakdown path between inner and outer conductors. Considerable blackening was also observed in these areas when the connectors were disassembled.

Connectors of these types experienced RF voltage breakdowns at power levels as low as 40 watts at the RANGER frequency of 960 Mc. This breakdown power level represented a deficiency of 30% below the system operating level. Since it was desirable to increase the power rating of the connectors beyond the system level of 60 watts, a connector redesign was initiated.

Recognizing the inability to eliminate voids between cables and connector dielectrics, the connector redesign utilized well established techniques of high voltage design. (Before the redesign was complete, however, it was discovered that TNC connectors of this type were commercially available.) These design techniques included the overlapping dielectric interface modification illustrated in Fig. 8. As shown, the breakdown path was forced "around corners" with a resulting increase in path length. The dielectric void was designed to be in the lower electric fields near the outer conductor. This modified connector-cable assembly was tested at power levels up to 140 watts without detectable breakdowns.

The connector RF breakdown problem outlined above represents a general class of problems associated with all interfaces between coaxial microwave components. This problem area usually arises from the method of manufacture which provides for the fabrication of the microwave device to which connectors are attached. To circumvent this problem, a philosophy which includes the connector in the device design is required and in the most obvious instance would extend a monolithic dielectric throughout the device and into the connector mating area. The need for the solid dielectric, as provided by a type TNC connector, is underscored by the following.

Air-dielectric, female N-connectors were employed on the inside and outside of the bell-jar vacuum test apparatus in the form of female feed-throughs. The first detectable breakdowns were traced to the female N-connector on the vacuum side of the bell jar. This malfunction occurred at varying power levels between 35 and 55 watts and necessitated replacing the vacuum-side female N-connector with a TNC-connector. With the installation of the new type N-to-TNC feedthroughs along with the newly designed TNC flight connectors, all test and flight connectors exposed to vacuum were of the TNC variety with solid dielectrics, and connector malfunctions disappeared.

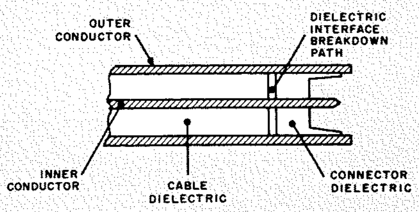
The remaining RF components were successfully cycled through this test configuration and were reinstalled on the spacecraft. However, RF power dropouts persisted in subsequent retesting of the assembled RANGER communications and TV subsystem. The magnitude of the dropouts varied from several tenths of a db to near one db, as initially observed, but were reduced in frequency.

Since the four-port hybrid was the only unpressurized air-dielectric, microwave device remaining on the spacecraft, all activity centered on this component in a re-examination of the problem.

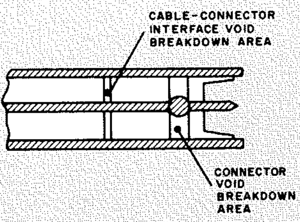
This device was a purchased component of conventional design consisting of a rigid, rectangular, cross sectional center conductor supported by teflon beads within a hollow rectangular annulus. The annular ring comprised the mechanical package and outer conductor as an integral part. Four TNC connectors attached to the ring section provided the means of making external connections. A typical connector-ring cross section viewed from the center of the ring appears in Fig. 9.

SUMMARY OF TEST CONCLUSIONS

Before entering into a description of the successful effort surrounding the four-port hybrid, a summary of the important results and conclusions of the testing program up to this time is beneficial. First, it was found necessary to perform fully instrumented breakdown retests on individual components. Second, because of the mission profile which required high-powered communications at the end of 66 hours, all breakdown retesting had been conducted at vacuum levels near or below 10^{-4} torr. The retests were performed with the knowledge that a vacuum level of 10^{-4} torr was sufficiently low to safely insulate against ionization breakdowns but would permit multipactor breakdowns. This led to the finalization of a multipactor testing procedure which uncovered connector deficiencies along with a marginal performance of the purchased four-port hybrid. The marginal performance of the latter component was of the form of small, infrequent power dropouts of several tenths of a db separated in some instances by hours.



TYPE A



TYPE B

Fig. 7—RANGER flight RF connectors.

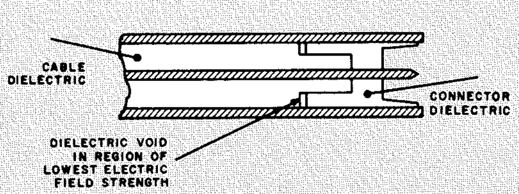


Fig. 8—RANGER connector designed to impede RF voltage breakdown.

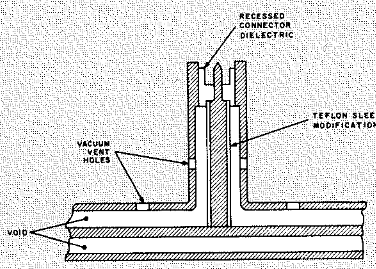


Fig. 9—Four-port hybrid connector cross section.

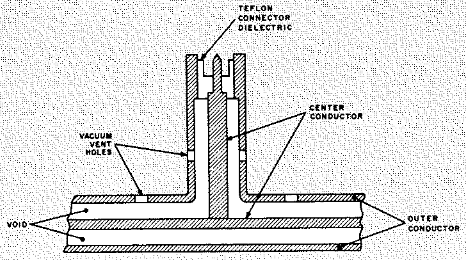


Fig. 10—Solid-void dielectric modification of four-port hybrid.

MODIFICATIONS TO THE FOUR-PORT HYBRID

The continuation of RF breakdowns within the four-port hybrid resulted in the following modification. The alteration consisted of the insertion of a 0.020-inch teflon sleeve around the center conductor as illustrated in Fig. 10. Although most insulators are good secondary emitters and the emission properties of teflon were unknown, this modification was completed in an attempt to disrupt the "resonance" of a multipactor discharge by 1) altering the electric field strength and 2) changing the transit time between boundaries. Further multipactor testing of the modified four-port hybrid revealed the absence of RF voltage breakdowns, and the integration of the RANGER spacecraft proceeded.

This modified four-port hybrid became the object of the intense reexamination of the RF breakdown problem mentioned earlier. It now was obvious that there was a deficiency in the multipactor testing which manifested itself in the form of RF breakdowns in simulated missions of an assembled spacecraft.

An examination of the device's characteristics, accumulated from the many hours of tests performed on an isolated four-port hybrid and on an integrated four-port hybrid, led to the following conclusions: The 0.5-db total insertion loss of the four-port hybrid required approximately 8 watts dissipation within the four-port hybrid with both the transmitters on. This level of power dissipation, it was reasoned, could lead to internal surface heating with subsequent release of trapped gases. If this were the case, the presumption of achieving a safe operating vacuum level was obviously in jeopardy. In addition to this, a question was raised about the outgassing efficiency of a small component, such as the four-port hybrid, while immersed in the atmosphere of a large space vehicle. The spacecraft might, in effect, be outgassing into the four-port hybrid. Verification of this condition would explain the successful operation of an isolated four-port hybrid and marginal performance of an integrated four-port hybrid.

To determine the feasibility of the existence of a dangerous outgassing condition, the four-port hybrid was placed in the test configuration used in the multipactor investigation. Now, however, the four-port hybrid was required to operate through a cycled pressure environment as illustrated in Fig. 11. This new requirement exposed the four-port hybrid to the hazards of a partial vacuum, ionization breakdown centered about the critical pressure of 1 torr depicted in Fig. 2. The term *critical pressure* is used here to indicate the pressure level that coincides with the minimum breakdown voltage or power level that exists in the vicinity of 1 torr.

The dashed line in Fig.11 represents a typical pressure profile achieved in actual test. Care was exercised to prevent rapid change through the critical pressure area, but this was not as acute as might be expected. The very high incident power of 120 watts had the effect of broadening the range of pressures at which breakdowns would occur. This was underscored by the erratic performance obtained with the four-port hybrid through pressures between 100 torr and 10^{-2} torr.

With 120 watts applied to each input connector, the modified four-port hybrid exhibited sustained breakdown over a 100-torr range centered about 1 torr. Moreover, it was determined that the four-port hybrid suffered from ionization breakdown at incident power levels as low as 14 watts at 960 Mc.

The above disclosures illustrated the need for a new design eliminating the air dielectric of the existing four-port hybrid. The success achieved with the solid dielectric, coaxial connectors naturally led to the consideration of a solid dielectric, coaxial four-port hybrid. To impose a solid dielectric in the existing four-port hybrid required a change in the conductor radii to maintain the 50-ohm characteristic impedance. Therefore, a more suitable means of achieving the solid dielectric appeared to be in a newly designed stripline device. However, little was known about the maximum RF power capability of a stripline device operating through the required vacuum environments.

In an effort to answer these questions and with the apparent advantages of stripline as an incentive, a new design was completed. Particular attention was focused on details to insure the exclusion of breakdown possibilities. Since the stripline represented a homogeneous, solid dielectric transmission line which formed the ring or rat-race portion of the four-port hybrid, it was inherently immune from breakdowns. However, the attaching TNC connectors represented a potential trouble spot. To eliminate the possibility of breakdowns forming along the interface between the connector and stripline dielectrics, a dielectric overlap technique similar to that successfully employed in the connector redesign of Fig. 8 was used. This was accomplished by forming a recess in the stripline dielectric which received the connector dielectric, as illustrated in Fig. 12.

Another potential trouble area was

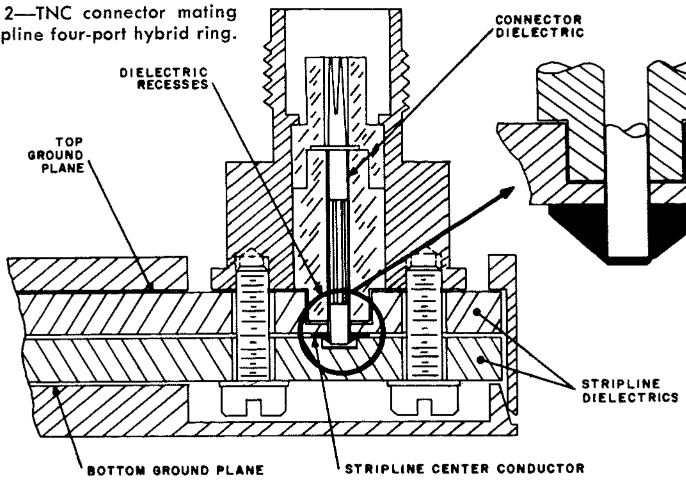


Fig. 12—TNC connector mating in stripline four-port hybrid ring.

the method of connecting the connector center-conductor to the stripline center-conductor. A suitable connection (Fig. 12) was realized by extending the connector center conductor beyond the stripline center conductor to facilitate inclusion of a solder fillet.

The completed stripline four-port hybrid (Figs. 13 and 14) was first subjected to the ionization breakdown test at incident power levels of 120 watts. This test revealed no thermal problems along with a noticeable absence of RF voltage breakdowns. Also, no RF dropouts were detected during the course of a 9-hour multipactor test in which ambient test chamber temperatures were raised to 55°C. Thus, the stripline four-port hybrid ring eliminated the breakdown problem in this component.

Since the multipactor testing techniques had failed to detect all breakdown possibilities of the coaxial four-port hybrid, all components that successfully passed this test were subsequently exposed to the cycled pressure environment of the ionization breakdown tests. It was determined through these two tests that no RF dropout problems existed in the other components utilized in the 60-watt RF lines. Thus, it was evident that the two pressurized 60-watt power amplifiers, the pressurized 50-ohm dummy load, the solid dielectric cable assemblies, and the solid dielectric four-port hybrid ring possessed the ability to operate through all vacuum environments. This ability fulfilled the RANGER mission requirements as was demonstrated in simulated missions of the TV and communications subsystem, of the spacecraft in its final integration tests prior to launch, and finally in the successful operational mission of RANGER VII.

CONCLUSION

Two important results emerge above all others as a result of the microwave communications effort on the RANGER program. The first and the most obvious is the recognition of the RF voltage breakdown problem itself, and the establishment of an adequate design and testing program to combat the problem.

The second and more subtle result of this effort is the contradiction that may arise between the planned spacecraft mission and the device design. The planned mission usually dictates the component design. Although the TV and communications subsystem of RANGER did not have to operate through the launch environment, successful operation in the space environment was not attained until the four-port hybrid achieved the capability of successful operation through the launch environment.

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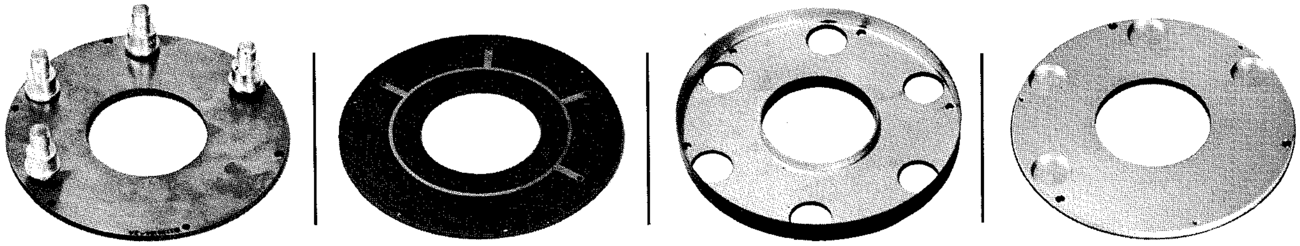

Fig. 13—Components of stripline four-port hybrid.

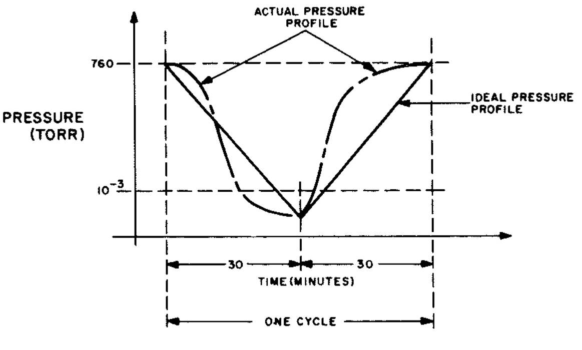

Fig. 11—Cycled pressure environment of the ionization breakdown test.

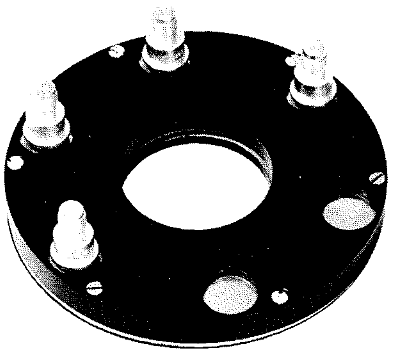

Fig. 14—Stripline four-port hybrid.

RELAY SATELLITE COMMUNICATIONS SYSTEM

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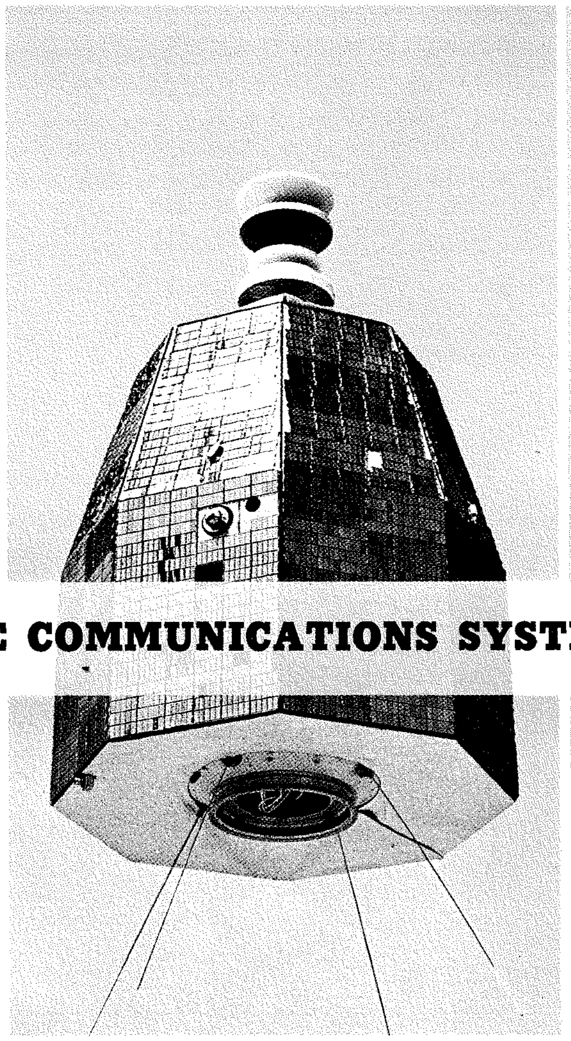

Fig. 1—RELAY satellite.

Long-distance, high-capacity communications systems may be established using microwave repeaters in earth-orbiting satellites. The problems associated with the operation of these repeaters in the environment of outer space have been explored using the RELAY satellites. In this paper, the performance objectives and design criteria for the repeaters in the RELAY satellite are given, and the ground-station network is described. It is shown that the ground station network, the launch and orbiting constraints, and the outer-space environment impose severe restrictions upon the design of this satellite system. The experience with the RELAY satellites has shown, however, that a proper consideration of these restrictions, in addition to careful processing and testing, can result in long-lived performance in the space environment.

Prior to 1961, space programs like Score and Courier had successfully demonstrated the feasibility of incorporating satellites in long-range communications systems. In the summer of 1961, RCA was selected by the National Aeronautics and Space Administration to build the Relay satellite for use in another experimental communications project.

The heart of the communications system in the Relay satellite was to be an active repeater which would receive and retransmit communications signals between six ground stations located in the United States, Europe, and South America. Communications signals to be evaluated were an assortment of television signals, multichannel telephony, and other communications that might be of interest.

Presently, two Relay satellites are in orbit. Relay I was launched on December 13, 1962, and Relay II on January 21, 1964. The final configuration of these satellites is shown in Fig. 1.

SYSTEM DESIGN CONSIDERATIONS

The high cost of launch vehicles generally results in a system which consists of a simple low-powered satellite of high reliability and highly complex ground stations. The Relay program has followed this logic. The Relay ground stations use low-noise receivers, high-power transmitters, and large steerable antennas. The Relay satellite, on the other hand, is weight- and power-limited with the design emphasis placed on simple circuitry and a minimum of parts. Thus, in the repeater the receiver input stage is a conventional mixer, the transmitter power is 10 watts, and the antenna is nonsteerable with an essentially omnidirectional radiation pattern.

As in many other communications systems which use line-of-sight transmission, FM was the method of modulation selected for the Relay system. The dy-

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namic range of the RELAY satellite is 40 db. This range accommodates not only large and small ground stations, but also the ellipticity of the orbit. The perigee of RELAY I is 800 miles, and the apogee is 4,500 miles.

Thermal Noise

Thermal noise is introduced into the communication path at the front-end of the satellite receiver and ground-station receivers. The satellite-to-ground portion of the communication path determines the total system noise, since this is where it is most difficult to improve the system. The principal limiting factor is the effective radiated power of the satellite transmitter. The use of a simple front-end in the satellite receiver in conjunction with ground-transmitter powers of from 1 to 20 kw generally ensure that the satellite-to-ground path contributes most of the noise.

The parameters of the satellite-to-ground path must be adjusted to satisfy several conditions. First, the demodulated signal-to-noise ratio must be adequate to satisfy the objectives of telephony, television, and record service. The service quality must be maintained in spite of varying conditions such as slant range, satellite antenna gain versus look angle, tracking losses, precipitation, etc. Satisfactory service quality requires that the demodulating equipment remains above threshold during these varying conditions. A conventional FM limiter-discriminator combination has a threshold of 12 to 14 db. For other methods of demodulation, such as FM with feedback or phase-locked loops, the threshold value will improve. The Communications Committee on International Radio (CCIR) recommends a telephone signal-to-noise ratio of 50 db; however, they do permit severe weather conditions to degrade the service quality provided the frequency of occurrence is not too high.

With the parameters of the satellite-to-ground path established, it is necessary to insure that the ground-to-satellite path does not significantly increase the overall noise limits. The noise contribution of the ground-to-satellite path can be expressed by:

$$T_e = T_g\left(1 + \frac{T_s P_s}{T_g \frac{P_g}{N^2}} K\right) \qquad (1)$$

where: T_e is the overall noise temperature of the system, T_g is the overall noise temperature of the ground station, T_s is the noise temperature of the front end of the satellite receiver, P_s is the power delivered to the antenna of the satellite transmitter, P_g is the power delivered to the antenna of the ground transmitter, N is the multiplication index of the satellite, and K is a factor determined by certain system conditions.

For the RELAY repeater, the multiplication index N is 3; in a linear repeater, it would be 1. This factor modifies power (P_g/N^2) in order to show the penalty incurred in the effective ground transmitter power by index multiplication in the satellite. If the ground transmitter had adequate bandwidth (and did not limit the system bandwidth), the same overall performance could have been achieved with a transmitter power $1/N^2$ as large, provided that all other factors had remained the same.

The K factor is used to account for a variety of situations. The value of K is one if the slant ranges for both paths (to and from the satellite) are equal, if the ground antenna(s) has equal gains at equal frequencies, and if the satellite receiving and transmitting antenna gains are essentially isotropic. If these conditions are not fulfilled, then the value of the K factor must be adjusted accordingly.

Typical noise-determining parameters for a large ground station and the RELAY satellite are:

$T_g = 50°\text{K}$ $\qquad P_g = 10$ Kw
$T_s = 6{,}960°\text{K}$ $\qquad N = 3$
(noise figure of 14 db) $\qquad T_e = 112.1°\text{K}$
$P_s = 10$ watts $\qquad K = 1$

Phase Distortion

With FM, the information is contained in the phase of the carrier wave. The combined phase characteristics of the satellite repeater and the ground equipment determine the amount of signal distortion introduced in the system. The important phase characteristic, called *group delay* (or *differential time delay*), is defined as the rate-of-change-of-phase with frequency. If the group delay is frequency-dependent, signal distortion will result. The level of distortion will depend on the group delay and the characteristics of the signal. For example, if the frequency-dependent group delay is large, ringing and other distortions will occur in a television picture. In the RELAY system, the important distortions caused by group delay are 1) video-to-audio crosstalk, and 2) intermodulation distortion in multichannel telephony.

The video-to-audio crosstalk is important because the audio portion of a TV program is carried on a 4.5-Mc subcarrier located just outside the 3.5-Mc video band. Group delay causes modulation of subcarriers by components in the video spectrum, principally the TV line rate. The result is spurious modulation in the audio channel. In the RELAY satellite, group delay is sufficiently small so that the video-to-audio crosstalk is about 60 db below the audio test tone.

Intermodulation distortion in multichannel telephony is caused by group delay. In multichannel telephony, the individual voice channels are combined to form a composite frequency-division-multiplex baseband. The group delay causes harmonics and difference-frequency components to be generated in the baseband. The resulting intermodu-

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J. Kiesling

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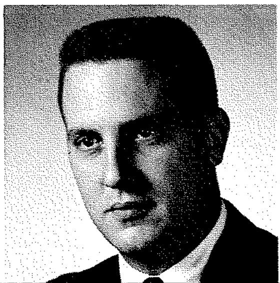

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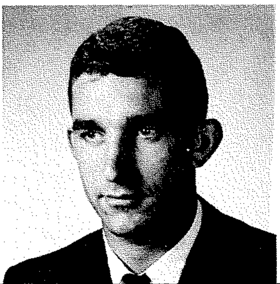


lation noise is similar to white noise provided there are many contributing channels. The intermodulation distortion affects the quality of the system just as if it were thermal noise. Fortunately, this type of distortion can be reduced through phase equalization of the system path. The total intermodulation distortion for 300 channels allowed in the RELAY system is 7,500 pw. The RELAY satellite contributes about 2.000 pw of intermodulation noise in the wideband mode. In the narrowband or two-way mode. the intermodulation noise contribution of the RELAY satellite is negligible.

Coherent Crosstalk

Another type of crosstalk is due to FM-AM-PM conversions. This form of crosstalk occurs if more than one modulated carrier is present simultaneously in the satellite during operation in the two-way mode. The generation of crosstalk involves the conversion of frequency variations into amplitude variations by imperfect gain flatness of the equipment; i.e., circuit gain is not constant with frequency. If this incidental amplitude modulation is present in a device whose phase response is amplitude-dependent, then the modulation on one carrier is transferred to the other carrier(s), and coherent crosstalk results. Crosstalk between the two narrowband channels in the RELAY system is typically 50 db.

COMMUNICATIONS REPEATER

The communications repeater, as finally evolved, is comprised of the major components shown functionally interconnected in Fig. 2. Figs. 3, 4, and 5 illustrate the mechanical configurations of the major components.

Basically, the operation of the repeater is as follows: A television signal at 1,725 Mc or dual-carrier telephony signals at 1,723.33 Mc and 1,726.67 Mc are received at the satellite receiving antenna at levels from —40 to —80 dbm (Fig. 2). The signal(s) is split by means of a 3-db hybrid, half the power being fed to each receiver. The *on* repeater first down-converts the signal to 70 Mc, then amplifies it, then triples its frequency deviation, and finally up-converts the signal to 4,170 Mc at a level of 5 to 7 mw. The output from the up-converter and a CW microwave-tracking beacon at 4,080 Mc are fed into a 10-watt TWT that operates at saturation with a gain of 35 db. The TWT output is transmitted to the ground station by means of the transmitting antenna.

Fig. 2—RELAY repeater, showing redundancy of the configuration.

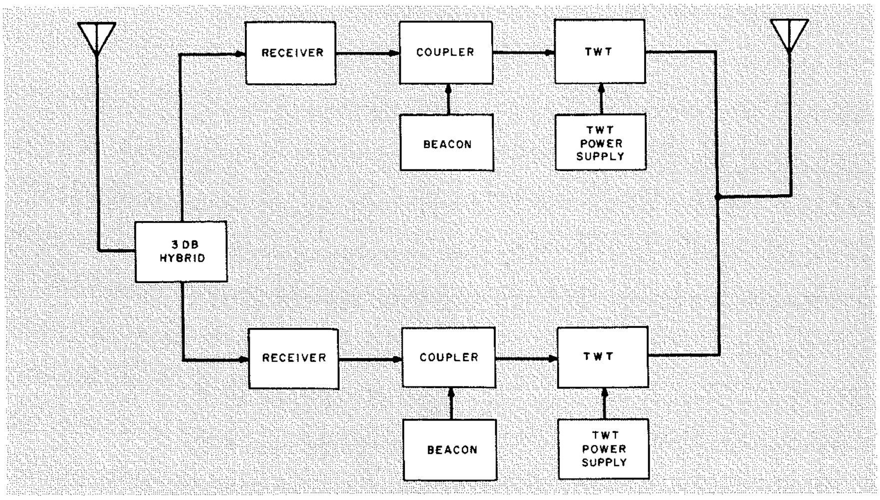


Fig. 3—Wideband receiver, beacon, and coupler.

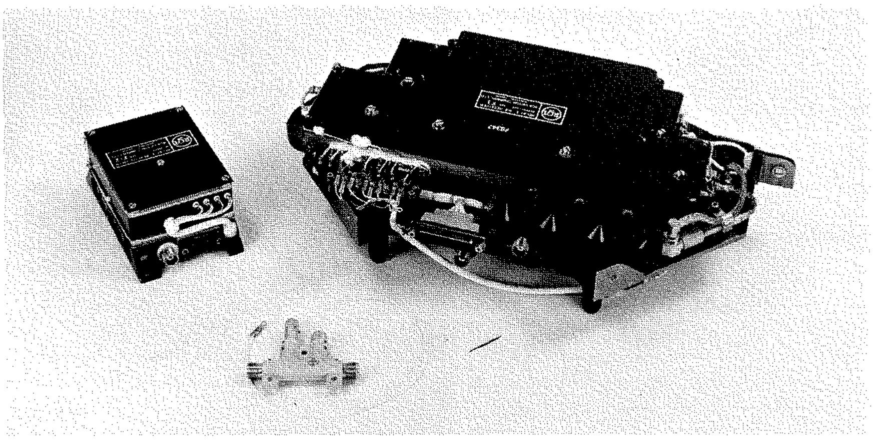


Fig. 4—Traveling-wave tube and its power supply.

Receiver

The receiver is a solid-state heterodyne-type circuit providing high efficiency, temperature stability, and high reliability. Fig. 6 shows the subunits included in each of the receivers.

The incoming signal to the receiver is routed through a bandpass filter which acts as a preselector. The 1-db bandwidth of 27 Mc is achieved by three tuned, mutually coupled transmission-line cavities. The filter insertion loss is 0.4 db. The filter bandwidth is made much wider than the maximum signal bandwidth of 13 Mc to minimize group delay. The purpose of the preselector is to reject undesired and interfering signals. The output signal of the input filter is coupled through a coaxial impedance transformer to a single-ended low-noise crystal mixer (down-converter).

Although certain frequency restrictions required the use of two independent local oscillators (one for the receiver and one for the transmitter), the frequency scheme was selected primarily on interference considerations. At the same time, an attempt was made to minimize local oscillator multiplier stages by the use of quadruplers and triplers, rather than doublers, for the generation of the required microwave frequencies for up- and down-conversion. For example, the receiver local-oscillator chain includes two quadrupler stages, and the transmitter local-oscillator chain uses a quadrupler and two tripler stages.

Both varactor multipliers in the receiver local-oscillator chain are self biased, with the second varactor mounted in a cavity tuned to 1,655 Mc. The transmitter local-oscillator chain, which provides a higher-power output, uses fixed bias stages, with the last two varactor multipliers mounted in separate cavities.

The local oscillator stages of the transmitter and the receiver use a grounded-base transistor and a fifth-overtone quartz crystal. Both oscillators have a frequency stability of better than 2 parts in 10^5 over the temperature range.

The output signal of each local oscillator stage is amplified in subsequent amplifier stages. The receiver local-oscillator stage delivers 10 mw of power to a Class-A buffer amplifier which supplies 150 mw of drive to the multiplier chain. The receiver local-oscillator chain delivers about 15 mw at 1,655 Mc to the down-converter. The 1,655-Mc pump signal from the receiver local-oscillator chain is injected into the down-converter by means of a capacitive probe which is adjusted to obtain 0.5-mamp crystal current. A quarter-wavelength choke at the output of the down-converter rejects RF and local-oscillator signals, and passes the 70-Mc difference frequency to the main IF amplifier. The mixer conversion loss is 5.5 db.

The IF amplifier uses AGC to maintain a constant output level of 100 mw ±0.5 db over the input dynamic range of −49 to −89 dbm. An inductive *T* network matches the impedances of the mixer output stage and the first low-noise IF stage, and limits the input noise bandwidth to 30 Mc. The IF amplifier consists of two low-noise common-emitter stages (with an overall gain of 20 db and a noise figure of 4 db), eight stages with AGC, a driver, and a power amplifier. The overall gain is between 69 and 109 db, and the minimum 1-db bandwidth is 16 Mc. The overall receiver noise figure, up to and including the hybrid, is 14 db. The first six stages with AGC are common-emitter stages that are video-tuned and RC-coupled. Each stage incorporates selective, degenerative collector-to-base feedback to improve response. The remainder of the stages are conventional synchronously tuned amplifiers. The video-tuned stages and synchronously tuned stages were selected in preference to stagger-tuned stages and other types in order to minimize the AM-to-FM conversion.

A portion of the output of the IF power amplifier is detected and amplified to derive the AGC voltage applied to the bases of the controlled IF amplifier stages. A control voltage, derived from the AGC voltage, turns off the repeater when no input signal is present for a continuous period of 3 minutes.

The output of the IF amplifier is fed to the IF tripler subunit which operates in one of two modes as determined by the command-controlled IF switch. For one-way television or 300-channel telephony, the signal is routed through a wideband tripler and adder to the up-converter. For 24-channel two-way telephony, the two signals are separated by filters, fed into separate limiting amplifiers and triplers, and recombined in the adder. The separate signal paths in the narrowband (two-way) mode provide for individual level settings and avoid the spurious-signal generation of a single tripler. If two signals of equal amplitude at frequencies of ω_1 and ω_2 were fed into a single tripler, the output frequencies would be unwanted frequencies which can be filtered (ω_1, ω_2, $2\omega_1-\omega_2$, and $2\omega_2-\omega_1$), unwanted frequencies not easily filtered ($2\omega_2+\omega_1$ and $2\omega_1+\omega_2$), and desired signals ($3\omega_1$ and $3\omega_2$). The $2\omega_2+\omega_1$ and $2\omega_1+\omega_2$ frequencies which cannot be filtered easily are higher in level than the desired frequencies and are not usable because they contain information from both input signals. Instead, the use of a separate narrowband signal path for each channel allows the use of amplitude limiting in each signal path, permitting a 10-db level difference to be accommodated between the two received carrier levels for equal carrier level output. This simplifies ground-station operation by eliminating the need for adjusting the output level of the ground transmitter, which is inconvenient for intermediate-altitude satellites.

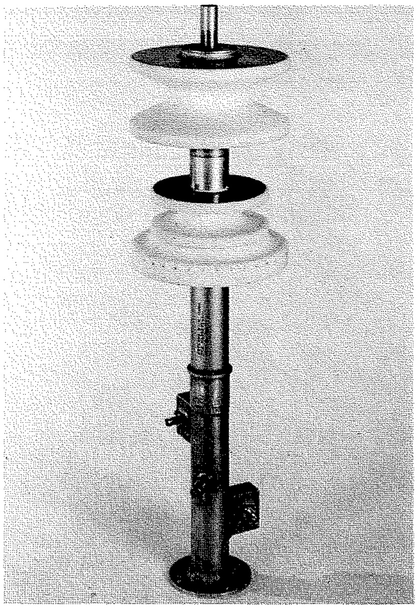


Fig. 5—Antenna assembly.

The narrowband bandpass filters are similar, except for a center-frequency difference of 3.33 Mc. An RLC network is used to match the parallel 20-ohm characteristic impedance of the filter combination to the 75-ohm output impedance of the IF amplifier. This type of matching network was used rather than a mutually-coupled inductive transformer in order to maintain a low VSWR outside the passband of the filters to reduce reflections of IF noise back into the IF amplifier. A 19-db minimum return loss was achieved with this network, thus minimizing undesirable AGC interaction. Each filter is of a modified image-parameter design which uses a toroidal form inductance. Because component density is high, mutual coupling is minimized by individual filter-section shielding partitions and the self-shielding properties of the toroidal coils. The bandwidth of each filter is 2 Mc, and the "skirts" afford 40 db of attenuation at the center frequency of the opposite channel.

Each limiting amplifier consists of two common-base tuned amplifiers connected in cascade. The first stage, a driver, operates in a nonlimiting mode to make up for the high insertion loss of the filter. The second stage maintains a con-

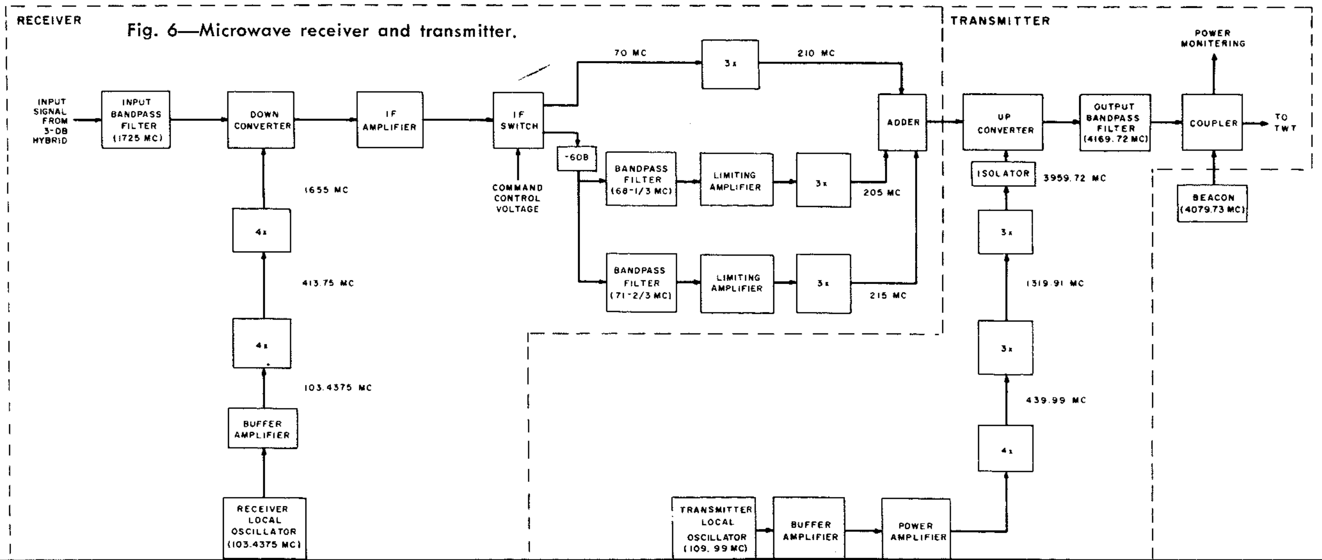


Fig. 6—Microwave receiver and transmitter.

stant output over a 10 db range of input power.

The wideband signal path and both narrowband paths have separate triplers. Except for center frequency and bandwidth, the triplers are essentially the same; therefore, only the wideband tripler is discussed. Originally, varactor triplers were used to reduce power consumption. However, after extensive testing, they were found to be unstable over the required temperature range, and were replaced by less efficient transistor triplers. All triplers are of the common-base configuration, employing a pi-input matching network. The wideband tripler matching network is tuned for optimum return loss, as was required of the narrowband filter-matching network. The output of each tripler is tuned to three times the input IF frequency, and is coupled into the adder stage.

The adder consists of two independently driven common-base amplifiers coupled together at their collectors. It serves as an impedance-matching network between the triplers and the up-converter and provides a means of combining the narrowband signals. At the same time, it provides at least 16 db of isolation between the two narrowband triplers and provides a 1-db bandwidth of 25 Mc. The adder drives a double-tuned circuit, which resonates with the capacitance of the adder transistors and the capacitive input impedance of the up-converter.

Transmitter

The subunits of the transmitter are shown in Fig. 6. The transmitter local-oscillator signal and the IF output signal from the adder are mixed and converted to 4,170 Mc. This microwave output signal is amplified in a TWT power amplifier and is transmitted to the ground stations.

The transmitter local-oscillator stage delivers 40 mw at approximately 110 Mc to a two-stage amplifier which drives the multiplier chain at a 1.2-watt level. The transmitter local-oscillator chain supplies 75 mw at 3,960 Mc to the up-converter through a ferrite isolator.

The up-converter uses a balanced phase-shift modulator. Two pill-type varactors are located on opposite ports of a stripline hybrid, tuned to the frequency of the transmitter local-oscillator chain (see Fig. 7). The pump signal is coupled into the hybrid through a ferrite isolator. The output signal is provided at the opposite port. The varactors are switched in a push-pull manner by the tripled IF signal from the adder. With one varactor in the forward biased state, the corresponding hybrid port presents a short-circuit condition to the oscillator frequency because of the action of the quarter-wavelength choke. Thus, the local-oscillator signal is forced to flow through the opposite hybrid leg to the output. When the varactor is reverse biased during the next half of the IF cycle, the varactor capacitance resonates with the inductive stripline stub at the local-oscillator frequency and allows the local-oscillator signal to flow to the output. In this manner, the oscillator signal is phase shifted ±90° at the tripled IF rate. Using Fourier analysis, the output waveform is expressed by the equation:

$$V_o(t) = A_L\left[\frac{2}{\pi}\sum_{n=1,3,5\ldots}^{\infty}\frac{1}{n}\sin(\omega_L + n\omega_S)\,t - \sin(\omega_L - n\omega_S)\,t\right] \quad (2)$$

Where: A_L is the amplitude of the pump signal, ω_L is the frequency of the pump in radians/sec, and ω_S is the tripled IF frequency in radians/sec.

Letting $n = 1$, Eq. 2 reduces to:

$$V_o(t) = \frac{2}{\pi}A_L\,[\sin(\omega_L + \omega_S)\,t - \sin(\omega_L - \omega_S)\,t] \quad (3)$$

Eq. 3 shows that the oscillator signal is isolated from the output. The desired output at 4,170 Mc ($\omega_L + \omega_S$) is ideally 3.9 db (20 log $2/\pi$) below the pump level. Since the up-converter is terminated in a bandpass filter tuned to the upper sideband, the available power output is increased by making use of the second harmonic of the pump generated in the up-converter. By varying the length of the line between the up-converter and the filter, the phase of the signal produced by the frequency difference between the $2\omega_L$ frequency and the reflected out-of-band lower-sideband ($\omega_L - \omega_S$) frequency is adjusted for optimum addition to the primary ($\omega_L + \omega_S$) frequency.

The up-converter produces a variety of unwanted signals. The output filter is designed to pass the upper-sideband ($\omega_L + \omega_S$) signal over the required 25-Mc, 1-db bandwidth and to reject the unwanted signals. The filter consists of three slug-tuned cavities which are mutually coupled and supplied with inductive type input- and output-coupling loops. The 20-db bandwidth is 110 Mc, with a nominal passband insertion loss of 0.4 db.

The coupler provides a means of coupling the microwave communications signal and a tracking beacon output into the TWT for simultaneous amplification. It is a coaxial directional coupler, in which the microwave signal is fed straight through and the beacon is fed through the coupling arm which has a directive inductive coupling loop. The coupling arm is adjusted to provide 10-db coupling and a minimum of 10-db directivity.

A TWT was selected as the output power amplifier of the repeater because it provides a satisfactory balance between efficiency, power, bandwidth, gain, low distortion, weight, ruggedness, and reliability. The tube was designed by RCA to operate in the space environment and to meet the following basic performance requirements:

Frequency Range	4,050 to 4,250 Mc
Saturated Output Power	11 watts
Saturated Gain	35 db
Overall Efficiency	21%
Weight	4.5 pounds

The ratio of helix diameter to electron-beam diameter is designed as small as possible for the optimum conversion of DC to RF power. The helix is embedded inside a fluted-glass vacuum tube of small diameter. The small helix diameter allows the use of: (1) helical RF input- and output-couplers which are inserted through the vacuum-tube envelope, 2) external RF attenuators which eliminate possible dissipation and outgassing problems, and 3) periodic focusing magnets which are small and light in weight. The combination of a small helix diameter and a small helix-to-beam diameter ratio requires a narrow beam to avoid excessive helix interception and resulting inefficiency. For this reason, the electron gun must supply a good laminar beam from a relatively small cathode so that beam convergence is not too high for sharp magnetic focusing of the beam. Although a small cathode requires less heater power, the cathode depletion is greater and the beam-current density is higher, thus increasing the operating temperature and decreasing the tube life. Because of this problem, a cathode of barium-strontium car-

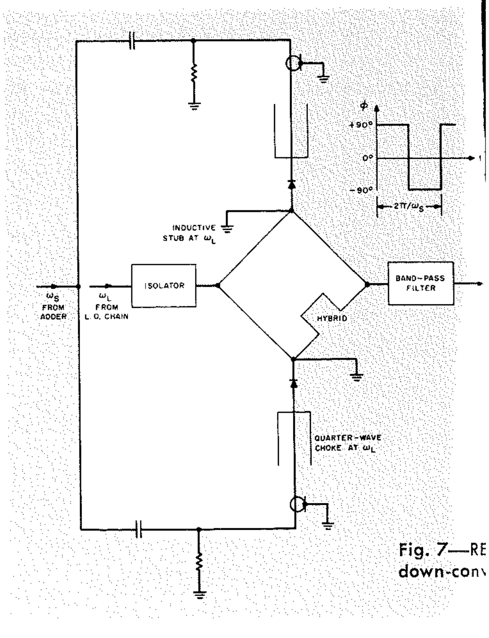


Fig. 7—RE
down-con

bonate over base metal N132 was designed for very low beam-current density (55 ma/cm^2) and low cathode operating temperature, with beam-shaping characteristics adequate for sharp focusing. The low operating temperature slows down the rate of depletion of the emissive cathode material, further ensuring long tube life. The electron gun, similar to the Pierce convergent type, uses an efficient dark heater and supplies a beam with an excellent laminar flow.

Since the major source of heat in the tube is the 35 watts of power dissipated in the collected beam, the collector is electrically operated at spacecraft ground, enabling good thermal conduction to a thermal radiator and the spacecraft heat sink. Originally, it was intended that the tube components outside of the vacuum envelope would be operated in a nonpressurized enclosure. However, random power-output perturbations were revealed during early vacuum tests on the tube. The TWT development program was then modified; a pressurized tube was designed and qualified for the RELAY program, and two of these tubes were flown on RELAY I. Simultaneously, work was performed to eliminate the need for the pressurization. Part of the difficulty resulted from the ionization of gas accumulated between the helical RF output coupler and the glass envelope. Another difficulty was due to an induced charge on the glass bulb caused by the 1,100-volt potential difference between the helix and the grounded tube capsule. As outgassing occurred, the small amount of accumulated gas was ionized by the potential difference between the glass bulb and the RF output coupler. Power output fluctuations were eventually eliminated in the unpressurized tube by painting silver stripes on the outside of the glass bulb to maintain the bulb at ground potential and by using materials of lower vapor pressure. One pressurized tube and one unpressurized tube were successfully flown on RELAY II.

To maintain repeater redundancy, each repeater is supplied with a separate DC supply voltage of 22.5 volts, regulated to better than 1%. Each TWT requires a power supply to convert the regulated 22.5 volts DC to the proper filament, collector, helix, and anode voltages. The TWT power supply consists of an input low-pass filter, a DC-to-AC inverter operating at 2.5 kc, two transformer-rectifier-filter circuits, a solid-state time-delay circuit and switch, and telemetry circuitry. Since the TWT was found to tolerate power supply input variation of 3% with no more than 0.5 db output variation, no additional regulation was needed in the TWT power supply. The time-delay circuit provides a 3-minute delay between the application of filament and collector voltages and the application of anode and helix voltages. Because the spacecraft mission did not require the TWT power supplies to be turned on during launch or the first spacecraft orbit, the unit was left unpressurized, and high-voltage components (other than the corona-free oil-impregnated high-voltage transformers) were not potted to save weight. The complete power supply weighs 4.4 pounds, and it operates at 75% efficiency, while supplying the following electrode voltages and currents for the TWT:

Filaments	4.25 volts @ 1.6 amp
Collector	950 volts @ 45 ma
Helix	2,030 volts @ 1 ma
Anode	1,500 volts @ 0.1 ma

Beacon

The 4,080-Mc CW beacon signal, used by the ground stations for acquisition and tracking, is generated in essentially the same manner as the transmitter pump signal. A 10-mw local oscillator drives a buffer stage, a power amplifier, a quadrupler, and a cavity-mounted times-nine multiplier (instead of two tripler stages as in the transmitter local oscillator chain). The beacon supplies 0.5 mw of power to the coupler.

Antennas

Both repeater antennas, one to receive and one to transmit, are mounted in a single assembly (Fig. 5). Each antenna provides an omnidirectional pattern about the spin axis from 40° to 125°. The gain is approximately 0 db, referred to a circularly polarized isotropic antenna. The polarization sense of the two antennas is opposite.

SYSTEM PERFORMANCE

As this article is written, the RELAY I satellite has orbited the earth more than 4,300 times. During this time, the RELAY ground stations have implemented an intensive measurement program to evaluate the satellite and overall system performance.

Wideband Performance

Satisfying the system criteria specified in Table I required an accurate identification of signal and noise powers along the communication path. After RELAY I was launched, received carrier power and noise measurements were compared with theoretical predictions to confirm the accuracy of system calculations. During satellite orbit No. 200, the Pleumer-Bodou ground station in France measured received carrier powers that agreed with the calculations to within 2 db. Confirming measurements were repeated at the Goonhilly station in England during satellite orbit No. 285, and at the Nutley station in New Jersey during satellite orbit No. 565. The reasonably close correlation between predicted and measured received carrier powers indicates that ground station and satellite equipment performance is well understood.

Table II presents the parameters involved in a typical performance calculation for communication between the Andover, Maine and Goonhilly, England ground stations. Using estimations of system parameters such as appear in Table II, a satellite noise figure of 14.5 db was measured at Goonhilly during satellite orbit No. 1075. This noise figure agrees closely with the 14.0-db noise figure measured during prelaunch tests.

Table III compares group-delay design objectives with prelaunch and postlaunch measurements. Also included in Table III is the calculated intermodulation noise represented by that group delay for 300-channel FDM-FM telephony.

The measurements indicate agreement among the different parabolic group delays. However, the measured linear and ripple group delays vary from ground station to ground station. This suggests that some components of group delay are generated by the ground station equipment. The measured group delays represent significantly more intermodulation noise than the satellite group delay measured before launch.

These group-delay measurements, along with their associated intermodulation noise contributions, indicate that system performance may be significantly improved by equalizing the group-delay components introduced by the ground stations.

TABLE I—System Objectives

Characteristic	*Large Station*	*Small Station*
Thermal Noise		
Narrowband* (1.3 Mc), pw	7,500	50,000
Wideband** (25 Mc), pw	7,500	50,000
Intermodulation Noise (Psophometrically Weighted)		
Narrowband, pw	7,500	7,500
Wideband, pw	7,500	7,500
Intermodulation Plus Thermal Noise		
Narrowband, pw	15,000	57,500
Wideband, pw	15,000	57,500
Audio Crosstalk, dbm.	−55	−55
Video S/N, weighted, with preemphasis, db	43	not required
Audio S/N, db	50	50

* 12 telephone channels in each direction.
** 300 telephone channels in one direction or TV.

TABLE II—Typical Parameters for Andover-to-Goonhilly Communication*

Parameter	*Ground-to-Spacecraft*	*Spacecraft-to-Ground*
Frequency, Mc	1,725	4,170
Modulation	FM	FM
Transmitter power, w	10,000	10
Diplexer and cable loss, db	0.5	1
Transmitter antenna gain, db	50.2	−1
Space loss (5,000 naut. miles), db	176.5	184.2
Ellipticity loss, db	1	1
Receiver antenna gain, db	−1	58.4
Received signal power, dbm	−58.8	−88.8
Received noise density, dbw/Mc	−101.5	−118.6
Received noise bandwidth, Mc	23	24.3
Receiver noise power, dbm	−87.9	−104.7
Predicted *S/N*, db	29.1	15.9
Threshold, db	12	7.7
Margin (*S/N* minus threshold), db	17.1	8.2

* Communication System Performance Criteria Revision IV 17 May 1962.

TABLE III—Group-Delay and Intermodulation Noise Comparison

Note: All noise powers are calculated from the group delay and do not include pre-emphasis.

	Linear Group Delay nsec/Mc	Linear Group Delay Noise, pw	Parabolic Group Delay nsec/Mc^2	Parabolic Group Delay Noise, pw	Ripple Group Delay nsec (peak-to-peak)	Ripple Group Delay Noise, pw	Total Estimated Intermodulation Noise, pw
Design objective	0.294	140	0.378	6,600	0	0	6,740
Prelaunch Measurement WCS-2 (25°C)	−0.1	16	0.21	2,000	0	0	2,016
Andover, Satellite Orbits No. 200, 247, 495	1.875	5,500	0.208	2,000	10 (5- to 6-Mc period)	800*	8,300
Pleumeur-Bodou	0.67	700	0.21	2,000	14 (12-Mc period)	6,000*	8,700

* D. P. Sullivan, *Space Technology Laboratories Report No. 8614-6030-RV-00*, June 1962.

Group-delay measurements are not the most accurate indicators of noise generated by the system. The noise-to-power ratio measurement specifically measures the thermal noise plus the intermodulation noise in the system's baseband and is a direct measure of the system's quality.

To measure the noise-to-power ratio, every input baseband channel is filled with noise. A comparison of the noise in the corresponding output channel, first with the noise present and then with it absent at the corresponding input channel, yields the system noise-to-power ratio for that channel. Thus, this ratio is a measure of the amount of intermodulation noise appearing in a given channel from all the other channels.

The Relay intermodulation noise objective was 7,500 pw with 300-channel noise loading and an RMS deviation of 675 kc. During satellite orbit No. 1,237, Goonhilly ground-station personnel measured the noise levels of Table IV.

The total noise (with pre-emphasis) of 16,000 pw is only 1,000 pw above the system total noise specification of 15,000 pw. Without pre-emphasis, intermodulation noise (like thermal noise) has a triangular voltage spectrum.

In addition to the wideband measurements already mentioned, the signal-to-noise improvement with video weighting is a measure of the system performance. Video weighting should yield a 12.3-db signal-to-noise improvement for line pictures with a triangular noise spectrum. Weighting improvements much less than 12.3 db indicate either a nontriangular thermal-noise spectrum in the baseband of the receiving ground station (which may result from operation near the demodulator threshold), improper FM threshold improvement behavior of the ground station receiver, or an equipment defect in the satellite or ground station. The video-weighting improvement measured by Nutley ground-station personnel at different times during satellite orbit No. 1,440 was between 10.7 and 13.1 db. This agrees fairly closely with theoretical predictions. It is not known to what extent the correct weighting has been measured at the other ground stations.

The Relay system objective for video-to-audio crosstalk was 50 db or better. Prelaunch measurements on the Relay satellite indicated better than a 60-db ratio. The crosstalk ratio measured by Goonhilly ground-station personnel during satellite orbit No. 740 was 47 db. Since the crosstalk ratio is a function of the group delay, the different crosstalk ratios may indicate different group delays at each measurement site. At the time of measurement, the Goonhilly station group delay had not yet been equalized.

Although Relay was not specifically designed for color television, prelaunch measurements on the satellite repeaters indicated that this capability existed. Color television signals have been successfully transmitted between the Andover ground station and the Relay satellite.

Narrowband Performance

In the narrowband mode, just as in the wideband mode, thermal noise in a channel degrades the signal quality. Table V compares some measured thermal noise levels at the Nutley and Goonhilly ground stations.

Relay II prelaunch measurements indicated a crosstalk level (due to FM-AM-PM conversion) of less than -58 dbm_0. (*dbm_0* is db relative to a point of zero reference level.) Measurements performed by various ground stations between satellite orbits No. 199 and 1,089 indicated crosstalk levels between -28.2 and -54.8 dbm_0. The crosstalk level of -54.8 dbm_0 measured by Andover ground-station personnel almost satisfies the objective of -55 dbm_0. The higher crosstalk levels measured by the other ground stations may include significant ground station crosstalk contributions. Prelaunch measurements of crosstalk on Relay I were not accurate because of a test-equipment defect.

Noise-to-power ratio measurements were performed to evaluate the intermodulation noise for two-way telephony operation. As previously mentioned, the noise-to-power ratio design objective for intermodulation noise was 7,500 pw. During 13 satellite orbits between No. 774 and 1,862, Nutley ground-station personnel measured the noise-to-power ratio at baseband frequencies of 14 kc, 34 kc, and 56 kc. Ten of these thirteen measurements indicated less than 7,500 pw of intermodulation noise. Thus, the noise-to-power ratio objective for the narrowband mode was, on the average, met by the communication path between the Relay satellite and the Nutley ground station.

SUMMARY

Presently (June 1964), the Relay I and II satellites are performing satisfactorily. To the extent that the telemetry system can detect degradation, none of the communications equipment has degraded. No reports of degraded communications performance have been received from the ground stations. However, the Relay I satellite capability has degraded; first, because the solar-cell capacity was reduced by Van Allen radiation and second, because two electronic failures have occurred in the power-supply subsystem. One of the electronic failures rendered one of the three batteries inoperative, thus reducing the useful duty factor. The other failure involves a transistor switch providing power to the No. 1 repeater. This switch occasionally failed to completely disconnect the repeater, thus draining the batteries and rendering the satellite useless temporarily. This difficulty has plagued the satellite since launch. Fortunately, the switch has never remained in the leaky condition.

These troubles have been eliminated on Relay II. To date, the only evidence of a malfunction has been an apparently intermittent beacon. This intermittent cleared after a short time, but the difficulty may recur.

The experience gained with the two Relay satellites points to the effectiveness of the overall program which was characterized by conservative design, careful parts selection, and meticulous and thorough testing.

TABLE IV—Goonhilly Ground Station Thermal Plus Intermodulation Noise Measurement, Satellite Orbit No. 1237

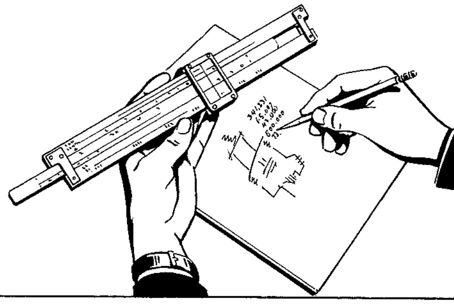
	Thermal Noise, pw	Intermodulation Noise, pw	Total Noise, pw
With pre-emphasis	—	—	16,000
Without pre-emphasis	16,000	24,000	40,000

TABLE V—Narrowband Thermal Noise Performance

Ground Station	Pre-emphasis	Channel No.	Thermal Noise Objective, pw	Measured Thermal Noise, pw
Nutley (Orbit 284)	yes	1	50,000	50,200
	yes	12	50,000	31,600
Goonhilly (Orbit 261)	no	2	7,500	2,200
	no	5	7,500	2,700
	no	11	7,500	7,000
Goonhilly (Orbit 1,962)	no	1	7,500	80
	no	12	7,500	300

Engineering and Research NOTES

BRIEF TECHNICAL PAPERS OF CURRENT INTEREST



An Improved Detection Technique for Random Access Discrete Address Communications

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Final manuscript received September 10, 1964

In a previous *Note*[1], an optimum system parameter adjustment was determined such that, for a specified number of active subscribers and an allotted system bandwidth, the probability of falsely generating a desired address was minimized. In that analysis, the receiver (synchronized to the desired transmission) was assumed to make its pulse/no-pulse decision on a single impulsive sample of the random width coincidence pulse stream. Thus, no use is made of the width of the received coincidence pulses; consequently, that method of detection, although practical, is amenable to improvement.

Owing to the random occurrence of undesired pulses on each address channel, a fortuitously generated coincidence pulse will generally be narrower than its constituent pulses. A desired coincidence pulse must be at least as wide as its constituent pulses. Improved performance accrues through the use of a receiver which integrates the coincidence pulse stream over precisely that interval during which the desired pulse/no-pulse is expected and accepts, as desired pulses, only those coincidence pulses which remain in progress over the entire interval. With integration detection, all coincidence pulses having a substandard width (which, with probability one, are generated by undesired traffic) are ignored. Thus, with integration detection, a false address results *only* when the fortuitous occurrence of undesired pulses produces a coincidence pulse which is in progress at the beginning of the desired pulse interval *and* remains in progress for (at least) the entire interval. Provided that a false pulse is in progress at the beginning of the desired interval on any single address channel, the probability that it remains in progress for (at least) the entire interval is[2]:

$$r_1 = 1 - \frac{\lambda q^o}{P_o} \tag{1}$$

Where: q_o, the probability that no pulse is in progress on any single address channel, is:

$$q_o = (1 - p_o) = \exp(-\lambda) \tag{2}$$

And, the system load factor, λ, is related to the system parameters by:

$$\lambda = \frac{Mnt}{PT} = \frac{Mn}{BT} = \frac{n}{E} = \alpha n$$

Where: M = number of interfering active subscribers, n = number of pulses per address, T = average period of address transmissions per subscriber, $B = p/t$ = a measure of overall system bandwidth, p = number of address channels, t = width of each rectangular pulse, and $E = 1/\alpha = BT/M$ = a logical definition for bandwidth expansion factor.

With integration detection, the probability that all n elements of the desired address appear (with statistical independence) simultaneously and continuously over the entire interval, and thereby generate a false address, is $\varepsilon = [p_o\, r_1]^n$, which, from Eqs. 1 and 2, can be expressed as:

$$\begin{aligned} \varepsilon &= [1 - (1+\lambda)\exp(-\lambda)]^n \\ &= [1 - (1+\alpha n)\exp(-\alpha n)]^n \end{aligned} \tag{3}$$

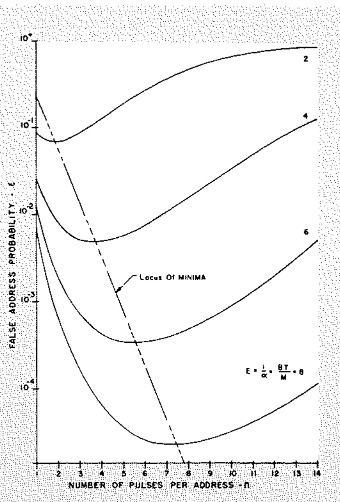

Fig. 1—n vs. ϵ

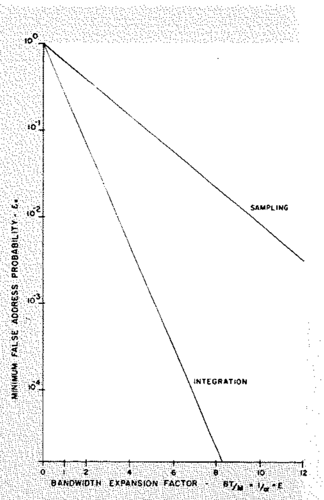

Fig. 2—E vs. ϵ_o

This result is illustrated in Fig. 1 for several values of bandwidth expansion. Of interest is the location of the minima of ε versus n. Upon taking the derivative of ε with respect to n, and equating the result to zero:

$$\frac{d\varepsilon}{dn} = \alpha^2 n^2 \exp(-\alpha n) + [1 - (1+\alpha n)\exp(-\alpha n)]\, ln\, [1 - (1+\alpha n)\exp(-\alpha n)] = 0$$

Or:

$$\frac{d\varepsilon}{dn} = \lambda^2 \exp(-\lambda) + [1 - (1+\alpha n)\exp(-\alpha n)]\, ln\, [1 - (1+\alpha n)\exp(-\alpha n)] = 0 \tag{4}$$

That value of λ which satisfies the constraint imposed by Eq. 4 minimizes ε for a given M and B. By numerical means, the optimum λ has been determined as $\lambda_o = 0.932642\ldots$ (Sampling detection optimized at $\lambda_o = ln\ 2 = 0.693147\ldots$.) By substituting the optimum value of λ into Eq. 3, the minimum false address probability which is obtained through optimizing a RADA system with integration detection is:

$$\varepsilon_o = (0.23948\ldots)^{n_o} \tag{5}$$

Where: the optimum value of n can be expressed in terms of the bandwidth expansion factor by $n_o = \lambda_o/\alpha = \lambda_o E$. Upon substituting the optimum value of n into Eq. 5, one obtains the minimum false address probability as a function of bandwidth expansion factor given by:

$$\varepsilon_o = (0.26368\ldots)^E. \tag{6}$$

Correspondingly, optimized sampling detection yields $\varepsilon_o = (0.61850\ldots)^E$. These results are compared in Fig. 2. When comparing optimized systems, one observes that, for the same false address probability, sampling detection requires about 2.774 times the bandwidth required for integration detection.

Acknowledgement: The author wishes to thank M. Landis for numerically solving Eq. 4.

1. R. C. Sommer, "On the Optimization of Random Access Discrete Address Communications," (Correspondence), *Proc. IEEE 52*,10; p. 1255, October 1964. *Also in:* (E&R Note), RCA ENGINEER, Vol. 10, No. 2, Aug.-Sept. 1964.
2. C. H. Dawson and H. Sklar, "False Addresses in a Random Access System Employing Discrete Time-Frequency Addressing," *1964 IEEE International Convention Record.*

Measurement of Inductor Q by Ringing-Circuit Technique

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The Q of an inductor is the ratio between the reactance and the resistance at a specified frequency. Measurement of the Q can be made on various types of AC bridges and with Q-meters. However, sometimes it is necessary to obtain the Q of an inductor where a standard type instrument of the correct frequency range is not available. The method of applying the "*ringing circuit*" may then be used. The additional equipment required consists of a square-wave or pulse generator and an oscilloscope. Since this equipment is usually available in a laboratory, the method can be applied readily, and the Q at the desired frequency determined.

The inductor can be hooked up in the ringing-circuit technique and fed from a high-impedance source (Fig. 1). The pulse or square-wave generator is operated at a low frequency, say about 1/40th the circuit resonant frequency. The inductor is then tuned to the desired resonant frequency by means of a low-loss capacitor. However, if the overall Q of the resonant circuit is desired, the capacitor that will be used can be substituted. The waveform of the circuit's self oscillation is observed on the oscilloscope, and is similar to Fig. 2. The decrement of the circuit is determined by measuring the ratio between the amplitudes of the pulses. For reasonably high-Q inductors, the amplitude between a high-amplitude pulse, say the second pulse, and the n^{th} *subsequent* pulse is measured. The ratio of the amplitudes is then recorded and the Q calculated:

$$\frac{I_1}{I_n} = \varepsilon^{\frac{n\pi}{Q}}$$

For example, suppose the ratio of the amplitudes between pulses *1* and *11* is 2. Then, $Q = 10\pi/0.693 = 45.3$. If the ratio of the amplitudes measured is 5, then $Q = 10\pi/1.61 = 19.5$.

In using the ringing-circuit technique, the generator and the oscilloscope *must not* load the circuit. However, if this load is known, then the unloaded Q of the circuit can be calculated.

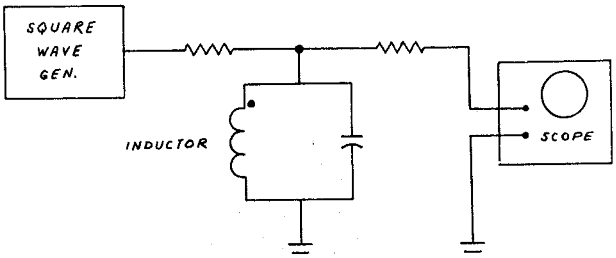


Fig. 1—Inductor in "ringing circuit."

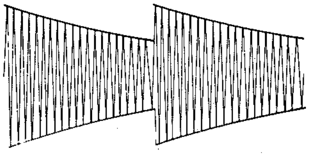


Fig. 2—Self-oscillation waveform.

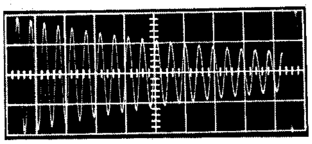
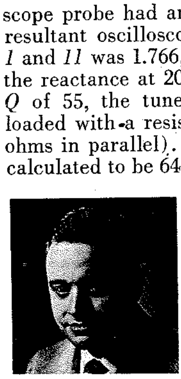


Fig. 3—Lab test trace.

In a laboratory test setup, a 70.5-mh coil was resonated with a decade capacitor to ring at a frequency of 20,000 cps. The generator was connected to the coil through a 5-megohm resistor. The scope probe had an input impedance of 10 megohms. Fig. 3 is the resultant oscilloscope trace. In this test, the ratio between pulses *1* and *11* was 1.766, and the Q was 55. From the value of inductance, the reactance at 20,000 cps was calculated to be 8,850 ohms. For a Q of 55, the tuned impedance would be 487,000 obtained when loaded with a resistance of 3.3 megohms (5 megohms and 10 megohms in parallel). The Q of the coil without the external load was calculated to be 64.5.



M. Cardona

A New Effect in Superconductors

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Superconductors form the basis for many devices such as high-field magnets, computer memories, and a low-noise microwave amplifier.

All superconductors revert to the normal high-resistance state when they carry too high a current or are exposed to too high a magnetic field. *Soft* superconductors such as lead and tin revert to the normal state even at low currents and at magnetic fields of a few hundred gauss, a fortunate situation for computer memories because the memory elements can be switched with low drive currents. *Hard* superconductors such as niobium stannide, on the other hand, can withstand high currents and fields of as much as 180,000 gauss—which makes high-field magnets possible.

By measurements of the absorption of microwaves, we have observed a new effect in soft superconductors[1] and confirmed it in hard superconductors[2]. The effect is the persistence of superconductivity in a thin surface layer of the material at magnetic fields nearly twice those at which the interior of the material reverts to normal. The surface superconductivity depends strongly on the field direction; it is pronounced when the magnetic field is parallel to the surface and disappears when the field is perpendicular to the surface. The effect had been predicted theoretically by others a few weeks earlier for hard superconductors.

The significance of the new results are: 1) the knowledge of surface superconductivity may help to clarify the results of high-field measurements of hard superconductors; as a result, the development of superconducting magnets should be aided; 2) surface superconductivity in soft superconductors could reduce the speed of computer elements; low speeds could thus be obtained inadvertently, but with our new information this can be avoided by placing a thin film of normal metal on the surfaces of the superconductor; 3) new devices based on surface superconductivity are a possibility; an elementary example is a device that has three states—normal, superconducting, and surface-superconducting.

1. B. Rosenblum and M. Cardona, *Phys. Letters 9*, 220 (1964).
2. M. Cardona and B. Rosenblum, *Phys. Letters 8*, 308 (1964).

Transient-Free, Automatic Switch-Over Standby Power Supply

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A military communication system to be designed by RCA imposed difficult requirements on the standby battery power supply subsystem. The standby power supply had to be switched in automatically upon failure of the primary power — and performed without the usual transients or fall-off in power experienced when back-up power is switched in manually. It was critically important to assure continuous service in the overall communication system. A power supply meeting these demands was not available in the industry; conventional types were relatively crude, slow to switch-over and bulky. As a result, the project engineers in CSD produced an advanced design with the characteristics described below. Implementation of the design was subcontracted to Deltron, Inc. of Phila., Pa.

The standby power supply consists of lead-acid batteries in nine separate and isolated outputs packaged in one rack; the supply delivers a total power of 4400 watts and covers voltages from a nominal 6 to 50 volts and currents of 5 to 120 amperes. Power is supplied automatically for a maximum of one hour within a transient period of approximately 200 microseconds. After service, the standby batteries go "on charge" immediately and are fully charged within a 20-hour period, ready for reuse. However, the standby supply will again switch on automatically even though not fully charged, in case of a second primary power failure. After recharge, the batteries are held in their fully-charged condition electronically without harming the cells over a long term. Mechanical characteristics of the power supply are based on the ability to withstand two environments: 1) vibration in the range of 2 to 200 cycles, with a maximum acceleration of 5 G's and 2) half-sine-wave shock impulses of 10 G amplitude with 11 microseconds duration.

Design Approach: The batteries are employed in parallel with the output of an electronically regulated power supply and isolated by means of a switching element. To provide high reliability as well as transient free switchover, a bank of transistors was chosen for the purpose. This proved to be extremely practical because of the existence of very-low-saturation-voltage transistors developed for high-current switching; the transistors make use of an aluminum doped emitter to provide the desired characteristics. A family of transistors possessing these characteristics include the 2N2152 through 2N2159. The particular type used in this design is the 2N2156 with a maximum saturation voltage specification of 100 millivolts at a 5-amp collector current. To successfully parallel these transistors before application in the circuit, and not being able to use emitter equalizer resistors, the units had to be matched for beta within a tolerance of ±20%. Typical measurements based on collector currents of 10 amps/transistor yielded voltage drops in the vicinity of 80 millivolts; such a voltage-drop level vies favorably with good contactors having contact resistances in the range of 5 to 10 milliohms.

By paralleling the transistors at a current of 10 amps/transistor, a current-carrying capability of 120 amps was achieved with a voltage drop less than 100 millivolts. These excellent characteristics provide a measure of reliability not possbile with mechanical contactors.

Design Description: The total system (Fig. 1) is broken down into the following parts: main power supply; over- and under-voltage circuitry; battery switcher; and end-of-charge controller.

The main switching element consists of a bank of high-current switching transistors Q101, arranged with base resistors to equalize for the differing base diode characteristics. The parallel transistors had been previously selected for beta, matched to within ±20%.

To minimize standby power required from the battery when the power supply is furnishing the load current, the base drive for the switching bank is fed through a series control. Q105 is a Darlington driver for transistor Q102, both of which are arranged in a base-driving circuit which has a toggle action. The base of Q105 is fed from the collector of Q103, which in conjunction with Q104 forms a differential comparator amplifier.

All transistors, with the exception of the high-current switchnig units, are silicon with inherently low leakage current characteristics. This is extremely important in connection with the base circuit of the switching transistors. Under normal conditions the battery voltage is approximately the same as the output of the series regulator of the main supply so that there is no net voltage across the series transistors of the battery-switcher.

After a primary power failure, however, and following one hour of battery operation when the AC power is restored, the power-supply output voltage will be higher than the battery voltage; under this condition, Q101 will be reverse-biased as a transistor and, due to leakage, may turn itself on so that the power supply output is clamped to the battery voltage, resulting in a rapid full charge.

Simultaneously, the charge controller is feeding an additional current into the battery so that its voltage will eventually become larger than the output voltage of the power supply, biasing Q101 in the normal direction. Since this is a gradual action it will pass through zero, permitting transistors Q101 to assume an equilibrium temperature equal to that of the ambient. The leakage will, therefore, be small, even though the transistors are germanium types, so that the series base control is possible. Shunt elements are avoided since they would present conduction paths for the transistor when it is reverse-biased.

The differential amplifier turns on the switching element when the output voltage drops slightly below the setting of the power supply output which corresponds to the full-charge voltage of the batteries. Thus, the battery switcher acts as a second power supply in parallel with the main series regulated supply, but adjusted to supply a voltage slightly lower than the main power supply voltage.

After a power failure, as output voltage drops a small amount, the battery switcher turns on automatically with a toggle action; such action results from the high-gain current-amplifier which feeds the base of the series switching elements through a resistor which absorbs most of the voltage. The circuitry, with the exception of the differential amplifier, employs a switching mode rather than a dissipative Class A type of operation.

When Q101 is passing battery current to the load, its base transistor Q102 must be turned on, thereby bringing the collector of Q102 to the negative potential. Thus, a voltage from the positive input to that collector is developed nearly equal to the input battery voltage. The collector is therefore fed through a steering diode CR102 and a limiting resistor R103 to a Zener diode CR104 to provide a 6-volt signal, indicating battery operation.

When power is restored, provided that the under-voltage point had not been reached, relay K3 will immediately become energized and the regulated power supply will produce an output voltage at its terminals. As this voltage rises, it will bias the differential amplifier to cut off the drive to the battery switcher, thereby separating the battery circuit from the output terminals. The charger-controller (Fig. 3) resumes its function of charging the battery.

The battery is fed through a quasi-constant current source contained in the regulated power supply section of the system. This quasi-constant current circuit has two output current ratings selected by means of relay K2; one is a fast rate, capable of recharging the discharged battery within a 20-hour period, and the second is a trickle-charge rate capable of keeping the batteries fully charged for an indefinite period of time.

A precise analog comparator determines the state of charge of the battery; when full charge is reached, the comparator automatically transfers the battery to constant-current-trickle charge. As shown in Fig. 3 a reference Zener CR202 is used to compare transistor Q201 against a fraction of the output. As the output rises and reaches a preset voltage level, transistor Q201 will turn on; base current is then applied to transistor Q202, energizing relay K2. The moment that K2 is energized it will be seen that the contact in the sampling network causes the relay to hold in a latched fashion. A second contact on relay K2 shown on Fig. 1 is used to select the trickle-charge rate.

As polarization gradually subsides, the battery voltage returns asymptotically to its full-charge value. When temperature conditions or other variables reduce the trickle charge rate so that the battery is not in a full-charge condition, the voltage will drop below the full-charge voltage. This continues until transistor Q201 is cut-off, which simultaneously cuts off base current to Q202 and cuts off current to relay coil K2. This allows relay K2 to drop back in the de-energized position, again putting the battery under its 20-hour recharge rate.

The contact K2 in the sampling network is used, therefore, to produce a precision voltage operating relay of controllable differential; thus, several important benefits are achieved: 1) there is a high degree of assurance that the battery is fully charged, 2) the controller constantly monitors the battery terminal voltage with a considerable degree of precision; and 3) by properly setting the cutoff point, battery charging is eliminated.

Oscilloscopic verification of the transitions indicates that they are completely transient free, inasmuch as the power supply output has a storage capacitor across it which makes the output voltage drop relatively slowly after the loss of power. This permits the battery switcher (with a turn-on time of a few microseconds) to turn on without producing any visible transient. Similar transient free conditions exist during power-supply take over on the reapplication of primary power. The main power supply should be highly regulated and stable with environmental conditions, since an output voltage drop below the switcher turn-on point would cause the battery to supply the load through the battery switcher.

In this equipment, the main power supplies were built to a net regulation tolerance better than 0.05%. A similar specification limit was imposed on the battery switcher and the charger controller.

The transient free power supply described above may be used in other ground communications systems both military and industrial wherever automatic and fast switch-over between primary power and battery power is required.

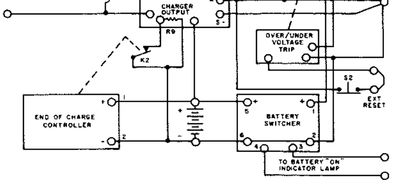


Fig. 1 — Overall power supply block diagram.

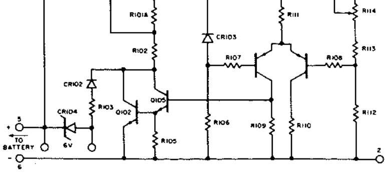


Fig. 2 — Battery switcher.

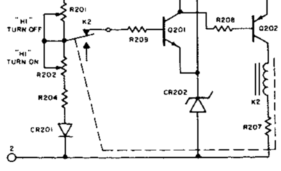
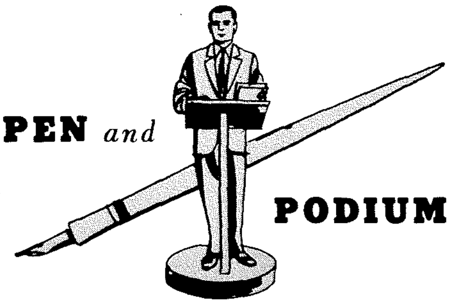


Fig. 3 — "End-of-charge" control circuit.

PEN *and* PODIUM



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Superconducting Energy Gap and Net Electron Drift Velocity as Functions of Temperature and Cooper-Pair Drift Velocity—R. H. Parmenter, L. J. Berton (Labs, Pr.) *RCA Review,* Vol. 25, No. 3, Sept. 1964

Superconducting Energy Gap of Lead Above H_c—Y. Goldstein (Labs, Pr.) 9th International Conf. on Low Temperature Physics, Columbus, Ohio, Aug. 31, 1964

Surface Superconductivity and Supercooling in Hg-Cd Alloys—M. Cardona, B. Rosenblum (Labs, Pr.) *Physics Letters,* Vol. 11, No. 2, July 15, 1964

Switching Times of the Current-Induced Superconducting-to-Normal Transition in Filaments of Tin and Indium—J. I. Gittleman, S. Bozowski (Labs, Pr.) *Physical Review,* Vol. 135, No. 2A, July 20, 1964

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Switching Times of the Current-Induced Superconducting-to-Normal Transition in Filaments of Tin and Indium—J. I. Gittleman, S. Bozowski (Labs, Pr.) *Physical Review,* Vol. 135, No. 2A, July 20, 1964

TELEVISION BROADCASTING

Color Television, Progress in—V. K. Zworykin (Labs, Pr.) MIFED Comite International de Television Congress Excot 1964; Milan, Oct. 17, 1964

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TELEVISION COMPONENTS

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Sync Generator, New Transistorized—R. J. Smith (BCD, Cam.) IEEE Broadcast Symp., Washington, D.C., Sept. 24, 1964

TUBE DESIGN; APPLICATION

Color Picture Tube, Development of the RCA 25-Inch 90-Degree Rectangular—A.M. Morrell, A. E. Hardy (ECD, Lanc.) National Electronics Conf., Chicago, Ill., Oct. 19-21, 1964

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Microwave Phototubes Using Transmission Electron Multipliers—D. J. Blattner, H. C. Johnson, G. A. Morton, J. E. Ruedy, F. Sterzer (ECD, Pr.) International Conf. on Microwave Tubes, Paris, France, Sept. 14-16, 1964

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TUBE MATERIALS; THEORY

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VACUUM TECHNOLOGY

Vacuum Technology—J. T. Mark (ECD, Lanc.) Graduate Student Group, Penna. State Univ., Sept. 22, 1964

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Buizert, H. antennas
Gore, V. instrumentation; lab equipment
Kaufman, Dr. communications, equipment components
Osborne, Dr. F. J. F. instrumentation; lab equipment
Roberts, G. communications, equipment components
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Shkarofsky, Dr. I. P. instrumentation; lab equipment
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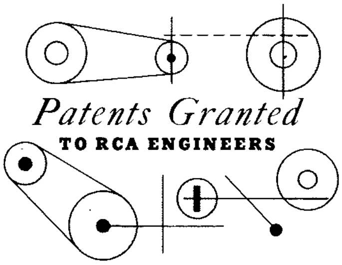

Patents Granted
TO RCA ENGINEERS

Based on Summaries Received over a period of about two months

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3,140,484—Range Sweep and Positioning Circuit, July 7, 1964; A. I. Mintzer (assigned to U.S. Government)

3,143,727—Magnetic Memory and Switching Circuit, August 4, 1964; W. L. Morgan, II (assigned to U.S. Government)

3,150,370—Radar Sequencer, September 22, 1964; A. J. Lisicky (assigned to U.S. Government)

3,154,720—Solid State Display Device, October 27, 1964; M. Cooperman

3,156,398—Tape Handling Apparatus, November 10, 1964; C. Lauxen and J. B. Long, Jr.

3,156,901—Shift Register Systems, November 10, 1964; A. J. Kline, Jr.

3,157,795—Transistor Inverter Utilizing an Impedance Directly Connected Between Base and Collector to Prevent Saturation, November 17, 1964; A. I. Pressman

3,157,797—Switching Circuit, November 17, 1964; C. R. Eshelman

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3,147,336—Slot Radiator in Wall of Coaxial Feed Having Mode Coupling Means at Slot, August 14, 1964; N. Nikolayuk

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3,147,435—Strip Line Phase Comparator, September 1, 1964; D. J. Blattner (assigned to U.S. Government)

3,150,325 — Wide Band Traveling Wave Parametric Amplifier, September 22, 1964; D. J. Blattner (assigned to U.S. Government)

3,153,172 — Automatic Brightness Control Using a Light Conducting Rod and Photocell, October 13, 1964; K. S. Ling

3,153,190—Method of Testing and Controlling the Gettering of Electron Tubes During Manufacture, October 13, 1964; R. L. Spalding

3,154,714—Anode Structure, October 27, 1964; L. R. Gormay

3,154,975—Apparatus for Pointing and Severing Wires or Pins, November 3, 1964; J. A. Chase

3,155,767—Connecting Arrangement in Electronic Modular Structures, November 3, 1964; H. F. Schellack

3,154,976 — Wire Shaping Apparatus, November 3, 1964; J. A. Chase

3,156,029—Electron Gun and Fabrication Thereof, November 10, 1964; J. O. Simon

3,156,030 — Cathode Insertion Apparatus, November 10, 1964; H. E. Natalis

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3,152,900—Art of Making Electron-Sensitive Mosaic Screens, October 13, 1964; P. E. Kaus

3,153,778 — Magnetic Core Binary Devices, October 20, 1964; A. W. Lo

3,154,679—Multiplying Devices, October 27, 1964; A. W. Vance

3,154,840—Method of Making a Magnetic Memory, November 3, 1964; R. Shahbender

3,155,886 — Solid State Superconductor Triode, November 3, 1964; J. I. Pankove

3,156,816—Electrical Circuits, November 10, 1964; W. F. Kosonocky and J. J. Amodei

3,156,893 — Self-Referenced Digital PM Receiving System, November 10, 1964; A. Harel

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3,153,228—Converting Systems, October 13, 1964; J. T. Winkler

HOME INSTRUMENTS DIVISION

3,155,779 — Stereophonic Phonograph System, November 3, 1964; J. A. Tourtellot

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DATES and DEADLINES
Professional Meetings and Calls for Papers

Meetings

Jan. 5-8, 1965: Solid-State Physics; Inst. of Physics and Physical Soc.; U. of Bristol. *Prog. Info.:* D. A. Greenwood, H. H. Wills Physics Lab., Royal Fort, Bristol 8: Administration Assistant, IPPS, 47 Belgrave Sq., London, S.W. 1, England.

Jan. 12-14, 1965: 11th Ann. Symp. on Reliability and Quality Control, IEEE-ASQC-IES; Fountainebleu Hotel, Miami Beach, Fla. *Prog. Info.:* H. Reese, Burroughs Corp., P.O. Box 305, Paoli, Pa.

Jan. 28-30, 1965: 48th Ann. Mtg., The Math. Assoc. of America; Denver-Hilton Hotel, Denver, Colorado. *Prog. Info.:* H. M. Gehman, Univ. of Buffalo, Buffalo, New York.

Feb. 3-5, 1965: 6th Winter Convention on Military Electronics, G-MIL, L.A. Sect. IEEE; International Hotel, Los Angeles, Calif. *Prog. Info.:* Dr. R. Ashby, Autonetics Div., North American Aviation, 3370 Miraloma Ave., Anaheim, Calif.

Feb. 17-19, 1965: Intl. Solid State Circuits Conf., IEEE, G-CT, Univ. of Pa.; Univ. of Pa. and Sheraton Hotel, Phila., Pa. *Prog. Info.:* G. B. Herzog, RCA Labs., Princeton, N.J.

March 10-12, 1965: Particle Accelerator Conf., G-NS, et. al., IEEE; Shoreham Hotel, Wash., D.C. *Prog. Info.:* R. S. Livingston, Oak Ridge Natl. Lab., P.O. Box X, Oak Ridge, Tenn.

March 22-25, 1965: IEEE Intl. Convention, IEEE, All Groups TOC Committees; Coliseum and N.Y.-Hilton, N.Y., N.Y. *Prog. Info.:* Dr. E. L. Harder, IEEE Hdqs., Box A, Lenox Hill Station, N.Y., N.Y.

Calls for Papers

April 5-7, 1965: Second Space Congress, Canaveral Council of Tech. Societies; Cocoa Beach, Fla. **Deadline:** Abstracts, 100 wds; summaries, 500-1,000 wds, **1/15/65.** *TO:* Dr. L. E. Mertens, Tech. Program Chairman, RCA-Missile Test Project, M.U. 741, Bldg. 423, Patrick AFB, Fla.

Apr. 13-15, 1965: Natl. Telemetering Conf., IEEE, AIAA-ISA; Shamrock Hilton, Houston, Texas. **Deadline:** Abstracts, approx. **1/15/65.** *TO:* R. W. Towle, Philco Corp., Western Dev. Labs., 3825 Fabian Way, Palo Alto, Calif.

Apr. 14-15, 1965: 1965 Electronics and Instrumentation Conf. and Exhibit, IEEE and ISA, Cinc. Sect.; Cincinnati Gardens, Cincinnati, Ohio. **Deadline:** Abstracts, approx. **11565.** *TO:* J. R. Ebbeler, Avco Corp. 2630 Glendale-Milford Rd., Cincinnati, Ohio.

Apr. 21-23, 1965: Southwestern IEEE Conf. and Elec. Show (SWIEEECO), Region 5; Dallas Memorial Auditorium, Dallas, Texas. **Deadline:** Abstracts, approx. **1/15/65.** *TO:* E. F. Sutherland, Genl. Radio Co., 2501-A Mockingbird Lane, Dallas, Texas.

Apr. 21-23, 1965: 1965 Intl. Nonlinear Magnetics Conf., (INTERMAG), IEEE; Sheraton Park Hotel, Wash., D.C. **Deadline:** Abstracts, approx. **1/15/65.** *TO:* E. W. Pugh, IBM Components Div., Poughkeepsie, N.Y.

May 4-6, 1965: 5th Ann. Packaging Industry Conf., IEEE; Milwaukee Inn, Milwaukee, Wisc. *FOR DEADLINE INFO.:* IEEE, Box A, Lenox Hill Station, New York, N.Y.

May 5-7, 1965: Electronic Components Conf. (ECC), IEEE-EIA; Marriott Motor Hotel, Wash., D.C. **Deadline:** Abstracts, approx. **1/1/65.** *TO:* B. Schwartz, IBM Components Div., Poughkeepsie, N.Y.

May 6-8, 1965: 6th Natl. Symp. on Human Factors in Electronics, IEEE, G-HFE; Sheraton Hotel, Boston, Mass. *FOR DEADLINE INFO.:* IEEE Hdqs., Box A, Lenox Hill Station, N.Y., N.Y.

May 10-12, 1965: Natl. Aerospace Electronics Conf. (NAECON), IEEE, G-ANE, AIAA, Dayton Section, Dayton, Ohio. **Deadline:** Abstracts, approx. **1/1/65.** *TO:* IEEE Dayton Office, 1414 E. 3rd St., Dayton 2, Ohio.

May 13-14, 1965: Symp. on Signal Transmission and Processing, IEEE, G-CT; Columbia Univ., New York, N.Y. **Deadline:** Abstracts, **1/15/65.** *TO:* Dr. L. E. Franks, Bell Tel. Labs., North Andover, Mass.

May 19-21, 1965: Power Industry Computer App. Conf. (PICA), IEEE, G-P and Florida Westcoast Section; Jack Tar Hotel, Clearwater, Fla. *FOR DEADLINE INFO.:* G. W. Staff, American Elec. Power Serv. Corp., 2 Broadway, N.Y.

June 7-9, 1965: 1st Ann. IEEE Communication Convention (GLOBECOM VII), IEEE, G-ComTech, Denver Boulder Section; Univ. of Colo. and NBS Labs., Boulder, Colo. **Deadline:** Abstracts, **2/15/65.** *TO:* Wm. F. Ulaut, NBS, Boulder, Colo.

June 21-25, 1965: San Diego Symp. for Biomedical Eng., IEEE, US Naval Hosp.; San Diego, Calif. *FOR DEADLINE INFO.:* D. L. Franklin, Scripps Clinic and Res. Found., La Jolla, Calif.

June 28-30: 7th Natl. Symp. on Electromagnetic Compatibility, IEEE, G-EMC; Waldorf-Astoria, New York, N.Y. **Deadline:** Papers, approx. **2/15/65.** *TO:* D. Fidelman, Electromag. Meas. Co., Farmingdale, N.Y.

Aug. 23-27, 1965: 6th Intl. Conf. on Medical Elec. and Biological Eng., (IFMEBE) IEEE; Tokyo, Japan. **Deadline:** Abstracts, approx. **4/30/65.** *TO:* Dr. L. E. Flory, RCA Labs., Princeton, N.J.

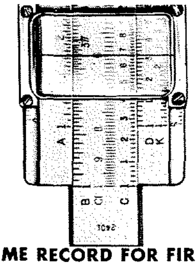
Aug. 24-27, 1965: Wescon, IEEE, WEMA; Cow Palace, San Francisco, Calif. **Deadline:** Abstracts, approx. **4/15/65.** *TO:* IEEE L.A. Office, 3600 Wilshire Blvd., Los Angeles, Calif.

Aug. 30-Sept. 1, 1965: Antennas and Propagation Intl. Symp., IEEE, G-AP; Sheraton Park Hotel, Wash., D.C. **Deadline:** Abstracts, approx. **3/2/65.** *TO:* Dr. R. J. Adams, Serach Radar Branch, Naval Res. Lab., Wash., D.C.

Sept. 22-24, 1965: Intl. Convention on Military Electronics (MIL-E-CON 9), IEEE, G-MIL; Wash. Hilton Hotel, Wash., D.C. *FOR DEADLINE INFO.:* IEEE Hdqs., Box A, Lenox Hill Station, New York, N.Y.

Be sure DEADLINES are met — consult your Technical Publications Administrator for lead time needed to obtain required RCA approvals.

Engineering NEWS *and* HIGHLIGHTS



RCA EARNINGS UP 26%, SET ALL-TIME RECORD FOR FIRST NINE MONTHS OF 1964

RCA sales and earnings in the first nine months of 1964 achieved an all-time high for any comparable period in the company's history. Operating profits after taxes for the first nine months climbed by 26% to a record of $55,800,000, as compared with $44,200,000 in the same 1963 period. After-tax profits for the third quarter increased 23% to a new high of $18,200,000, compared with $14,800,000 for the 1963 third quarter. The third-quarter record marked the fourteenth consecutive quarter in which profits were higher than in the same period of the preceding year.

Sales for the nine-month period rose to a new record of $1,330,500,000 compared to $1,314,000,000 for the 1963 period, an increase of 1 percent. For the third quarter, sales totalled $431,400,000 as against $436,700,000 for the third quarter of the preceding year, a decline of 1%. The decline was caused by the general falling off in defense business.

Operating earnings per share of common stock for the nine months were a record $1.02 compared with 80 cents for the same period in 1963. For the third quarter of 1964, operating earnings per share were 33 cents, another all-time record for any comparable period. Per-share operating earnings for the same period last year were 27 cents.

The nine-month period record was attributed to these principal trends:

1) The crossover into profitability of the company's electronic data processing business during the third quarter of 1964, ahead of the previously established schedule. RCA is the third major company to report a level of profitability in the computer industry, with a volume of equipment sales and rentals expected to exceed the $100 million mark by the end of 1964.

2) Record sales and profits by RCA home instruments for the nine-month period, exceeding the previous all-time levels for 1963. Despite increased competition throughout the industry, RCA color sales continued to maintain a strong leadership position, with a 37% rise in factory unit sales, and a 50% increase in unit sales from distributors to dealers over the same period last year.

3) Record sales and profits for the National Broadcasting Company, with profits for the nine months up by about 20% over the 1963 period. For NBC, the third quarter was the twelfth consecutive period of record-breaking profits. NBC's unparalleled leadership in news programming combined with high reception to its fall entertainment schedule to generate an exceptionally strong advertiser response for the new broadcasting season.

4) A nine-month sales record for the company's electronic components and devices activity, paced by the continually mounting demand for color picture tubes, with profits for the period the highest in 14 years.

5) Record nine-month sales and earnings for RCA Communications, Inc., with sales up 15 percent over the previous nine months of 1963.

RCA's growing sales and profit strength is evident in the fact that sales from the commercial sectors of our business have all but completely offset the continued decline in government sales, which were down by approximately $70 million during the first nine months of 1964.

RCA's continuing cost control program is cited as a major factor in the rising earnings picture. Since 1961, RCA has focussed on a program of rigid production and marketing efficiencies for all operating divisions and services. This encompasses the most advanced methods of planning and control, including the utilization of computers. This program, which is under constant review, is contributing importantly to the maintenance of profit objectives.

RCA COLOR TELEVISION DEMONSTRATED IN EUROPE TO SHOW VALUE OF NTSC SYSTEM

RCA color television using American (NTSC) standards converted to 625 lines, 50 fields has been demonstrated to European Broadcast Union (EBU) officials considering the adoption of standards for European color television.

The demonstration unit was prepared under the supervision of the TV System Group of the Broadcast and Communications Products Division, Camden. It consists of an air conditioned, 35-foot semi-trailer containing RCA color TV equipment capable of providing live or tape-recorded pictures as well as pictures from films and slides. The pictures are displayed on color receivers from the RCA Victor Home Instruments Division.

BCP officials reported that the excellent color pictures produced in the demonstration unit, particularly those that were recorded and played back, are expected to have a significant salutary effect on the consideration given by EBU to NTSC-type standards for European television.

(*Editor's Note:* For background information on color TV in Europe, see the concluding section of J. W. Wentworth's article "Color Television—The First Ten Years," in RCA ENGINEER *10*-2, Aug.-Sept. 1964.)

EXTRA SPEED REQUIRED FOR PATENT DISCLOSURES MADE UNDER DOD CONTRACTS

The Department of Defense recently accelerated the time schedule under which contractors must report inventions made under DOD government contracts. Accordingly, RCA is now required to supply disclosures of inventions to the Government within four months of the date of conception, together with a decision as to whether RCA elects to file a patent application on each particular invention.

Since processing of inventions requires considerable time, it is essential that patent disclosures on contract inventions be submitted to Domestic Patents, RCA, Princeton, N. J., immediately after conception. Failure to do so could not only jeopardize RCA's relations with government agencies, but could adversely affect government contract payments.

Within DEP, Division AR&D Coordinators have been designated to expedite such reporting. However, the responsibility rests primarily with the individual inventors to assist RCA to meet its contractual responsibilities.

B. P. MILLER RECEIVES NASA "PUBLIC SERVICE AWARD" FOR WORK ON RANGER

On October 9, 1964, Bernie Miller, RANGER Project Manager at the Astro-Electronics Division, Princeton, was presented NASA's Public Service Award by James E. Webb, NASA Director. The award reads as follows: *"For outstanding leadership as RCA* RANGER *Project Manager, and for his major contribution to the world's first high resolution photography of the moon on July 31, 1964. For management of the RCA team which was responsible for the design and development of the television camera system on board the* RANGER *VII spacecraft."*

This was the only award made by NASA to an industrial contractor this year.—*J. C. Phillips.*

W. W. THOMAS GETS AOA AWARD

William W. Thomas of DEP Central Engineering, Camden, has been given the *Robert H. Sterns Award* of the American Ordnance Association (AOA). He was honored for *"outstanding personal leadership and achievement in support of the national program of simplification, efficiency, and standardization of engineering documentation."*

A graduate aeronautical engineer and an RCA employee since 1959, Mr. Thomas has staff responsibility for documentation practices, drawing standards, and configuration management procedure for all DEP divisions. He is Chairman of the AOA Technical Documentation Division. He has also served on several national advisory committees and in the winter of 1962-63 spent 5 months as a full-time consultant to the Logistics Management Institute in its study of defense standardization for the Secretary of Defense.

LIND NAMED FELLOW OF SMPTE

A. H. Lind, Mgr., Studio Equipment Engineering, Broadcast and Communications Products Div., Camden, was named a *Fellow* of the SMPTE on Sept. 29, 1964, at their 96th Semiannual Convention in New York City.—*R. N. Hurst.*

DEP MEASUREMENT PROGRAM EXCEEDS INDUSTRY STANDARDS

The Measurements Engineering Laboratory of DEP Central Engineering, Camden, was one of twelve participants in a recent industry-wide Measurement Agreement Program sponsored by the National Conference of Standards Laboratories (NCSL).

The results of the round-robin comparison were reported at the NCSL's second-biennial meeting last month. They showed that in all cases the DEP measurement capability at the reference standard level to be well within the accuracies assumed for the various parameters. In several cases the DEP Laboratory's results equalled those of the National Bureau of Standards.

In addition to servicing all DEP divisions, the Measurements Engineering Laboratory supplies measurement and calibration services on request to Electronic Data Processing, the RCA Service Co., and to Broadcast and Communications Products.

STAFF ANNOUNCEMENTS

Research and Engineering: Effective October 1, 1964, **R. H. Edmondson** was appointed Staff Engineer, reporting to **Dr. G. H. Brown,** Vice President, Research and Engineering.

Broadcast and Communications Products Division: Effective October 16, 1964, the organization of the Engineering Department, reporting to **W. C. Morrison,** Chief Engineer, was announced as follows: **N. C. Colby,** Manager, Communications Products Engineering; **H. N. Kozanowski,** Manager, TV Advanced Development; **A. H. Lind,** Manager, Studio Equipment Engineering; **A. C. Luther,** Manager, Tape Equipment, Projector and Scientific Instruments Engineering; **H. S. Wilson,** Manager, Microwave Engineering; **J. E. Young,** Manager, Broadcast Transmitting Equipment Engineering; and **J. E. Young,** Acting Manager, Engineering Administration and Services.

ECD Special Electronic Components Division: Effective October 1, 1964, the organization of the newly established Integrated Circuit Department, reporting to **L. R. Day,** Acting Manager, was announced as follows: **D. W. Chace,** Manager, Product Administration; **R. D. Lohman,** Manager, Integrated Circuit Engineering; and **R. A. Wissolik,** Manager, Integrated Circuit Products Manufacturing.

Effective October 1, 1964, the organization of Integrated Circuit Products Manufacturing, reporting to **R. A. Wissolik,** Manager, was announced as follows: **H. L. Eberly,** Manager, Manufacturing; **L. P. Fox,** Manager, Production Engineering; **R. R. Giordano,** Manager, Production and Material Control; and **P. Greenberg,** Manager, Quality and Reliability Assurance.

Effective October 1, 1964, the organization of Integrated Circuit Engineering, reporting to **R. D. Lohman,** Manager, was announced as follows: **I. H. Kalish,** Manager, Design; **B. V. Vonderschmidt,** Manager, Application; and **F. M. Yates,** Administrator, Engineering Projects.

ECD Technical Programs: Effective September 1, 1964, the staff of **E. O. Johnson,** Manager, Engineering, was announced as follows: **R. B. Janes,** Manager, Advanced Development; H. V. Knauf, Manager, Photomask Operation; and D. H. Wamsley, Staff Engineer.

LICENSED ENGINEERS

G. E. Dunn, HI, Indpls., PE-8522, Ind.
J. C. Johnson, DEP-CSD, Tucson, PE-2775E, Pa.
W. D. Kouns, DEP-CSD, Camden, PE-13257, N. J.
A. L. Lea, RCA Ser. Co., Cherry Hill, PE-18415, Ohio
J. E. Matlin, DEP-ASD, Van Nuys, PE-11860, Calif.
M. B. McVernon, RCA Ser. Co., Riverton, PE-37470, N. Y.
G. A. Metz, RCA Ser. Co., Riverton, PE-5824E, Pa.
R. S. Milne, DEP-CSD, Camden, PE-10347E, Pa.
M. Mitnick, RCA Ser. Co., Cherry Hill, PE-1180E, Pa.
W. H. Paul, ECD, Lancaster, PE-10304E, Pa.
C. L. Rintz, ECD, Lancaster, PE-9294E, Pa.
E. A. Roloff, RCA Ser. Co., Riverton, PE-10262, Missouri
W. F. Schacht, RCA Ser. Co., Cherry Hill, PE-12594, N. J.
R. E. Schell, RCA Labs., Pr., PE-13629, N. J.
K. L. Shaw, RCA Ser. Co., Riverton, PE-18606, Texas
G. L. Stitely, DEP-AED, Princeton, PE-1952, Del.; PE-13634, N. J.
H. D. Twitchell, Jr., DEP-ASD, Burl., PE-19758, Mass.
J. W. Tyler, RCA Ser. Co., Riverton, PE-24581, N. Y.
M. Wauters, RCA Ser. Co., Riverton, PE-29959, N. Y.
J. G. Weaver, RCA Ser. Co., Riverton, PE-7158E, Pa.
H. H. Wittenberg, ECD, Lancaster, PE-10188E, Pa.
M. K. Wilder, BCD, Camden, PE-A-13270, N. J.

COST REDUCTION ACTIVITIES CONSOLIDATED BY DEP

DEP Cost Reduction activities, which since 1959 have produced total savings exceeding $174 million, have been merged and reoriented into a single program, with savings goals established for each activity. **Bert Fein,** Manager of DEP Management Engineering, is in charge of the consolidated program. Each of the four DEP divisions has appointed a cost reduction coordinator. Working within each division with the divisional coordinator are men responsible for the various phases of the overall program. These include: engineering standardization, design review, employee suggestions, employee training, value engineering, zero defects, value analysis, and overhead control.

MILLIONTH 1964 RCA TV SET SOLD

The sale of RCA Victor TV receivers in 1964 passed the million mark more than a month ahead of last year's record pace.

. . . PROMOTIONS . . .

to Engineering Leader & Manager

As reported by your Personnel Activity during the past two months. Location and new supervisor appear in parenthesis.

DEP Communications Systems Division

W. R. Trimmer: from Senior Project Member Tech. Staff to *Ldr., Tech. Staff* (D. C. Maxwell, Cambridge).

RCA Service Company

K. F. Wenz: from Ldr., Engrs. to *Mgr., Pulse Radar Eng.* (R. T. Platt, Jr., Instrumentation Support Eng., MTP).

R. E. Hunter: from Installation and Modification Engr. to *Mgr., C&E Eng.* (A. C. Cowan, Jr., Operations and Eng. White Alice).

F. C. Minning: from Engr. to *Ldr., Engrs.* (R. T. Platt, Jr., Instrumentation Support Eng., Missile Test Project).

Electronic Components and Devices

J. A. Schramm: from Eng. Ldr., Prod. Div. to *Eng. Ldr., Manufacturing* (D. Watson, Somerville)

W. Greig, Jr.: from Engr. to *Engr. Ldr. Prod. Dev.* (A. Rose, Somerville)

B. Czorny: from Engr. Prod. Dev. to *Eng. Ldr., Prod. Dev.* (A. Rose, Somerville).

Broadcast and Communications Products Division

M. S. O. Siukola: from Engr. to *Ldr., Design and Dev. Engrs.* (R. L. Rocamora, Antenna Eng., Camden).

NEW SOUTHWEST WING ADDED TO RCA LABS IN PRINCETON

Construction of a new laboratory wing has been completed at the RCA Laboratories, Princeton, N. J. Ground was broken for the new wing on September 30, 1963, by **Dr. Elmer W. Engstrom,** RCA President.

Scheduled for occupancy in late 1964 and early 1965, this new wing provides 39,000 square feet of additional space to house some forty laboratories, as well as providing office and shop space. When opened for business, the three-story, air-conditioned facility will bring the Laboratories' total space to 472,000 square feet.

Dr. James Hillier, Vice President, RCA Laboratories, pointed out that it will help to relieve the substantially increased pressures upon existing laboratory space resulting from the continued growth of the RCA Laboratories staff and the establishment at the David Sarnoff Research Center of advanced development groups associated with RCA's various manufacturing divisions.

He added that the David Sarnoff Research Center, as the central research organization for all of RCA, has more than doubled in size since its establishment in Princeton in 1942, and that the research staff itself has increased approximately 35% in the last five years alone.

The construction of the Southwest Wing is part of a long-range building program that will include in its next phase the southward extension of the new wing with a main entrance facing Fairview Avenue, the principal access to the Laboratories. This phase will also include the expansion of the cafeteria facilities.

In the later phases of this program it is planned that the Southwest Wing will be continued farther southward to join with Building No. 3 and integrate that building into the main Laboratories building complex. When this is completed, it will enable all of the administrative offices of the David Sarnoff Research Center to be located in a single area.

DR. HILLIER NAMED TO EDITORIAL BOARD OF "INDUSTRIAL RESEARCH"

Dr. James Hillier, Vice President, RCA Laboratories, was recently named to the expanded Editorial Advisory Board of *Industrial Research* magazine. Dr. Hillier was one of twenty prominent scientists, engineers, and research administrators appointed. The new board members include two Nobel Prize winners, three university presidents, three members of the President's Science Advisory Committee, and the president of the National Academy of Sciences.

DEGREES GRANTED

J. R. Allen, DEP-CSD	MSEE, University of Pennsylvania
E. de Haas, DEP-AED	Dr., Tech. Sci., Technological Univ. of Eindhoven, Netherlands
J. B. Feller, DEP-CSD	MSEE, University of Pennsylvania
J. Tullai, DEP-AED	MSEE, Villanova University

"VITA"—THE VOLUNTEERS FOR INTERNATIONAL TECHNICAL ASSISTANCE, INC.

VITA is a nonprofit, membership organization with headquarters in Schenectady, New York, incorporated under the laws of the State of New York in 1960. It is made up of engineers and scientists who want to actively use their skills for the betterment of mankind and yet cannot leave their homes and employment for service overseas. Now, on a part-time and volunteer basis through VITA, these concerned individuals can serve as the back-home technical brains of the Peace Corps and other service organizations.

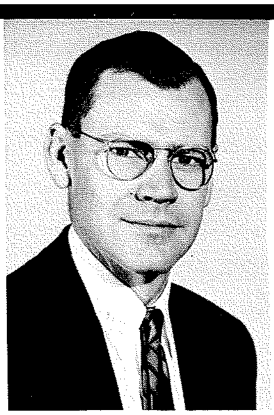
The VITA membership now comprises over 1,000 scientists, engineers, and others who have contributed their professional services. Financial support is drawn from the contributions of individuals and corporations, from foundations, and from proceeds of contract work performed for agencies with financing for that purpose. VITA does not charge recipients in the developing nations for its services. VITA seeks to apply the technical competence of American scientists and engineers to specific projects of direct relevance to the developing countries, in direct partnership with individuals and groups in those countries.

Where VITA participants are concentrated in sufficient numbers within relatively small geographical areas, VITA Chapters have been formed. Nine such chapters are now in operation—the original chapter in Schenectady, New York and environs; Rochester, New York; Metropolitan New York City; Morgantown, West Virginia; New Holland, Pennsylvania; Peoria, Illinois; Santa Barbara, California; and two chapters in Los Angeles.

VITA has formed working relationships with several professional technical societies. The prospects for further development of this aspect of VITA have been recognized by the Engineers Joint Council, which supported a grant from the Engineering Foundation to VITA for this purpose.

More detailed information — including plans for future development of VITA, documentary summaries of VITA "cases," the *VITA Newsletter*, and similar material —may be obtained from: VITA, Inc., 1206 State Street, Schenectady, New York 12304. *Attention:* Mr. Bernard R. Carman.

H. H. Spencer



T. T. Patterson

H. H. SPENCER NAMED TPA AND ED REP FOR EDP, AND A CONSULTING EDITOR TO THE "RCA ENGINEER", REPLACING T. T. PATTERSON

H. H. Spencer, Staff Engineer, RCA Electronic Data Processing, Camden, has been named as Technical Publications Administrator and RCA Engineer Editorial Representative for EDP. In addition, he will serve as a Consulting Editor to the RCA Engineer editorial staff. In all these functions, he replaces **T. T. Patterson,** who recently left EDP Engineering for a new assignment in EDP Cherry Hill. The Editors welcome Mr. Spencer, and offer their thanks and best wishes to Mr. Patterson, who for several years has been especially active in RCA Engineer and RCA technical paper activities.

(For other RCA Engineer Consulting Editors and EDP Editorial Representatives with whom Mr. Spencer will be associated in these activities, see the inside front and inside back covers, respectively.)

H. H. Spencer received his BSEE in 1923 and his MSEE in 1924 both from MIT. He joined RCA in 1946 at the dissolution of the National Defense Research Committee where he had served during World War II as Chief of the Guided Missiles Division. His initial assignment at RCA was government contracting in the field of research and development with particular emphasis on radar-television aids to air navigation and traffic control and on the application of the same techniques to fire control. Over the next 15 years he had varied assignments in the marketing area, an early one of which was the support of the idea that RCA investigate the application of digital electronic computers to business operations. During this period he served as Mgr., New Products Administrator, Mgr. BIZMAC Market Planning, Mgr., EDP R&D Contracting, and Mgr., Computer Equipment, Data Communications and Custom Projects Dept. In 1962, he returned to Engineering as a Staff Engineer. Among his current duties, he coordinates patent activities in EDP and in other divisions where work closely related to EDP's interest is going on, and coordinates EDP technical aid to foreign licensees. He encourages engineers taking graduate studies in the preparation of their theses and assists them in obtaining approval. He is a Registered Professional Engineer in Massachusetts, and a Member of the ASME, IEEE, and Franklin Institute.

The "RCA Engineer" Reprint Service

Since the inception of the RCA Engineer in 1955, articles contained in the journal have been reprinted in increasing numbers. Today, a majority of the papers published in the RCA Engineer are reprinted on order for various RCA activities. While the RCA Engineer itself as a matter of policy cannot be distributed outside RCA, reprints can and are regularly utilized for distribution to government and industry customers, at technical meetings, as background information for seminars, etc.

Reprints vary from single articles up to reprint booklets consisting of several related articles housed in covers. The Editorial Staff of the RCA Engineer provides the service of ordering such reprints, designing special covers as needed, etc. Cost of these reprints is nominal, since master negatives of all previously published articles are retained for that purpose.

Quantities of each reprint are ordered by the various activities providing requisitions for them, and such bulk quantities are the property of those groups; however, token quantities of recent reprints are available from the editorial office of the RCA Engineer. Single copies can usually be furnished without charge upon request through the limited stocks available. The RCA Engineer prepares and distributes to its Editorial Representatives catalog lists of available reprints, as well as price lists and ordering information for quantity orders. For such information contact the Editorial Representative in your activity—or contact the RCA Engineer editorial office, Building 2-8, Camden. Details on reprint policies, information on special booklet-type reprint collections, and suggestions for their utilization can be so obtained.

RCA Engineer articles receive RCA clearance through the RCA review and approval cycle just like any RCA technical paper for "outside" publication. RCA Engineer articles are copyrighted and are the property of RCA, and the copyright protects the reprints also so that the reprints may be widely utilized without additional special clearances. Reprints may be ordered *only* through the RCA Engineer, which controls and administers policies for their preparation and use.

NEW YORK CITY HIGH SCHOOLS BEGIN THIRD YEAR OF PARTICIPATION IN INDUSTRY-SCIENCE PROGRAM

More than 500 high school science students from all boroughs of New York City recently had an unusual opportunity to examine satellites and other space-age devices in the first of a lecture series designed to bring students into direct contact with the latest achievements in science. **Dr. T. Todd Reboul** of RCA brought with him to the lecture hall at Stuyvesant High School models of four satellites, an ion propulsion engine, a solar seeking mechanism and other instruments under development as part of the nation's space program.

That Saturday lecture marked the beginning of the third year of the industry-education program proposed originally by RCA Board Chairman **David Sarnoff** and launched jointly with the New York City Board of Education in 1962. Since that time, eleven other major industrial firms have participated in the program, and seminars and lectures have been held in Brooklyn schools in 1962-1963, Queens schools in 1963-1964 and will be continued at Stuyvesant High School during the current school year. The program is under the direction of Samuel Schenberg, director of the Office of Science Education, New York City School System. Other companies which will participate in the program during the current school year include Bell Laboratories, Esso Research, IBM, and U.S. Steel.

Editorial Representatives

The Editorial Representative in your group is the one you should contact in scheduling technical papers and announcements of your professional activities.

DEFENSE ELECTRONIC PRODUCTS

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B. Singer *Data Communications Engineering, Camden, N. J.*

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