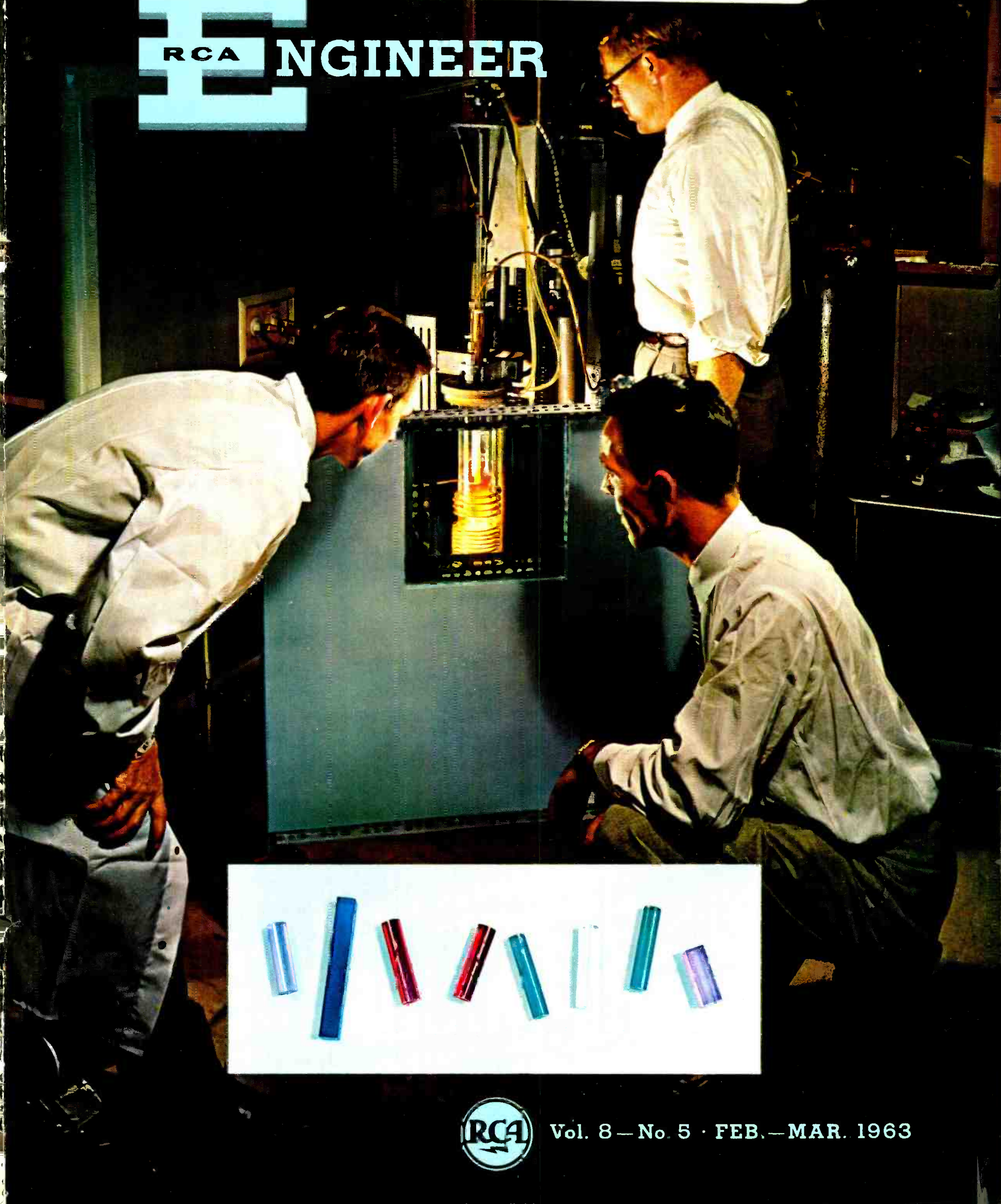
RCA ENGINEER



Vol. 8 — No. 5 · FEB.—MAR. 1963

OUR COVER

. . . shows Jordan S. Doyle (l.), Anthony G. Salerno (r.) and Donald R. Nichols (rear) of the Semiconductor and Materials Division, Somerville pulling a calcium fluoride laser crystal boule in an RF-powered, quartz-tube Czochralski crystal-growing furnace. The inset shows finished calcium fluoride laser rods; their doping is, l. to r.: dysprosium, samarium, thulium, uranium, samarium, neodymium, samarium, gadolinium.

LASERS . . . MASERS . . . RCA

There was a Mutt and Jeff cartoon that appeared several years ago in which Mutt was shown trying to coax Jeff into climbing the light beam of a powerful searchlight. Jeff never doubted he could do it, but demurred eventually on the grounds that Mutt would probably switch the light off when he got half way up.

In some ways, the development of *lasers* and *masers* has put the electronics industry in the same position as Jeff. We, too, are presented with a powerful light beam that, figuratively, we are sure we can climb. So strong is our faith, in fact, that unlike Jeff, we have not hesitated to clamber on board despite the many uncertainties we face.

By this faith, we are moving boldly to add to a remarkable tradition of accomplishment embodied in such modern wonders as radio communications, phonographs, computers, color television, and remotely controlled space vehicles.

With the *laser* and *maser*, we believe we can extend this tradition, creating high-speed optical computers, gigacycle communications networks, precise surgical and metallurgical tools, powerful radiation weapons, and instruments that call forth new phenomena or catalyze new materials.

To do so, however, will require more than faith and a tradition of accomplishment. The control of the electromagnetic spectrum above microwave and the complex way by which *lasers* and *masers* provide us access to it are both poorly understood. To succeed, therefore, we are going to require plenty of hard work in the fundamental as well as applied sciences.

In that connection, as this issue of the RCA ENGINEER so ably demonstrates, RCA is meeting its responsibilities vigorously. In Princeton, the Laboratories has recently developed a new rare-earth laser that can be driven continuously by natural sunlight. The Semiconductor & Materials Division is perfecting techniques to produce laser crystals of the highest optical quality. The Electron Tube Division is exploring sophisticated schemes to modulate lasers at infrared frequencies. Defense Electronics Products is developing radar, missile-tracking, communications, and computer systems based on laser principles.

Clearly, RCA is doing its share and can expect to reap its share of the benefits that may accure to this effort. Clearly, too, it will take a combination of all three attributes — faith, tradition and hard work — to transform the phenomenon of *L*ight *A*mplification by *S*timulated *E*mission of *R*adiation from a scientific fact to a vital force in the advancement of human society.

Dr. James Hillier
Vice President
RCA Laboratories
Radio Corporation of America

RCA ENGINEER

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PAPERS

NOTES

DEPARTMENTS

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

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

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

RESEARCH AND DEVELOPMENT is big business. The bill for this expanding industry in the United States during the 1960's is estimated to spiral to $120 billion. But R&D management has correspondingly become complex. In addition to the spiraling growth, other factors which have complicated management's job include the need to write off nonrecurring costs over a shorter period, increased competition, and the high dollar investment associated with R&D.

The administrative control of technical effort is perhaps the most difficult of all managerial tasks. Management is faced with the dilemma of producing R&D results on an economical basis. Yet, the degree of control that can be exercised is a function of the type of work being performed and the tools available for controlling and measuring progress. A need exists to improve the relationship between the accountant's general ledger and the engineer's slide rule. A delicate balance must be achieved between the diametric

The Engineer and the Corporation

THE CHALLENGE OF ADMINISTRATING TECHNICAL PROGRAMS

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forces at work—*the motive for planned, orderly business profit vs. the basically random, hard-to-predict nature of (hoped-for) R&D progress.*

As will be discussed herein, the *technical administrator* can supply a working relationship with the technical staff that provides management the opportunity to effectively monitor, yet does not mitigate the conditions necessary to the technical staff for creative professional work.

R&D—MAXIMUM FREEDOM, OR STRICT CONTROL?

The literature contains many explanations on how R&D should be administered. These theories have evolved over the years and to some extent substantiate management's concern.

One supports the premise that the scientist must be allowed *maximum freedom.* This theory rebels against administrative control with the claim that scientific achievements are not the function of time schedules, cost, or planning. This theory also suggests that the desire to create is frustrated when subjected to poor working conditions and excessive managerial control.

A second theory expresses the philosophy that R&D can be *effectively administered.* It supports the feeling that in order for technical management to make sound decisions they must have an awareness of schedules, costs, and profits. They must be concerned with the proper selection of projects. The attainment of profitable results within a reasonable time and for a reasonable cost is certainly a practical goal for technical management. Evidence exists to support the philosophy that management control techniques have been effectively employed in connection with development projects. (The POLARIS program is a prime example.)

Although a universal approach does not exist, a middle-ground theory, blending the best of the two discussed above, can provide a desired flexibility required to administer R&D effort. The *technical administrator* is an important key in implementing such an approach.

PROBLEM: THE TECHNICAL PERSONALITY VS. ADMINISTRATIVE RESPONSIBILITY

R&D progress *does not* flow in a regular or repeating pattern—and thus tends to be unpredictable. Change is the norm, and the resultant side effects can touch the entire organization.

In general, the nature of the technical personality results in little interest in the administrative viewpoint. It is the exception, rather than the rule, to locate an individual who performs effecively both as scientist and business manager. The logic and discipline developed through formal training and work experience in the application of the scientific method to a specialized field of technical knowledge is not easily transferred. The factors which inhibit technical personnel from generating ideas are not well defined. Concern has been expressed that management is making "organization men" out of their engineers and scientists—a concern more prevalent in R&D dealing in state-of-the-art techniques.

A hypothetical, but not-unfamiliar example will illustrate: Consider Joe Brown, a creative, talented engineer, who has ambition to advance in the organization. He is appointed responsible engineer on an important development project. Prior to this appointment, his primary concern was with details of complex technical problems. His selection was based on his recent MS degree, the discovery of several significant patents, and his ability to work effectively with his associates. The results of preliminary investigations were extremely favorable and a substantial development contract followed.

Success appeared inevitable. As the project proceeded Joe was called upon regularly to discuss progress of the effort with his management—matters of schedules, expenditures, cost to complete, and technical problems. Joe began to realize something to which he had given only little consideration over the last few months. Before his appointment, he had particularly enjoyed contact with complex technical problems and the challenge of developing effective solutions. That continued exposure to the discipline of scientific work and the satisfaction derived from converting theory and innovation into hardware had always stimulated him. Now, although his management was satisfied with the project's progress to date, Joe was dissatisfied. For example, several technical questions directed to him during administrative meetings made him aware that *he was not really close to the project technically*. His time was being consumed by a myriad of meetings, progress reports, re-estimates, coordination with the project engineer, and other administrative tasks associated with keeping the job on the track. Because time had to be allocated to all these activities, he was no longer making the needed technical decisions that he was both competent to make, given the time, and that he actually found more satisfying. And, the future promised more of his time taken by administrative problems. Unfortunately, Joe found himself disliking what he felt was now a role of shuffling papers, keeping superiors and subordinates happy, and developing magic numbers. *Is this, then, the best role for Joe?*

While the above example may overemphasize the dilemma for the sake of making a point, such shifts in attitude and difficulties in best utilizing talent are today very real problems for R&D management. Effort expended by scientists and engineers on administrative matters *reduces the overall technical capacity of an organization.* Granted, the technical man must have both respect for and an awareness of the business aspects of his task. But, it is also reasonable that the technical staff should not be administratively burdened to the point where the organization experiences a loss in creative technical output. Thus, management must always ask itself *whether they can afford to use their technical staff to monitor and control R&D.*

FINDING ANSWERS: THE TECHNICAL ADMINISTRATOR's ROLE

The techniques employed to collect, analyze, and forecast costs, schedules, workload, and manpower, and to generally administer a technical task are sufficiently complicated to warrant the use of individuals trained and skilled in their application. Varied titles have been used. Herein, *technical administrator* will serve to describe the function as a composite. Regardless of specific job description, the technical administrator assists his operating management with their job of effectively administering engineering and scientific work. The important overall project responsibility and authority remains within the framework of the technical hierarchy. However, the degree of participation in administrative details by technical personnel can be limited through intelligent delegation.

An effective administrator can substantially relieve scientists and engineers of nontechnical activities, thus making them available for technical assignments more suitable to their education, experience, and interests. Without undermining the structure of line management, and with guidance on technical details, the technical administrator can minimize red tape and make R&D effort operate in a much more productive manner.

His field of knowledge is necessarily broad, yet oriented to technical effort. Although not a specialist like the pricing expert, the purchasing agent, or the contract administrator, *he will utilize an intimate knowledge of all operating functions*—including cost estimating, proposal preparation, scheduling, manpower planning, business forecasting, value engineering, and techniques for recording, monitoring, and controlling costs. All this is necessarily influenced by a basic appreciation and understanding of the nature of engineering and scientific thought and the individuals associated with it. With such knowledge and an ability to communicate effectively, he can truly assist the engineer and also accomplish his management role.

A CHALLENGE FOR ENGINEERS: TO PROFIT FROM THE POTENTIAL OF THE TECHNICAL ADMINISTRATOR

By nature, technical personnel appreciate the ability to eliminate red tape and will tend to attach greater significance to those functions of the administrator that effectively do this. While the administrator must appreciate operating situations and the stifling effect that rigid controls have on the creative process, proper control and procedure are essential management tools. Thus, the good administrator will work to instill in the technical staff not just an appreci-

WILLIAM A. THACKER received his BS in 1955 from the University of California, Berkeley, and his MBA from the University of Southern California in 1960. He has seven years experience, including two with the DEP Data Systems Division. His experience includes manufacturing engineering, industrial planning, industrial engineering, organization planning, and technical administration in engineering and research organizations. At present, he is responsible for assisting the Manager of the Digital Development and Design Engineering in the coordination of internal administration, cost, schedule, manpower planning and control of engineering tasks, and the development and implementation of internal operating procedures. At Data Systems Division, Mr. Thacker has participated in and has been responsible for the project administration of several military electronic computer projects including the MIPIR Computer. He participated extensively in the development of the RCA 4100 Product Plan. Prior to 1961, Mr. Thacker worked at the Atomics International Division of North American Aviation, Inc. as a member of the Organization Planning Staff. He was responsible for conducting operational evaluations of individual operating functions and related management control systems. In addition, he participated actively in the Division's Cost Reduction Program. Prior to 1957, he worked at North American Aviation, Inc. as an industrial engineer. The title of Mr. Thacker's Master Thesis is "Development, Implementation, and Results of a Program to Reduce Overhead Costs in a Medium-Size Research Organization".

ation of his obvious value (less red tape) but an intelligent understanding of the real purpose for intelligent controls. In other words, the administrator is regularly "selling" his programs.

The Technical Administrator—by design—endeavors to anticipate problems, not just wait for them. In developing practical solutions, he makes constant use of knowledge of the profit motives and basic objectives of the company. In so doing, he can develop for the technical staff an understanding of and cooperation with other operating functions.

To further understand the role and the potential of a technical administrator, consider the following set of basic attributes—a sort of *modus operandi:*

He must be capable of operating as much on his own as possible, with a minimum of direction, and with consistent and reliable results.

An effective working relationship with the technical staff is the most important factor in determining the technical administrator's success in an organization—a matter of confidence and rapport. A professional interest in administrative problems is a start. A foundation of successful experiences over a period of time is the formula for establishing rapport. Engineers should expect the administrator to use thorough approaches and provide recommendations that are based on sound logic.

The technical administrator must be sensitive to individual personalities.

Because the technical man may not always understand what the administrator is attempting to accomplish—it may appear to be red tape—the administrator must be prepared to defend his position in a positive and logical fashion. Mere reference to an operating procedure is not adequate.

New methods may be required to solve problems. Often the administrator must function as a management consultant.

Take a quick look around your organization. Do you see someone in Joe Brown's situation? Let's assume that a technical administrator had been assigned to Joe. Many examples could be provided to demonstrate the advantages gained. An example might be the establishment of a detailed working schedule. Given the overall parameters, knowing the contract requirements, and being familiar with the nature of the job, the technical administrator with inputs from Joe could prepare the schedule for his review. If an automated technique like PERT is involved, most of the details could be provided by the administrator. Periodic re-estimates, because of changes in scope, and proposals for new jobs can be made easier for the technical man with the help of a competent administrator.

What type of individual makes a competent technical administrator? Ideally, he has a formal education and work experience in the fields of engineering or science and business administration or economics. A flexible personality equips him for the frequent and rapid changes that take place in R&D. The lack of specific job definition that must exist in order for his function to serve a useful purpose. His success will depend a great deal on the communication skills he has developed. He must have the ability to quickly grasp what the technical man wants and to provide an adequate and timely solution. The administrator should have the capacity for understanding the type of technical work being done so that he can intelligently discuss job progress, plan future workload, order materials, etc. In performing his assignments, he must continually search for modern administrative techniques to make the technical man's job easier.

IN CONCLUSION . . .

A *professional* administrator can assist management in planning, organizing, directing, coordinating, and controlling engineering and scientific effort. It is not intended that such an individual be useful at all levels but his existence in some form within the line organization provides a means for effectively administering technical projects.

Management by exception is an important principle to be followed in the use of the technical administrator. *The benefits to be derived are a function of the status and recognition afforded this important task by management.* Acceptance must not be forced on operating functions but like the professional engineer, the position must not be prostituted. A set of hard and fast rules for the technical administrator does not exist, *nor should they be developed.* Rather, it is the effective, creative application of sound, flexible principles that the Technical Administrator proves of greater value to the technical staff and to management.

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TWO RCA MEN ELECTED IRE FELLOWS

The two RCA men appearing on this page have been honored for their professional achievements by being elected *Fellows* of the Institute of Radio Engineers. This high honor is bestowed each year by invitation upon those who have made outstanding contributions to the field of electronics.

... for contributions to the development of electron tubes

JOSEPH T. CIMORELLI received the B.S. degree in 1932 and the M.S. degree in Electrical Engineering in 1933 from the Massachusetts Institute of Technology. He joined RCA in 1935 in Harrison, N. J., as a member of the Receiving-Tube Application Laboratory. From 1940 to 1944, he worked as a field engineer, and in 1944 was appointed Manager of the Receiving-Tube Application Laboratory in Harrison. He transferred to Camden in 1953 as Assistant to the Vice-President and Director of Engineering of the RCA Victor Division, and later became Administrative Engineer on the Staff of the Vice President, Product Engineering. In 1956 he returned to Harrison as Manager of Engineering for the Receiving Tube Operations Department of the Tube Division. In 1958, Mr. Cimorelli became Manager of Receiving Tube Manufacturing in Harrison, Woodbridge, Indianapolis, and Cincinnati. In 1960, he resumed the position of Manager, Engineering, Receiving Tube Operations. Prior to being named Fellow of IRE he was a senior member. Mr. Cimorelli is also a Fellow of the Radio Club of America.

... for contributions to military communications

SIDNEY METZGER received his B.S.E.E. from New York University in 1937, and his M.E.E. from Polytechnic Institute of Brooklyn in 1950. From 1939 until 1945 he worked for the United States Army Signal Corps Laboratories at Fort Monmouth on radio communications equipment. From 1945 to 1954 he was at the ITT Laboratories as division head in charge of commercial and military multiplex microwave systems. In 1954, Mr. Metzger joined the RCA Laboratories at Princeton, N. J., where he worked on communications problems for military systems. He transferred to the Astro Electronics Division when it was formed in 1958. He was in charge of the group which supplied the satellite and ground based radio equipment for Project SCORE ("the Talking Atlas") in 1958; was responsible for the communications system and equipment for TIROS; and is now responsible for the communications system and equipment for Project RELAY. He is a Fellow of IRE, a Senior Member of the American Rocket Society and a member of Tau Beta Pi and Sigma Xi. He is a member of the Communications Committee of the American Rocket Society and of the United States Committee for Study Group IV, CCIR (on space communications).

[*Ed. Note: Mr. Cimorelli helped pioneer the early plans and policies for the RCA ENGINEER. Both Mr. Metzger and Mr. Cimorelli have authored professional papers for publication in the RCA ENGINEER.*]

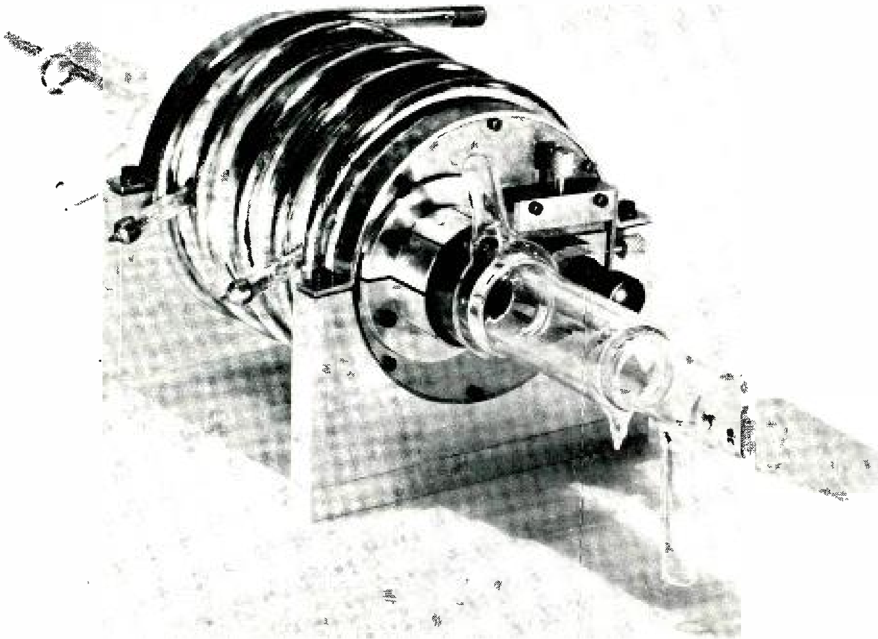

Fig. 1—An elliptical cavity for testing continuously operating solid-state lasers. The emitting crystal is placed at one focus, and the "pump" lamp is placed along the second focus of the elliptical cylinder. The glass tubing is a dewar which may be used to circulate a liquid coolant. The emitted radiation emerges along the axis of the dewar.

THE LASER—AN INTRODUCTION

In discussing lasers, an analogy might be made: When the transistor was announced, its applicability was immediately evident. It was clear from the start that this device could perform the function of a vacuum tube and thus replace it for many purposes. The laser, on the other hand, is an entirely new tool looking for work. So far, it has been considered primarily as a replacement part in systems already in operation using lower-frequency generators. It is to be hoped that in time it will more and more interest creative engineers who will invent entirely new applications, particularly suited to its novel properties. This paper provides a brief insight into those properties, as an introduction to the several papers on this field that follow.

DR. H. R. LEWIS

RCA Laboratories
Princeton, N. J.

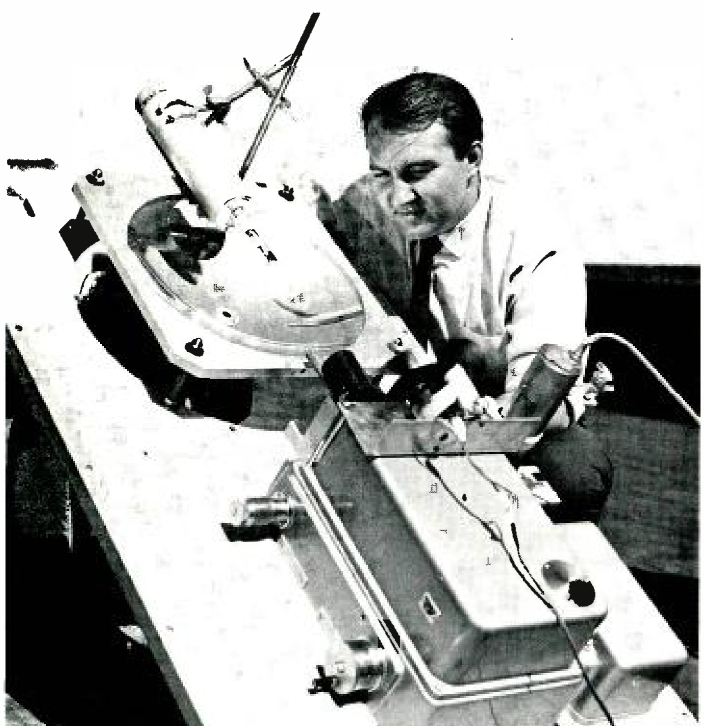

Fig. 2—Dr. Z. J. Kiss and the sun-pumped laser. The emitting crystal is mounted in the tip of the dewar, and the dewar is adjusted so that the crystal lies just beyond the focal point of an elliptical reflector directed at the sun. The emitted beam travels along the axis of the dewar into a detector.

PROGRESS IN TECHNOLOGY usually consists of a series of small improvements in well-known techniques. Making these improvements requires a great deal of hard work and ingenuity. Moreover, one may expect the effort expanded per unit improvement in any one field to increase as time passes because the easy inventions have all been made. Thus, the uncovering of an entirely new technique is greeted with considerable excitement.

Those who are most excited are, of course, the scientists and engineers who can see in the new development an easier road to other new achievements. Ultimately, when the initial flurry dies down, the real achievements can be separated from the sham, and the importance of the new invention can be properly assessed. That time is still a year or more away for one of the most recent technological innovations—the optical maser, or as it is also called, the *laser*. Yet, one can feel even now that most of the attention the laser has received to date is justified for at least two reasons: 1) it performs an entirely new function: the generation of coherent electromagnetic energy in the visible and infrared portions of the spectrum where coherence was previously unattainable and 2) it uses physical principles which were not exploited for practical purposes until very recently. A serious discussion of these two points must necessarily make large demands on the reader, the editor and the author. Fortunately, this is unnecessary here for the literature contains several detailed articles on lasers and masers[1-3] which the dedicated can read at their leisure. Here, we need only state how the laser differs from the old, familiar light sources, indicate why these differences may be significant for applications, and give a general notion of the operating principles involved.

COHERENCE AND ITS SIGNIFICANCE

In what way is the output of a laser significantly different from the output of, for example, a high intensity mercury arc? The answer lies in a careful description of the nature of the electromagnetic radiation characterized as light. Normal light sources—the mercury arc, or an incandescent light for that matter—produce an electromagnetic field which is irregular and unpredictable. The field at any point in space has a sinusoidal time dependence only for times much shorter than a microsecond. Then, an abrupt change occurs, and a new sinusoidal variation begins. This is in contrast to the situation for radio waves where the regular sinusoidal variation continues for long periods of time. This regularity is known as *coherence.* The laser output differs from that of conventional light sources by virtue of the high degree of coherence it displays. In fact, the laser produces electromagnetic waves that rival radio waves in their regularity.

Why is the coherence of laser radiation so important for certain applications? Crudely speaking, one can relate the coherence of a generator to the width of the frequency band which contains a significant portion of the energy emitted by the generator. (Here we are speaking of the character of the radiation *before* modulation). The relation is inverse: that is, the greater the coherence, the narrower the emitted band. Thus, one can change the question to read: *Why is the spectral purity of the laser so significant for applications?* There are three, simple, important answers. First, the spectral purity allows one to filter out noise in the receiver of any communications or radar system. To give an example, the filter for a typical laser need only be approximately 1 kc wide (or the width determined by the modulation rate if that is larger), but a filter designed to accept most of the energy in the "sharp" green line of a mercury arc must be 10^{10} kc wide. Clearly, the receiver must accept much more noise if the conventional source is used. A related point is that conventional heterodyning techniques may be used with the coherent laser, while they are ineffectual with normal light sources. Heterodyning is a convenient technique for limiting the bandwidth of the receiver, and the local oscillator adds power that helps in other ways to overcome noise. Finally, the coherence of laser radiation enables one to form the radiation into beams which may be as narrow as the diffraction produced by the largest optical element involved in the system. (The usual example: A laser on earth using a 4-inch lens will illuminate an area only two miles in diameter at the moon). By way of contrast, in a system using a conventional light source, the beam angle increases as the area of the emitter increases; thus, one cannot easily increase the intensity of radiation received at a distant point by increasing the area of such a source. The same coherence properties of a laser which provide such a narrow beam permit a condensing lens to focus the energy into an area with dimensions of a few wavelengths of the light, thus producing energy densities and electric field strengths orders of magnitude larger than those previously available. The technological and scientific consequences of this feature may well prove to be the most important of all.

HOW THE LASER WORKS

Finally, we come to the question of how the laser works. (The general principle underlying the maser is the same.) It may clarify the situation to compare and contrast the operation of lasers to the familiar process of fluorescence.

In fluorescence, individual atoms in a gas, liquid, or solid absorb energy from some source (frequently ultraviolet light) and then re-emit the energy, usually at a longer wavelength. Each of the atoms, and there may be 10^{15} to 10^{19} of them per cubic centimeter, emits independently and spontaneously; the light thus produced emerges in all directions in a broad frequency region. However, an atom can also emit its energy because of stimulation by an electromagnetic field of the appropriate frequency (i. e. the same frequency at which the atom would eventually emit spontaneously). Thus, the emission of light by one atom can cause the emission of light by another atom thereby amplifying the electromagnetic field. Providing reabsorption and other losses do not overcome the gain produced by the stimulated emission, we can then visualize a growing electromagnetic wave propagating through a fluorescing material. This amplification process, caused by a kind of cooperative fluorescence of many atoms, can be converted into a self-triggered oscillation—a laser—in the conventional way, that is, by introducing feedback and a resonant circuit. In a laser, the feedback and resonant circuit may be produced by placing the emissive material between two parallel mirrors. Useful power is coupled out by making one of the mirrors semi-transparent (Fig. 1).

Lasers of this kind have now been made using solids (ruby, calcium fluoride, or calcium tungstate, suitably doped with fluorescing centers; semiconductor diodes) and gases (helium-neon mixtures, cesium, etc.). Some are excited optically, some by radio-frequency discharges, some by passing DC currents through semiconductor diodes. They deliver energies in the range of 1 milliwatt to 1 watt at frequencies ranging from the ultraviolet to the near infrared.

RCA Laboratories has made a number of contributions of its own to the development of solid-state lasers. It has developed several new lasers including[4,5] Er^{3+} in $CaWO_4$, and Tm^{2+} in CaF_2. A relatively high power, continuously operating laser has been produced,[6] using CaF_2:Dy^{2+}, and a continuous laser using sunlight as the pump source has been demonstrated by RCA Laboratories—*the first device of its kind.* [See Fig. 2 and *Engineering & Research Notes,* this issue.]

CONCLUSION

Despite this progress, much improvement in lasers is still needed if this new field is to measure up to its advance publicity. The chances that such improvements can be made still seem large after several years of work in this area.

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DR. HENRY R. LEWIS received A.B., M.A., and Ph.D. degrees from Harvard University in 1948, 1949, and 1956 respectively. In 1950 and 1951 he was a teaching fellow in physics at Harvard. His Ph.D. thesis was a study of distortions observed in nuclear resonances using the molecular beam technique. From 1951 to 1953, and for a year after receiving the doctorate, he worked with the Operations Evaluation Group of the Massachusetts Institute of Technology on various problems in naval warfare. Dr. Lewis joined the Technical Staff of RCA Laboratories in 1957 and has worked in the fields of masers and paramagnetic resonance since that time. He is now head of the Quantum Electronics Group of the Electronic Research Laboratory. The present projects of the group include masers, lasers, and cyclotron resonance.

Some of the multiple problems in producing a successful laser system are: modulation of the beam, detection of the modulation, shot noise, continuous operation, heat dissipation, and operating frequency. Progress in solving the last three of these problems depends upon the proper choice of laser materials—the subject of this paper.

RESEARCH ON LASER MATERIALS

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RCA Laboratories

Princeton, N. J.

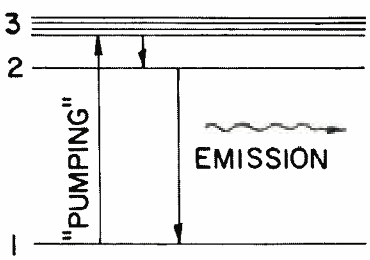


Fig. 1—Three-level energy scheme.

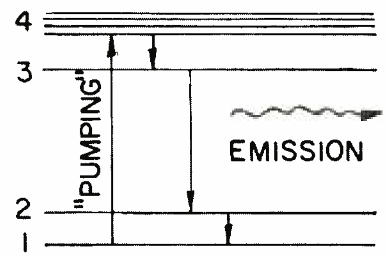


Fig. 2—Four-level energy scheme.

THE CONCEPT of the maser (and thus also the laser) was suggested in 1954 in a letter to the editor of *The Physical Review* by Gordon, Zeiger and Townes[1] in which they presented the idea of "*A*mplification by *S*timulated *E*mission of *R*adiation." Acronyms coined from these initial letters create *maser*, with *m* for microwave, and *laser*, with *l* for light. (The latter is, of course, also called the optical maser.)

The laser idea has opened up numerous potential applications: for example, narrow-beam, multichannel communication, long-distance communications (to the moon, etc.), spot welding, cutting of tissues for medical purposes, and photocatalysis of chemical reactions. These depend upon the nature of laser output—a very-high-energy, highly directional, coherent beam of light.

CRITERIA FOR LASER MATERIALS

Materials for use in solid-state lasers must fulfill several criteria. From a qualitative point of view they should exhibit the appropriate electronic energy level scheme, should be capable of being fabricated into a mode selecting structure, and should be of suitably high optical quality to minimize optical losses. Also the material should exhibit a high luminescence efficiency and the correct relationship between lifetimes of states.

These criteria have been met by using as an emitter an ion that is introduced as an impurity into a host having the necessary optical and mechanical properties. The emitting systems employed to date for lasers have one of the energy-level schemes illustrated in Figs. 1 and 2.

Three-Energy-Level Scheme

Fig. 1 is a three-level system and represents such real systems as ruby (chromium doped aluminum oxide). When this system is not excited, all the electrons are in level *1*; on illumination of the material, electrons are excited or pumped to level *3*. Level *3* is shown as a band, rather than a line, to indicate that a wide range of frequencies can be used to pump electrons to this state. These excited electrons decay to state *2*, which is a line, indicating that when the electrons go through the radiative transition from state *2* back to state *1*, the radiation emitted has a very small frequency range or is monochromatic.

If the life time of state *2* is sufficiently long to allow electrons to be supplied to it at a rate faster than that at which they leave it, a population inversion will finally result, at which time more electrons will be in state *2* than in state *1*. It is at this time that emission by stimulation can occur, the stimulation being accomplished by photons having the same energy as the difference in energy between states *2* and *1*. It is now apparent that these systems are self-stimulating and that if there is a long optical path in the laser structure, very-high-energy beams can be obtained. A long optical path may be formed by a rod with optically flat, parallel ends as the laser element and silvering both ends—one end 100-percent reflecting and the other about 98-percent reflecting.

The foregoing is a simplified explanation of the operation of a ruby laser; however, ruby is *not* the best possible material for a laser element. Its defect is that the luminescent transition (level *2* to level *1*) terminates at the highly populated ground state (level *1*), which has to be depopulated to obtain the inversion necessary for stimulated emission. This depopulation of the ground state requires very high pumping energies.

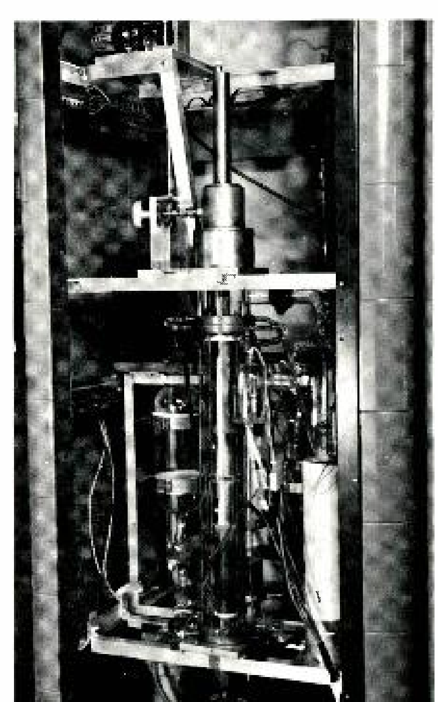
Four-Energy-Level Scheme

The four-level scheme for a more efficient laser element is shown in Fig. 2. This scheme differs from Fig. 1 in that the radiative transition is between levels *3* and *2*—the terminal state (level *2*) for this transition is not the ground state. This additional level (level *2*) obviates drastic depopulation of the ground state to produce a population inversion between levels *3* and *2*. Therefore, stimulated emission can be obtained with very much lower pumping energy.

The four-level scheme is approximated in reality by the divalent rare-earth ions excluding europium and ytterbium. These ions are difficult to prepare because of their highly reducing nature and have been known for only a short time.

The trivalent rare-earth (lanthanide) ions in a number of hosts have shown laser action. Their level scheme differs from the scheme of Fig. 2 in that level *4* is not a broad, continuous band, but is a composite of sharp absorption lines. This sharp-line nature of the absorption process means that existing light sources do not match the absorption spectrum of these materials well, and very high intensity lamps must be used to absorb a useful amount of energy into the sharp absorption lines. One way of overcoming this disadvantage of the trivalent lanthanides is to provide another adsorbing process in the host material in such a way that energy absorbed by this second

Fig. 3—Vertical-vacuum Bridgman furnace.



process is transferred to the rare-earth ion from which it is emitted. Included among the possibilities for producing this desired absorption band are doping with a second impurity ion, making use of lattice absorption of the host, or using a metallo-organic compound.

Host Material

Upon selection of an ion with the necessary emitting properties, this ion must be incorporated into a host which will show suitable optical properties and will allow fabrication (such as cutting and polishing) into the necessary shape. The host should be stable under the operating conditions. Since the trivalent lanthanides are very stable ions, the principal considerations, in addition to those mentioned above, are those of matching the ionic size and charge of the lanthanide to that of the host. These reasonably simple requirements have led to the use of calcium, strontium, and barium fluorides, tungstates and molybdates as crystalline hosts. Also the trivalent lanthanides have shown laser action in glass matrices.

Since the divalent lanthanides are very powerful reducing agents, the host should be resistant to having their component ions reduced.

PAUL G. HERKART received the BSEE and MSEE in 1935 from the Massachusetts Institute of Technology. From 1935 until 1943 he was engaged in power-tube manufacturing at the RCA Victor Division in Harrison, New Jersey. Since 1943 he has been with the RCA Laboratories, Princeton, New Jersey. Mr. Herkart's early work at the Laboratories was on television pick-up tubes, storage tubes, and color television. He later became interested in materials and lead a group working in the technology and production of semiconductor materials, thermoelectrics, associated alloys, and other materials with diverse uses. Mr. Herkart is now Head of the Materials Technology Group at the Laboratories. Present work for this group continues on semiconductor compounds and other alloys and compounds, and has expanded into thin-film production of cadmium sulfide, epitaxial growth of gallium arsenide, phosphors, and laser materials. Mr. Herkart is a Member of the IEEE and of Sigma Xi. He has received three RCA Achievement Awards. He is author and co-author of several papers.

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PREPARATION OF LASER MATERIALS

The steps in preparing a finished laser element involve the synthesis and doping of the host material, and growth of a single crystal (or molding an amorphous host). These will now be discussed. (The final steps of cutting and polishing, and coating of faces will not be covered.)

Chemical Synthesis

The materials that have received the most study at RCA Laboratories are calcium tungstate and structurally related compounds, and the alkaline earth fluorides and halofluorides. These materials have all been prepared by molten-salt or solid-state reaction. Since the materials composed of alkaline earth fluorides must be free of oxygen contamination, special precautions have to be taken in their preparation.

The divalent lanthanides were considered to represent a class of emitting ions of great interest because of their favorable absorption bands. The normal valance state for a lanthanide is *3*, except for europium and ytterbium in which it may be *2* or *3*. The remaining members of the series exhibit large negative-reduction potentials for the trivalent to divalent couple and, therefore, can exist only in hosts containing ions having even larger negative-reduction potentials — i.e., materials which are even more difficult to reduce. This consideration together with requirements for charge compensation and for ionic diameter matching indicates that suitable cations may be those of the alkaline earths.

To prepare an alkaline-earth fluoride doped with rare-earth ions, single-crystal optical-grade alkaline-earth fluoride is melted together with the appropriate quantity of rare-earth fluoride in an inert atmosphere. Since single-crystal optical-grade alkaline-earth fluoride is not always commercially available, satisfactory material may be synthesized by reacting the appropriate alkaline-earth chloride with potassium bifluoride in a melt. The rare-earth ions introduced into these fluorides are found to be primarily in the trivalent state. With the exceptions of ytterbium and europium, as mentioned, samarium is the most easily reduced; it can be obtained in the divalent state by melting in a hydrogen atmosphere or in the presence of carbon. The former process is the more desirable, as it avoids inclusions of very small carbon particles. Thulium can also be reduced in a like manner, though not so readily or completely. The remaining rare earths are reduced by remelting the trivalent material in the presence of an appropriate quantity of rare-earth metal or through the use of photo-reduction techniques employing intense gamma radiation. At present, it appears as though a purer and more homogeneous product is obtained through photo-reduction than by means of metallic reduction.

The alkaline-earth halofluorides, (MXF), where M is an alkaline earth and X a halide other than fluoride, are chemically stable and possess tetragonal symmetry. These compounds are synthesized by fusing together equal molar portions of halide and fluoride. X-ray analysis has shown a homogeneity range below the limit of detection; also, the melting points are from 900 to 1000°C.

The alkaline-earth tungstates, molybdates, and vanadates doped with trivalent rare earths are well known as luminescent materials. While these compounds will not support the more desirable divalent state of the lanthanides, they are of interest as hosts of trivalent ions because they exhibit laser action either through "line pumping" or by pumping in the lattice absorption region. A further advantage of these materials is that crystals possessing good optical and thermal properties can be prepared.

These compounds are readily prepared in air at 1000 to 1200°C by reacting the alkaline-earth carbonate, hydroxide, or oxide with MoO_3, WO_3 or V_2O_5. These materials are prepared as powders or polycrystalline masses. The trivalent rare-earth ion is incorporated by substitution for the divalent alkaline-earth ion. The imbalance of charge resulting from this substitution is compensated for by either vacancy formation or the incorporation of a monovalent ion. There is some spectrometric evidence that the

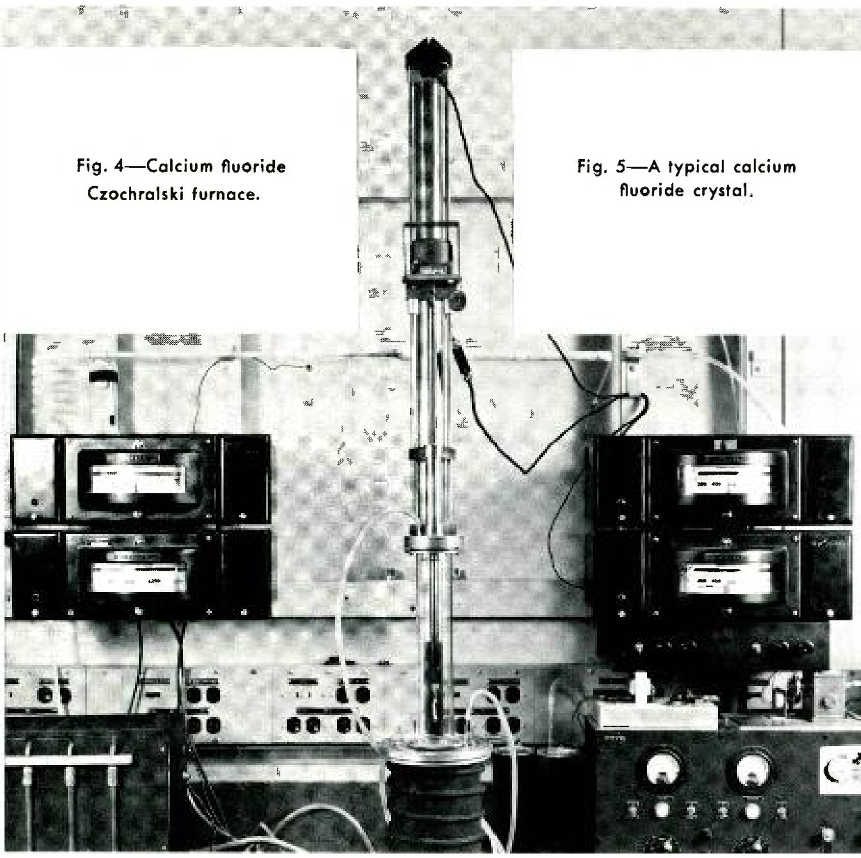

Fig. 4—Calcium fluoride Czochralski furnace.

Fig. 5—A typical calcium fluoride crystal.

Fig. 6—Calcium tungstate Czochralski furnace.

monovalent ion compensation produces a better material.

Single-Crystal Growth from a Melt—Bridgman Technique

Single crystals of materials for use as lasers have been grown from the melt by two methods: the *Bridgman* (now to be described) and the *Czochralski* (covered in the next section).

In the Bridgman technique, the material to be crystallized is contained in a crucible, the bottom of which is tapered to a point or a small capillary. The crucible is supported in a vertical cylindrical furnace which is generally constructed with two zones of different temperature connected by a thermal gradient, the slope of which is adjusted as required for the material under investigation. The upper zone of the furnace is held at a temperature above the melting point of the material, while the lower zone is held at a temperature below the melting point. The crucible is slowly lowered from the upper zone to the lower zone of the furnace and then slowly cooled to room temperature. A modification of this technique is a system in which the temperature gradient is moved while holding the crucible stationary. A second modification is a system which operates in a horizontal plane. A photograph of a typical Bridgman furnace designed specifically for operation under inert gas atmosphere or high vacuum is shown in Fig. 3.

Research efforts led to the discovery of a number of compounds of the general type, MXF, where M = Ca, Sr, Ba, and X = Cl, Br, I. This class of compounds would be suitable hosts for divalent rare-earth cations and are not hygroscopic. Small laminar single crystals of calcium chlorofluoride and strontium chlorofluoride have been prepared by the vertical and horizontal Bridgman techniques. Larger single crystals of divalent rare-earth-doped barium chlorofluoride and barium bromofluoride have been grown. These crystals are adequate for spectrometric investigations. The barium halofluoride crystals appear to be little affected by moisture; however, the micaceous characteristics of these crystals may present an optical polishing problem.

Single-Crystal Growth from a Melt—Czochralski Technique

In the second method that has been utilized to prepare laser single crystals from the melt, the Czochralski technique, a suitable seed is dipped into a melt and the seed is withdrawn at the same rate as the crystal is grown. The diameter of the growing crystal and the rate of growth are largely dependent upon the temperature gradients and material under investigation. Single crystals of calcium fluoride, barium fluoride, and strontium fluoride are very easily pulled by the Czochralski technique, as are the materials calcium tungstate, calcium molybdate, strontium tungstate, strontium molybdate and lead molybdate. The problems encountered in growing the fluorides, however, are different from those encountered in growing the tungstates and molybdates.

The melting point of calcium fluoride is 1402°C. This limits the possible crucible materials. Another problem associated with calcium fluoride is the presence of oxygen. The slightest amount of oxygen-containing material dissolved in the melt reacts to form calcium oxide, which precipitates in the growing crystal as very fine inclusions, causing Tyndall scattering of the light. Carbon and other inclusions will also result in optically imperfect crystals.

The furnace (Fig. 4) employed for pulling these crystals must be tightly closed to eliminate moisture and oxygen from the air. At RCA Laboratories, pullers originally designed for silicon single-crystal growth have been used. They employ a resistance heater made of graphite. The graphite heater, the crucible (a molybdenum shell encased in thick graphite), and the seed holder are enclosed in a water-jacketed steel cylinder. The pulling and rotating mechanism passes through a doubly gasketed Teflon seal located well out of the hot zone. Because alkaline earth fluorides are active scavengers for oxygen, it is necessary to keep quartz, alumina, and other oxides out of the hot zone. After loading, the furnace is evacuated, moderately heated, and then flushed with pure helium, or helium with added hydrogen. These gases are purified by a series of absorbents, catalysts, and cold traps.

The purified gas at just above atmospheric pressure is used during crystal growth. The furnace must be leak-tight, otherwise enough air to cause Tyndall scattering-centers in the crystal will diffuse in against the pressure within the furnace. As additional insurance against oxide contamination, lead fluoride is incorporated in the melt.

The furnace is purged and then slowly brought up to temperature. When the lead oxide and fluoride have completely volatilized the seed is dipped into the melt and slowly withdrawn. A typical CaF_2 crystal is shown in Fig. 5. Calcium, strontium, and barium fluoride single crystals doped with lanthanide, and rare earth-alkali metal cations have been prepared.

The Czochralski furnace (Fig. 6) used in growing single crystals of the tungstates and molybdates differs from the furnace utilized for preparing the alkaline earth fluorides. The former compounds lose oxygen readily and must be pulled in an atmosphere containing oxygen. The melting point of these materials is above 1600°C. This temperature is too high for a platinum crucible, and iridium or rhodium containers must be used. In this method, RF heating of the crucible is employed. The tungstates and molybdates are subject to strain, and are annealed in air. To achieve partial annealing during the growth process, a thermally insulating wall is built up outside the crucible. This tends to protect the crystal from drafts and to retain the heat around it. The usual crystal is ½ inch diameter and 3 inches long. Single crystals of laser quality of calcium tungstate, and of calcium, strontium, and lead molybdate activated with all the trivalent lanthanide elements have been prepared.

Laser action has been observed in a number of these systems at RCA Laboratories[2] including, for the first time, laser action in trivalent erbium-doped calcium tungstate.

[*Editor's Note:* As a result of cooperative effort between the RCA Laboratories and the Semiconductor and Materials Division, Somerville, N. J., several of the laser materials mentioned in this paper are available from the latter Division.]

Single-Crystal Growth from High-Temperature Solutions

Single-crystal growth from high-temperature solutions is a technique which has proved useful in systems where no seed crystals are available—and more important, in systems in which there is a phase transition below the melting point. The latter condition precludes crystal growth from a melt, making it necessary to resort to a method in which crystal growth can occur below the phase transition temperature.

The essential features of high-temperature-solution single-crystal growth are identical with the familiar aqueous-solution method. The only important differences are in the use of higher temperatures (500 to 1500°C) and in the solution medium—usually a salt that melts at high temperatures, such as the alkali-metal halides, and several carbonates, fluorides, and oxides. Single-crystal growth is promoted by decreasing the solubility of the solute in the solvent either by slow evaporation or by slowly cooling the solution. It was felt that a more efficient calcium tungstate structural-type host could be found in materials in which the trivalent rare earth cation presents a less drastic disturbance in the crystal lattice. Spectrometric studies of mixed alkali metal—rare-earth tungstates indicated that the luminescent efficiency of trivalent rare earth activators was considerably higher in lithium-lanthanum tungstate, $(Li_{0.5}La_{0.5})WO_4$, than in calcium tungstate. Small single-crystal platelets and fine needles have been prepared by slowly cooling a high-temperature lithium chloride melt.

Aluminum oxide (sapphire) and chromium doped alumina (ruby) single crystals are grown by the conventional flame fusion technique. Chromium, copper, erbium, thulium, and ytterbium doped aluminum oxide thin plates have been prepared from a high-temperature flux which is cooled very slowly. Single crystals of gallium oxide containing chromium, europium, and samarium as minor impurities were also prepared in this manner. These crystals were in the form of needles or thin laminar plates.

OTHER POSSIBLE MATERIALS

Investigations have also been carried out on a variety of materials, other than those already mentioned. For the most part they consist of glasses, complex fluorides, and chelates.

From the considerations of optical quality and the elimination of the need for single crystals, glasses are obviously attractive as hosts for lasers. The major disadvantage of present glass lasers is the broad spectral lines resulting from the low degree of atomic order characteristic of glasses. Studies are underway to find means of decreasing the line widths of the emissive states. Some difficulties have been encountered in obtaining homogeneous material from small-size melts. Up to the present, it has been possible to introduce only trivalent lanthanides into glass systems.

One method for obtaining lasers containing other trivalent lanthanides is to induce broadband rather than line pumping. One way of accomplishing broadband pumping is to combine the lanthanide with an organic molecule to form a quasi-aromatic system (chelate) involving the lanthanide ion. In such a ring configuration, the absorption can take place in the broadband of the organic portion; the energy can then be transferred to the lanthanide for emission. Preliminary studies have demonstrated that this energy transfer does take place. A serious problem for the construction of laser systems arises from the fact that chelates absorb very strongly and hence must be diluted with a transparent substance. Present experiments employ, as the diluent, plastics or organic solvents which are frozen to form a glass. Such glasses exhibit much narrower emission lines than do inorganic glasses.

A number of fluoride compounds containing two cations can be prepared and doped with either di- or trivalent lathanides. These materials are stable in the atmosphere and offer the lanthanides crystallographic sites of high coordination which they prefer.

CONCLUSION

Numerous promising materials for laser devices have been prepared, and many have exhibited laser action. Many of the materials mentioned herein may prove to be useful in laser devices. However, it is too early to be certain which material will eventually be most suitable for a particular laser application. Research at RCA Laboratories, and of course in other laboratories, is being continued in an effort to help answer this question. One of the objectives of research at RCA Laboratories is to learn how to prepare a laser material which possesses absorption and emission levels that suit the application.

ACKNOWLEDGMENT

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DEVELOPMENT OF EFFECTIVE LASERS

The precise characteristics of a laser crystal are critical in achieving effective laser action. Considerable effort of the Advanced Development Group, SC&M, Somerville, is bent toward minimizing strain and impurities in grown laser crystals. Also described herein are the effects on laser performance of the polish, parallelism, and reflecting surfaces of the prepared laser rod and the geometry of the pumping system, as well as the construction of absolute and relative threshold measurement apparatus.

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The exact materials characteristics of a laser crystal are critical in achieving laser action. In a given material, even though theory may predict laser action, the practical laboratory work toward this end can be unsuccessful because of imperfections and impurities in the available crystals. Since the probability of achieving laser action at available and efficient power levels is frequently marginal, any loss of power due to scattering by the impurities and imperfection in the crystal can be disastrous.

A second deterrent to successful operation is the presence of strain in the crystal as a result of differential thermal contraction when the boule is cooled to room temperature from its freezing point. In laser crystals, this strain causes refraction of the laser beam, scattering, and broadening of the line width. These effects tend to increase threshold input power, overheat the material, and thus reduce the probability of continuous operation. Even more disastrous is the possibility that the laser rods will shatter when cooled to optimum operating temperature, which is usually in the neighborhood of 100 °K.

Cuprous and silver halides, another class of materials, show substantial promise as laser modulators, but have also presented difficulties as a result of strain. Because such modulators operate on a principle employing a birefringence mechanism, any built-in strain birefringence tends to obscure the detection of the desired effect.

CRYSTAL-GROWTH DEVELOPMENT

Of the laser materials presently under development, alkaline-earth fluorides present both the problems of scattering by oxygen impurities and of strain. Some other operational lasers such as ruby, the scheelites, and the glasses are oxide systems, and oxygen does not result in scattering. In spite of these problems, however, these fluorides show promise because they are excellent hosts for both divalent and trivalent rare-earth ions which, because of their narrow fluorescent line widths, are desirable laser "dopes." (Only a few non-rare-earth ions, such as trivalent chromium and tetravalent manganese in corundum, have exhibited laser action up to now.)

Of all the possible alkaline-earth fluorides, only calcium fluoride and barium fluoride presently show good crystal properties with respect to high purity, stability, and lattice perfection. As a result, emphasis in this work has been directed toward these two hosts. Fig. 1 shows typical calcium fluoride standard laser rods cut from grown crystals. (See front cover.)

PREPARATION OF SCATTER-FREE CRYSTALS

The starting materials for laser crystals may be obtained from random pieces cleaved from boules grown for spectrometer prisms. After surface encrustations are removed, the material is checked for scattering centers by means of a narrow beam of light from a high-intensity carbon-arc lamp directed through the crystal. If scattering centers are present, the passage of the beam through the crystal is visible to the eye and appears to have a bluish tinge, the *Tyndall effect*. At this stage, scattering is usually negligible. After the crystal is grown, however, the Tyndall scattering may be quite pronounced. Thus, it can be concluded that the scattering centers are introduced during the growth process. The crystal shown at the top of Fig. 2 exhibits pronounced Tyndall scattering; the crystal at the bottom does not.

When these materials were first doped with uranium, the crystals showing scatter also exhibited the characteristic yellow fluorescence of the uranyl ion UO_2. Consequently, it was concluded that the major scattering centers were random precipitates of calcium oxide in the lattice, and efforts were made to design an oxygen-free system. The possible sources of oxygen were assumed to be due to leaks and to absorbed gases in the system and impurities in the furnace atmosphere and in the dopant used. In an effort to reduce the possibility of contamination from the system, an RF crystal puller was constructed to minimize undesirable heating. In an RF system, only the crucible containing the material is directly heated; the crucible pedestal is heated by conduction only. The chamber walls surrounding the crucible seldom reach temperatures above 300°C, provided they are placed sufficiently far from the crucible. In order to eliminate any atmospheric oxygen, the chamber is carefully evacuated and purged with helium purified by a titanium-chip liquid-nitrogen-trap system.

After a series of experiments, it was found that oxygen was entering the system in various ways:

1) by outgassing of the carbon crucible
2) from contaminated helium supply
3) from surface oxides on the raw material
4) as a result of oxidation of the rare-earth fluoride dopants with time
5) as a result of low helium flow rates
6) by degassing of the titanium chips
7) through microcracks in the quartz envelope
8) as a result of devitrification of the quartz pedestal and heat shield

Improved processing techniques and more rigid controls on all parts of the system were adopted to solve the oxygen problem.

Because it was suspected that carbon was not a good material for crucible use, tests were made to determine its attack by the melt. When an undoped crystal was pulled, there seemed to be no variation in color or scattering between the starting material and the grown crystal; however, this was not the case for doped crystals. The following tests showed a definite relationship between coloring and the amount of time spent in a carbon crucible:

A uranium-doped crystal was pulled, a small portion of its top was removed, and the bottom material was then remelted and regrown. The top portion was again removed, and the process was repeated. No detectable scattering was found at any time, but the coloring deepened with successive processing until the crystal was completely opaque to visible light after the third growth. Spectra of the crystals showed both the 2.51- and 2.613-micron laser lines; the relative ratio of the second wavelength to the first increased with processing. The 2.613 line produces continuous laser action and is thought to be due to interaction of two neighboring uranium ions. In these crystals, the carbon could possibly have enhanced the combination of two or more uranium ions, and thereby increased the prominence of the resonance. The doping level in these tests was 0.05 percent, although the 2.613 line normally occurs only at much higher doping levels.

It was evident that darkening of the crystal could only impair laser action because it would absorb pumping light, increase the laser threshold, and heat up the lattice uselessly. Experiments with samarium in carbon crucibles showed that darkening becomes even more pronounced when small amounts of hydrogen are added to the atmosphere. It can be concluded, therefore, that the rare earths attack the carbon. Tests with molybdenum crucibles have evidenced little darkening except in the presence of hydrogen.

ANNEALING

In general, the more ionic the bonding, the more susceptible a crystal is to thermal shock and the greater the care required after growth to prevent shattering. Normally, a calcium fluoride crystal is first grown with great care to avoid cracking, and is then annealed for several days to stabilize the strain-free condition. This method was initially used at Somerville, but was found to have the following disadvantages:

1. The removal of strain by this method is achieved by the introduction of dislocations. Although the effect of the dislocations is not now known, it is probable that in the future they will impose a limitation on obtaining low laser threshold and narrow beam widths.
2. Oxide formed on the crystal surface during growth diffuses through the crystal during annealing, introducing scattering centers and changing the crystal from the divalent to the trivalent state.
3. The crystal may crack before or during annealing.

One solution to the problem is to grow the crystal, allow it to crack to relieve the strain, and then pick out the big pieces. A more acceptable solution is the prevention of strain. Strain is introduced by a number of processes, such as differential thermal contraction during cooling, non-planar solid-liquid interface during growth, and distortion of the lattice during growth by dopant and contaminant atoms. Lattice distortion can be minimized by careful control of contamination.

Little can be done to eliminate strain introduced by the dopes, but this strain can be tolerated if the other sources are controlled. Previous studies on uranium-doped calcium fluoride also showed that the severity of cracking increased with the carbon concentration. Each dope introduces more or less strain. Thus, samarium- and uranium-doped calcium fluoride can be grown with minor annealing

Fig. 1—Calcium fluoride standard laser rods (about actual size) cut from grown crystals.

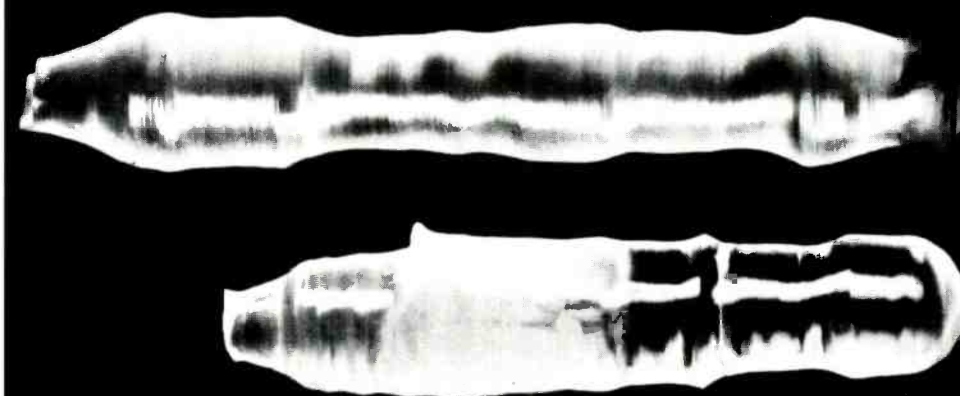

Fig. 2—Calcium fluoride crystals with (top) and without (bottom) scattering centers as shown by the Tyndall effect. (About actual size.)

Fig. 3—Strain birefringence in a laser output as shown under crossed polarizers.

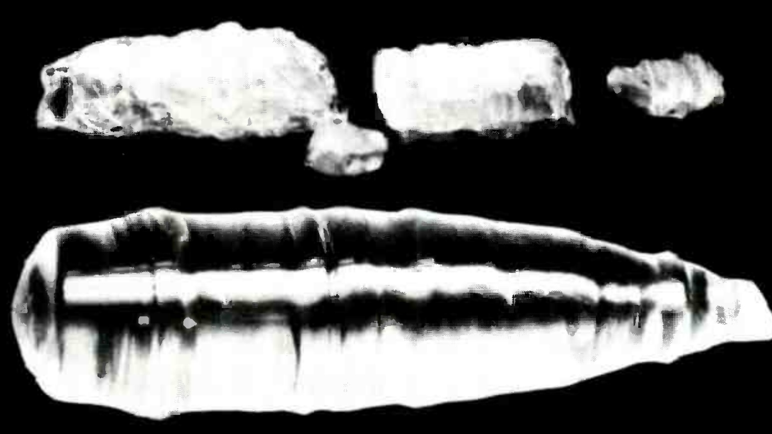

Fig. 4—Pieces (top) of crystal annealed by the original method, and a crystal (bottom) annealed by the improved procedure.

when an RF-powered puller is used, but holmium-doped crystals frequently crack during growth. Thulium-doped calcium fluoride is grown easily without cracking, but usually shatters on cutting. The strain in the host crystal seems to be a function not only of the dope, but also of the relative ratio of divalency to trivalency of the ion (the strain is generally higher with trivalent ions). This result appears reasonable since a trivalent state in the alkaline-earth fluorides requires an interstitial fluorine ion next to the trivalent ion; however, the assumption does not explain the ease of growth of trivalent uranium.

The maintenance of a planar solid-liquid interface is controlled by proper melt gradients, and is an old and familiar story in crystal growth.

In an effort to prevent differential contraction during cooling, an annealing susceptor was introduced into the growth chamber. The crystal was drawn up into the susceptor after growth, and the RF coil was also raised. The annealing tube was then heated to 850°C, and slowly programmed to room temperature over a sixteen-hour period. This method worked exceedingly well for uranium, samarium, thulium, and other dopes that could be pulled divalently.

A modified puller was then used in which the annealing susceptor was next to the growth susceptor. The RF coil was spaced nonuniformly to provide a uniform temperature over the growth region. The annealing-tube temperatures ranged from 1000°C at the bottom to 700°C at the top. Thus, the crystal never experienced a sudden temperature change during the cycle. During annealing, the temperature could be programmed down, or the crystal could be pulled slowly through the annealing susceptor, or a combination of both methods could be used. Fig. 3 shows the strain in an uncracked crystal annealed by the original method. Fig. 4 shows a crystal annealed by the improved method; the polariscope showed the crystal to be relatively free of strain. Also shown are pieces of a crystal annealed by the original method.

DIVALENCY AND TRIVALENCY

As previously mentioned, the rare earths may enter the lattice in doubly or triply ionized condition (and in some cases tetravalently). Although either state can exhibit laser action, the divalent state is more attractive with respect to thresholds as far as a cubic crystal is concerned because of its greater pumping efficiency. It appears that better calcium and barium fluoride crystals can be produced if divalency can be made to occur during growth. However, the crystal potentials of these lattices are smaller than those of the rare-earth fluorides, and trivalency is more likely. Forced divalency may be obtained by the use of a reducing atmosphere (H_2) and the introduction of metallic rare earths into the melt.

Some dopes such as thulium, which may enter in either state, have evinced the unusual segregation property of entering the crystal trivalently while being divalent in the melt. When the melt is quickly frozen, it remains divalent. Advantage has been taken of this property by pulling the crystal and, instead of breaking its contact with the molten region, slowly freezing the melt and pulling it out of the crucible as a single crystal; the entire melt is then annealed in the normal fashion. The resulting large section tends to crack, however, because the annealing is not adequate for the larger-diameter crystal. The difference between the upper portion and the lower is striking; the former is deep blue and the latter colorless. Attempts at rigid control of the ion state are presently being made.

LASER-ROD PREPARATION

The threshold may be simply defined as the point at which the stimulated emission is equal in magnitude to the spontaneous emission. Spontaneous emission is essentially independent of time, whereas the induced emission varies exponentially with the laser rod length. Because the actual rod length must be kept small to accommodate available pumping sources, it is difficult to achieve the required physical condition for threshold, except by causing the laser beam to be internally reflected many times.

To accomplish multiple reflection, the ends must be made flat, parallel, and scratch-free; any deviation causes a decrease of the effective length, and results in an increase of threshold. Thus, careful control of the polishing process is a necessity in obtaining good-quality lasers. For this reason, methods for extremely accurate preparation have been devised and set up, and are ready for operation.

Two different types of rods are being prepared at present. The first is the normal plane polished rod, which is used for devices that emit a plane wave having a very narrow angular spread. The second is a confocal type that emits a spherical wavefront. The confocal type is simpler and more quickly fabricated, and is being used because it produces lower thresholds.

REFLECTING SURFACE

To obtain gain in laser crystals, it is necessary for the radiant energy in the crystal to traverse the rod several hundred times. One method of obtaining gain is to place a totally reflecting surface at one end of the rod and 1-to-10-percent transmission surface at the other end. A large fraction of the energy is then contained in the rod, and a substantial gain in energy results. Both metallic and dielectric surface coatings are employed to obtain high reflectivity. The dielectric type is more desirable because of its lower losses as compared to metallic coatings. Dielectric coating may reduce the threshold and increase the output energy by an order of magnitude.

LASER CRYSTAL TESTING

Until now, laser threshold measurements have been difficult because of the lack of standards for comparing different laser materials or lasers made from different crystals of the same material. This unfortunate situation arose as a result of the extreme variability of the measurement system. Each measurement contains the following variables:

1) differing geometries of the laser rods,
2) various types and sizes of pumping sources,
3) different operating temperatures,
4) many types of reflectors.

This lack of standardization is caused by the desire to obtain a low threshold number. An attempt is being made to set up a standard relative test. In this test, only one type of pumping flash tube is used, 1000-joule xenon flash tube. The output is kept constant while light reaching the rod is varied by insertion of wavelength-independent screen filters. The operating temperature is maintained at 77 °K, and a standard laser geometry is used (i.e., 1-inch length, ¼-inch diameter). A disadvantage of such a system is the absence of absolute values. A test that would provide absolute values would require a wide variety of flash tubes and operating temperatures, and is not very practical. Another problem is a test for threshold for continuous operation. This problem is much more difficult and would require extremely good pumps, reflectors, and filtering and cooling systems.

In addition to thresholds, measurements are made on the strain, scattering, and fluorescence of the crystals.

MODULATOR CRYSTALS

The best prospect for achieving modulation of laser beams at and above the gigacycle range is with the electro-optic effect. This effect causes a retardation of the extraordinary ray with respect to the ordinary ray as a function of the electric field across a crystal.

Normal electro-optic effect modulators such as potassium dihydrogen phosphate (KDP) and ammonium dihydrogen phosphate (ADP) are severely limited



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by their narrow transparency range with wavelength, high dielectric constant, low angular aperture, and maximum modulation frequency. Crystals being developed in the RCA Advanced Materials activity far surpass these normal materials in all the above categories. These crystals, which incude CuCl, CuBr, CuI, CuCl:CuI, and AgI:CuI, are expected to achieve a successful gigacycle modulation of cw lasers. A comparative evaluation of cuprous chloride and potassium dihydrogen phosphate is given in Table I.

Some strain-free crystals of small volume have been made by the use of long annealing. Experiments are now under way to grow larger crystals without strain. Using the small-volume crystals, the Princeton Laboratories have shown modulation of continuous infrared radiation at 1 Gc, and modulation of a cw calcium-fluoride dysprosium-doped laser at 1 kc.

GROWTH OF SCHEELITE CRYSTALS

A puller has been constructed and is ready for growth of the scheelite family crystals: $CaWO_4$, $SrMoO_4$, $ZnWO_4$, $PbMoO_4$, $CdVO_4$, and others. It is hoped that incorporation of new techniques will permit growth of more perfect crystals than now available. Many of the potential lasers that should exist in this family have either been completely unsuccessful, or have not been operated continuously even though they theoretically could be. The major causes of these failures are the strain in the crystal and doping problems, which are probably not insuperable.

SEMICONDUCTOR INJECTION LASER

The rod-laser devices conventionally require very-high-energy lamp sources. The mechanical difficulties caused by the required optical pumping restrict the applicability of these lasers. In some cases, the devices are fragile. Optical pumping suffers from two serious disadvantages: the pumping efficiency is usually less than 1 percent, and heating resulting from the lamp spectrum which does not contribute to pumping increases the threshold and decreases the power output. As a result, the over-all efficiency of a conventional laser is approximately 10^{-3}. Moreover, the laser output must be modulated by external means to be useful for communications; this modulation may reduce the efficiency by an additional 50 percent.

The injection laser should not suffer from any of these disadvantages. The pumping efficiency should be substantially greater without the effect of extreme heating. In addition, the pumping source may be modulated to eliminate the need for an external modulator.

Laser action may be achieved in semiconductors such as GaAs and InAs, in which a direct band-to-band recombination takes place. The injection of minority carriers to affect a net recombination process is the equivalent of a temperature inversion. The semiconductor is fabricated into a cavity, and the usual surface coating is deposited to increase the laser path. The techniques are fairly straight-forward, and experiments are in progress to produce the most efficient laser geometries. To date, such devices have been successfully demonstrated using gallium arsenide and mixtures of gallium arsenide and gallium phosphide.

CONCLUSION: WORK ON OTHER NEW MATERIALS

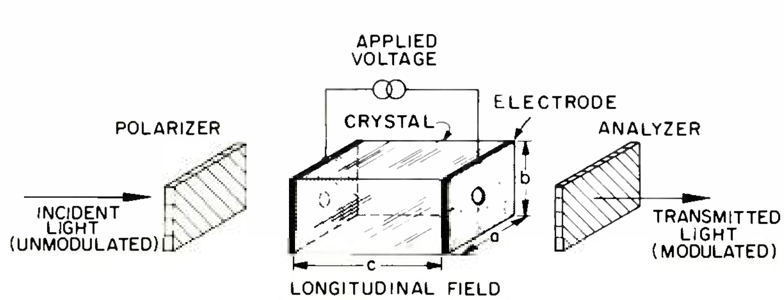
Exploratory investigations are underway to develop growth techniques for promising but previously unavailable crystals. Included in this group are special materials intended to provide higher radiation outputs and a wider range of wavelengths.

TABLE I—COMPARISON OF KDP AND CuCl

Parameter	*KDP*	*CuCl*
Half-wave retardation voltage, kv	7.5	6.2
Dielectric constant	15	8
Electro-optic coefficient r_{41} statvolt/cm	2.6×10^{-8}	18.4×10^{-8}
Angular aperture	acute problem	no problem
Transverse Pockels efficiency	none	yes
Dielectric loss tangent: highest report	0.05	0.2
lowest report	0.0005	0.0004
Limit of transparency, microns	0.5 - 1.3	15 - 20.5

General Observations:

1. The half-wave retardation voltage of CuCl is 6.2 kv, whereas that of KDP is 7.5 kv at DC for green light. Because the power requirement varies as the square of the half-wave voltage, CuCl has a large advantage in this respect.
2. Because CuCl has an electronic polarization and KDP an ionic polarization, the former suffers only a slight adverse effect at high frequencies. CuCl is useful above 30 Gc.
3. Because KDP has an optic axis, light must be collimated to 3 minutes or less; this requirement results in a severe alignment problem. On the other hand, CuCl is isotopic and can handle divergences to 30°.
4. KDP is extremely fragile and shatters easily with handling or temperature change. CuCl can withstand even rough handling, polishes easily, and is unaffected by sudden temperature changes.
5. The Q of CuCl is subject to control because the resistivity of the semiconductor material is sensitive to doping. KDP is an insulator; doping would affect its color but not its electrical properties.
6. The transmission of KDP stops at about 1.4 microns, whereas CuCl is transparent to 20 microns and beyond. CuCl can therefore be used with lasers operating in the infrared as well as the visible regions.
7. CuCl exhibits a transverse optical effect, and can therefore be used with long crystals to reduce the required voltage.
8. The dielectric constant of CuCl is half that of KDP; therefore the power losses and the perturbation effect are less in a cavity.



MODULATORS AND DEMODULATORS FOR LASER SYSTEMS

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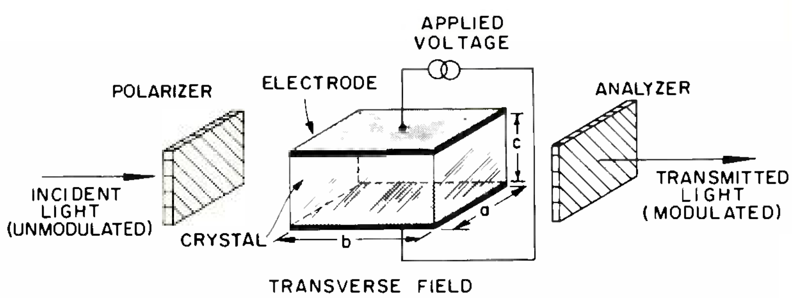


Fig. 1—Electro-optic modulators using longitudinal (top of page) and transverse electric fields.

As a step toward laser communications systems, RCA is investigating wideband laser modulators that utilize crystals exhibiting linear electro-optic effects. Work is also in progress on special microwave phototubes capable of demodulating light that has been modulated at gigacycle rates. (See also Engineering and Research Notes, Blattner, et al., this issue.)

As a result of the advent of lasers and the consequent effort to develop optical communication systems having bandwidths of thousands of megacycles, a need has arisen for wideband laser modulators and demodulators. The Microwave Applied Research Laboratory (MAR) of the Electron Tube Division is developing such modulators and solid-state demodulators under an Air Force program* directed by Dr. H. R. Lewis at RCA Laboratories. In addition, microwave phototube demodulators are being developed jointly by MAR and the Conversion Devices Laboratory of the Electron Tube Division under a Signal Corps program.† The work on these two programs is being carried out by H. C. Johnson, S. F. Miniter, L. M. Zappulla, and the authors in MAR, and Dr. J. E. Ruedy in the Conversion Devices Laboratory.

* Contract AF33(616)-8199, *Utilization of Coherent Light*, Electronic Technology Laboratory, ASD, Wright-Patterson AFB, Ohio.

† Contract DA36-039-sc-90846, *Quantum Detectors and Mixers.*

MICROWAVE MODULATORS

The wideband optical modulators being investigated at RCA use crystals that exhibit the linear electro-optic effect. In these crystals, the principal indices of refraction, n_α, are linear functions of the applied electric field, E. The changes in refractive indices are very small. Even for crystals exhibiting the largest known effect, $\Delta n_\alpha / E$ has a value on the order of only 10^{-11} meter/volt. Still, the induced birefringence may be large enough to produce phase differences of 180° between linearly polarized waves in crystals of practical size, and thus can be used to rotate the direction of polarization of a light wave.

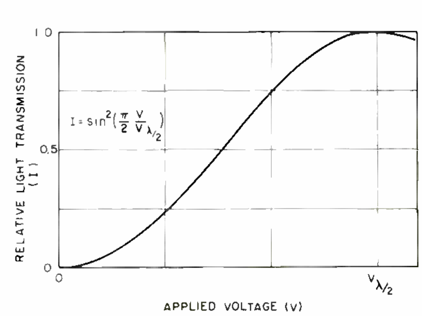
In the basic modulator, a crystal is aligned with its optic axis along the direction of the light beam and is placed between crossed polarizers (Fig. 1). When no electric field is applied, no light is transmitted through the analyzer. If an electric field is then applied across the crystal, the plane of polarization of the light passing through the crystal is rotated and light is transmitted through the analyzer, provided that certain relations between the direction of polarization of the light incident on the crystal, the directions of its crystallographic axes, and the direction of the applied electric field are satisfied. The intensity of the transmitted light as a function of the applied voltage varies as

$$\sin^2\left[\frac{\pi V}{2 V_{\lambda/2}}\right],$$

as shown in Fig. 2, where $V_{\lambda/2}$ is the voltage required for maximum transmission.

The modulators developed at RCA have used crystals of the dihydrogen phosphate family (ADP and KDP) and of cuprous chloride (CuCl). The dihydrogen phosphates were obtained from commercial suppliers; the CuCl crystals were grown by the Advanced Material activity of the Semiconductor and Materials Division at Somerville. The program for growing the CuCl crystals was initiated by J. L. Dailey and W. F.

Fig. 2—Light transmission as a function of applied voltage for electro-optic light modulator.



Schrotz of the System Engineering activity at Moorestown.

Because the dihydrogen phosphates are hexagonal crystals of class $\bar{4}2m$, they must generally be operated with the electric field applied in the same direction as the light beam[1], as shown in Fig. 1a. (Use of the dihydrogen phosphates in the transverse mode is difficult because the natural birefringence of the crystals generally exceeds the electrically induced birefringence by orders of magnitude.) Cuprous chloride, a cubic crystal, can be operated with the electric field applied at any angle to the light beam; for example, it can be operated with the field transverse to the light, as shown in Fig. 1b.[2,3]

The power, P_{in}, required to produce a voltage of amplitude V across the crystal can be calculated as follows:

$$P_{in} \approx \frac{V^2}{2}\omega\varepsilon'\frac{ab}{c}\tan\delta \qquad (1)$$

where: $\tan\delta$ is the loss tangent of the crystal material, ε' is the real part of the dielectric constant, and a, b, and c are the dimensions of the crystals as defined in Fig. 1. Values of $V_{\lambda/2}$, ε', and $\tan\delta$ for the various crystals used in RCA modulators are listed in Table I.

Fig. 3 shows a schematic diagram and a photograph of one of the KDP modulators. The KDP crystal is placed in a reentrant cavity operating in its lowest mode, and the light is transmitted through two small holes in the top and bottom of the cavity. The particular modulator shown operates at a center frequency of 1100 Mc, has a bandwidth of 10 Mc, and requires a driving power of 10 watts to produce a modulation depth of 10 percent. The depth of modulation can be increased to 20 percent if the light is passed through the crystal twice, as shown schematically in Fig. 4. The bandwidth of the cavity can, of course, be increased to any desired value by loading with lossy material. Because the driving power is approximately proportional to the bandwidth, a modulator of the type shown in Fig. 3 requires a driving power of about one watt per megacycle of bandwidth. Although this value is lower than has been reported by other workers, the power required for bandwidths of thousands of megacycles is still prohibitive.

There are several ways of reducing the power required to drive KDP modulators. For example, the light can be transmitted through the cavity many times, or traveling-wave modulators can use crystals many inches long. Although these possibilities are being actively investigated, a more promising approach to wideband low-power modulators may be the use of new materials like CuCl.

The power dissipated in a modulator is proportional to the real part of the dielectric constant ε' of the crystal (see Equation 1). For CuCl, ε' is only 8, compared to 20 for KDP. Furthermore, $V_{\lambda/2}$ for CuCl operated in the transverse mode can in principle be made arbitrarily small—at the expense of linear aperture—by reducing c/b. Also, the loss tangent of CuCl is lower than that of the dihydrogen phosphates.

Other advantages of CuCl include:

1) CuCl is transparent from 0.4 to 20.5 microns. None of the dihydrogen phosphates, by contrast, transmits wavelengths longer than 1.7 microns, and therefore they cannot be used to modulate infrared radiation beyond 1.7 microns.

2) The angular aperture of modulators using CuCl is orders of magnitude greater than the angular aperture of KDP or ADP modulators. Fig. 5 shows photographs of divergent light transmitted by a round ADP modulator and a wedge-shaped CuCl modulator. Only the very center of the pattern of the ADP modulator is usable, but the CuCl modulator produces nearly complete extinction and complete transmission for all light rays. Fig. 6 shows the calculated angular aperture of a modulator using a single CuCl crystal and of a "compensated" modulator using two crystals in series. The calculations are for crystals placed between crossed circular polarizers. The angular aperture of a KDP modulator is inversely proportional to the length of the crystal. Thus, although long crystals are attractive for power economy, their angular apertures are too small for many practical application.

3) Since the dihydrogen phosphates are generally used in the longitudinal mode, they must use either transparent electrodes (which are difficult to operate at high frequencies) or electrodes with holes (which operate only with fringing fields). As shown in Fig. 1, CuCl modulators using the transverse effect do not suffer from these limitations.

Fig. 3—Schematic diagrams and photograph of L-band re-entrant-cavity electro-optic light modulators.

TABLE I — Properties of Electro-optic Crystals

Property	$NH_4H_2PO_4$ *(ADP)*	KH_2PO_4 *(KDP)*	*CuCl*
$V_{\lambda/2}$, kilovolts*:			
longitudinal	9.6	7.5	6.2
transverse	—	—	7.2 c/b†
$\varepsilon'/\varepsilon_0$	12‡	20‡	8§
$\tan\delta$	0.005‡	0.0075‡	0.0015§

* low-frequency unclamped value, measured at $\lambda = 5461$ angstroms.
† E ⊥ (111) plane; dimensions b and c are defined in Fig. 1.
‡ Measured at x-band.
§ Measured at c-band.

The CuCl light modulators built at MAR have been operated in both longitudinal and transverse modes. Visible and infrared radiation (2.4 microns) were modulated. The highest modulation rate (1 Gc) was obtained in a reentrant cavity modulator of the type shown in Fig. 3c. The crystals used in these experiments were of good optical quality, but the largest reasonably unstrained samples had volumes of only a few cubic millimeters. Truly wideband low-power modulators will require unstrained crystals an order of magnitude larger. Work on preparing such crystals is now in progress at the RCA Laboratories and at the Semiconductor and Materials Division.

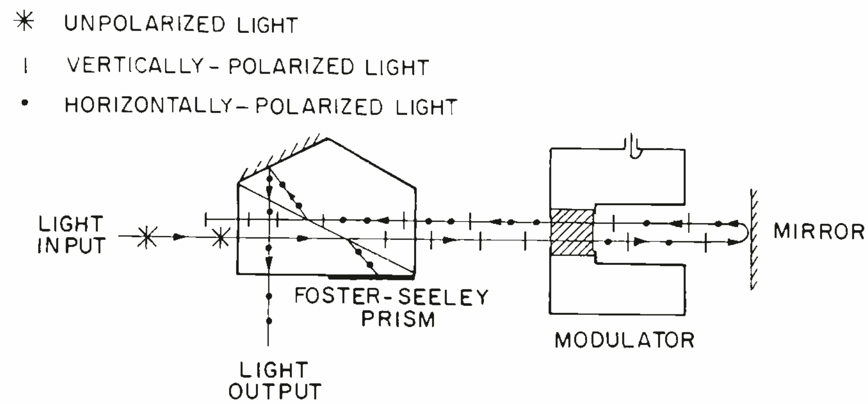


Fig. 4—Double-transit application of light modulator.

DEMODULATORS

The development of demodulators having gigacycle bandwidths is much further along than that of modulators. Special phototubes, photomultipliers, and semiconductor detectors are all capable of demodulating light that has been modulated at gigacycle rates. It seems likely that each one of these devices will be superior to the other two for a given system application.

One of the most important considerations in the choice of a light demodulator is the output signal-to-noise ratio S/N for a given number of incident photons. This power ratio can be written as

$$S/N = \frac{P_s}{P_{sh} + 4kTB} \qquad (2)$$

where: P_s = signal power delivered to the load; P_{sh} = shot noise power delivered to the load; k = Boltzman's constant; T = absolute temperature of load resistor; and B = receiver bandwidth.

For intensity-modulated light, and considering conventional photodiodes, microwave phototubes[4-6], and semiconductor photodetectors with no internal current multiplication, one can generally write:

$$P_s = \frac{m^2}{2} I_o^2 R_{eff} \qquad (3)$$

$$P_{sh} = 2\,e\,I_o B R_{eff} \qquad (4)$$

Where: m = modulation index of incident light; e = electronic charge; I_o = average photocurrent; and R_{eff} = effective output impedance as defined by Equation 3.

Normally, $4\,kTB \gg P_{sh}$, so the signal-to-noise ratio improves as R_{eff} is increased. As $R_{eff} \rightarrow \infty$, then $S/N \rightarrow (m^2 I_o/4eB)$, which is the best physically realizable signal-to-noise ratio. In conventional photodiodes and semiconductor detectors, the maximum value of R_{eff} is determined by the bandwidth requirement and the output capacitance. For bandwidths of the order of 1 Gc, the maximum value of R_{eff} is typically several hundred ohms. Microwave phototubes, on the other hand, have bandwidths of several thousand megacycles, and effective output impedance in excess of a megohm can be readily achieved.

The signal-to-noise ratio can be greatly improved by optical heterodyning[1-6], i.e. illuminating the photosensitive surface by a coherent local oscillator light beam having wavefronts parallel to the wavefronts of the light beam to be demodulated. In this case, both P_s and P_{sh} in Equation 2 are multiplied by a number proportional to the number of local oscillator photons incident per second on the demodulator. Thus, very good signal-to-noise ratios can be achieved, since P_{sh} can be increased to a value much greater than $4\,kTB$. Furthermore, this mode of operation provides automatic light filtering if a strong local oscillator is used.

However, optical heterodyning also has several important disadvantages. First, the frequency stability of both the transmitter and the optical local oscillator must be extremely good, so the complexity of both receiver and transmitter is much greater than with other systems. Secondly, if the light to be received is divergent, as it will be in all practical cases, it will be difficult to maintain a constant phase relationship between the signal and the local oscillator. If this phase relationship is not maintained, the signal-to-noise ratio will be degraded. Finally, since in optical heterodyning one must use double demodulation, the signal-to-noise ratio is deteriorated by approximately 6 db because energy is produced at various sideband frequencies that in most cases of practical interest cannot be recovered.

The signal-to-noise ratio of phototubes can also be improved by increasing the photocurrent by secondary emission multiplication before it passes through the output circuit. In this case[7]:

$$P_s = \frac{m^2}{2} I_o^2 D^{2\alpha} R_{eff} \qquad (5)$$

$$P_{sh} = 2\,e\,\frac{D^{\alpha}(D^{\alpha+1}-1)}{D-1} I_o B R_{eff} \qquad (6)$$

Where: I_o = current from the photocathode; α = number of stages of secondary multiplication; and D = secondary emission multiplication ratio per stage.

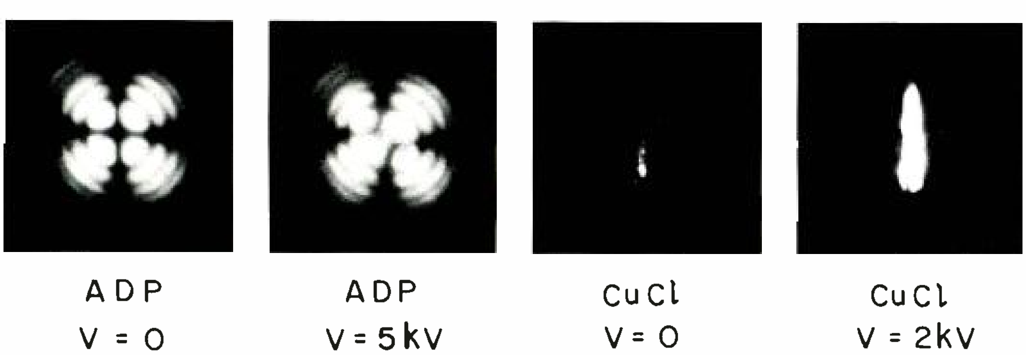
From Equations 5 and 6, the shot noise power is increased somewhat more than the signal power by the multiplication process. For typical values of D and α, the relative increase in shot noise power is only of the order of 30 percent. Thus (assuming negligible transit time effects, negligible dark current, and current gain sufficiently high that $P_{sh} \gg 4\,kTB$), the signal-to-noise ratio of a photomultiplier is 0.77 of the ideal value.

Microwave phototubes with helix-type RF output circuits have been built by the Microwave Lab and the Conversion Devices Laboratory. A brief description of these tubes is given in *Engineering and Research Notes*, this issue.[5]

Work is also in progress on microwave phototubes incorporating transmission-type secondary electron multipliers (Fig. 7). These multipliers should respond to modulation frequencies at least as high as X- or K_u-band.

One of the major disadvantages of all devices using photoemitters is that they cannot operate with wavelengths longer than about 1.3 microns. However, semiconductor photodetectors using small-bandgap material can operate out to wavelengths of several microns. For example, photodevices made from indium arsenide by Dr. J. T. Wallmark of the Special Devices activity of RCA Laboratories have been used in MAR to demodulate 2.4-micron radiation. It

Fig. 5—Photographs illustrating superior angular aperture of CuCl light modulator.



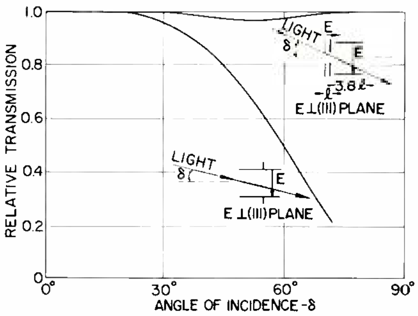


Fig. 6—Angular aperture of CuCl light modulators; relative light transmission as a function of angle incidence.

should be noted, however, that the ultraviolet response of semiconductor photodetectors is poor, while photoemitters have good response to ultraviolet radiation.

The most widely used high-frequency semiconductor photodetectors are p-n or p-i-n junction rectifier diodes mounted in encapsulations that permit illumination of the semiconductor material.[8] The diodes are biased in the back direction, and the only "dark current" is the saturation leakage current. When the diodes are illuminated with radiant flux having a frequency higher than the frequency of the fundamental absorption edge, photons are absorbed. Each absorbed photon produces an electron-hole pair. These pairs produce a current flow if they cross the junction where they are separated by the electric field across the junction. The quantum efficiency of a well-designed photodiode approaches unity; that is, each incident photon contributes one electron to the current flow.

Transit time for the charge carriers to traverse the junction limits the highest modulation frequency the diode can detect. Germanium photodiodes made by Dr. Wallmark were used in MAR to demodulate laser beats at 1600 Mc. Demodulation up to 11Gc with semiconductor photodiodes has been reported by Stanford University workers.

For video operation, the sensitivity of microwave phototubes and microwave photomultiplier tubes is much greater than the sensitivity of semiconductor photodiodes because of the small load impedance that must be used with the semiconductor diodes. On the other hand, if optical heterodyning is used and all sideband power can be utilized, the sensitivity of semiconductor devices can in principle approach that of an ideal receiver, since the quantum efficiency of semiconductors can approach unity. The maximum quantum efficiency of photoemitters is limited to about 0.2.

CONCLUSION

Wideband laser modulators and demodulators, like laser transmitters, are still in their initial stages of development. Because intensive work on all these devices is being pursued in many laboratories, it seems likely that practical wideband laser communication systems will become a reality within the next few years.

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DONALD J. BLATTNER received the BSEE in 1946 and the MA degree in Physics in 1949, both from Columbia University. For four years he taught physics and did solid-state research at Columbia. He joined RCA in 1953 as a member of the Microwave Tube Group at the Princeton Laboratories, and transferred to the Microwave Engineering activity of the Electron Tube Division in 1955. He has worked on low-noise and power traveling-wave tubes, backward-wave oscillators, ultra-high-speed computer devices and light modulators and demodulators. He is presently a member of the technical staff in the Microwave Applied Research Laboratory, Electron Tube Division, Princeton, N. J. He is a member of Sigma Xi and a Senior Member of the IEEE.

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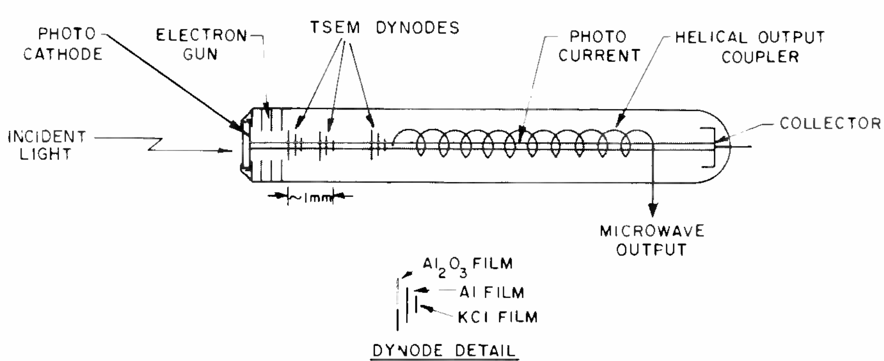
Fig. 7—Phototube using transmission-secondary-emission-multiplier (TSEM) amplification.



LASERS FOR RANGING APPLICATIONS

An engineering prototype of a laser ranging system has been built for the Signal Corps. As packaged, the prototype laser ranging system—sighting telescope, transmitter, and receiver optics—form a triangular about 5 inches on a side. In actual practice, this type of laser equipment could vary from a portable man-pack, or vehicle- or air-borne unit to use in conjunction with other sensors, for example, an acquisition radar. Potential civilian applications exist for such fields as surveying. Discussed herein are the basic considerations of laser ranging, and the prototype unit.

EDWARD KORNSTEIN graduated with a BA in physics from NYU in January 1951 and immediately joined the RCA optics group in the Optics, Sound and Special Engineering Section. He received a MS in Physics from Drexel Institute of Technology in 1954. During this period he was involved in the development and design of numerous optical systems for commercial and military applications. In 1957 he received a David Sarnoff Fellowship for study at Boston University. In 1960 he joined the Missile Electronics and Controls Division in Burlington, Massachusetts in a technical planning capacity and is presently Leader of the Electro-Optical Techniques Group, ACCD-Burlington. In addition to other programs, this group has been active in techniques leading to the development of laser systems and devices for the past two years. Mr. Kornstein is a member of the Optical Society of America, Society of Motion Picture and Television Engineers and the American Ordinance Association.

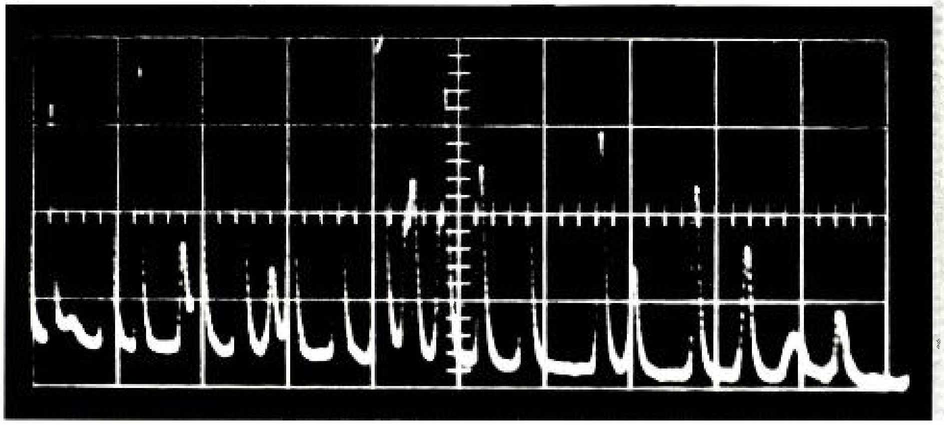


Fig. 1—Portion of a typical laser output pulse train. Sweep speed, 10 μsec/cm; peak power, 1 kw; material, neodymium in glass host; pump, GE FT524.

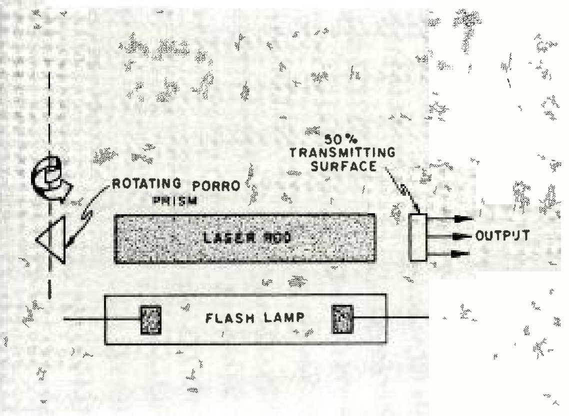


Fig. 2—High-peak-power laser transmitter.

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Since the demonstration of the ruby laser a little more than two years ago,[1] the potential application of these devices has fired many imaginations. At the DEP Aerospace Communications and Controls Division in Burlington, Massachusetts, a program to exploit the ranging capability of laser devices has led to the manufacture of engineering prototypes of a military laser ranging system for the U.S. Army Signal Corps. This very short time scale from the demonstration of a principle to the production of militarized equipment is indicative of the pace in this rapidly expanding field. (The principles and concepts of laser devices have been discussed in a previous issue[2] and the technical literature has provided many other excellent reviews and applications.[3])

While the ranging application is but one of the numerous possible uses of lasers, it seemed the most logical to exploit first, since the required auxiliary components and techniques have been fairly well developed for some time.

EARLY RANGING SYSTEMS

The key to the development of a laser ranging system is the means of controlling the transmitted pulse. Conventional pulsed laser systems produce a chain of pulses in a somewhat random fashion, as shown in Fig. 1.

Such an output can be utilized as a ranging system by performing a correlation between this outgoing pulse and the returning energy. Crude ranging systems were built using this principle. For example, the outgoing pulse could be displayed on one part of a dual beam cathode ray oscilloscope, the return beam picked up by a detector, such as a photomultiplier tube and displayed on the second trace. The tube face is photographed by a Polaroid Land camera; the two traces are then compared to determine the time elapsed between outgoing and returning traces. Such a system was proposed by Hughes Aircraft Company, California, as a ranging method.

An alternate approach proposed by TRG, New York, was to clip the outgoing signal so that only a few high-power pulses would be transmitted. Each time a pulse would be sent, a very rapid oscilloscope sweep would start and the return signal was placed on the Z, or intensity, axis of the cathode ray tube. If only one sweep was examined, it would be impossible to tell which signal was the true return and which was due to return of other pulses or noise. This sweep occurs many times in one flash of the exiting lamp, and since what is seen on the oscilloscope is the composite of several of these sweeps, only the true returns will appear at precisely the same place each time. The phosphor intensity at the true point will then be integrated over many sweeps and appear bright, while the intensity other places will correspond to the statistical average of other pulses, background light and tube noise, resulting in several low intensity points distributed at random across the trace.

Both of the above approaches seemed

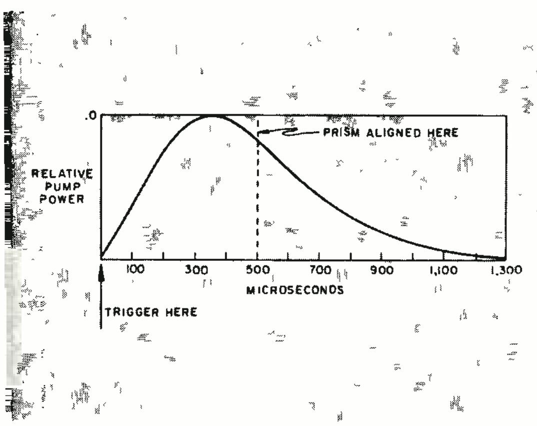


Fig. 3—Pump output and timing.

to be cumbersome and operative only under special conditions. It remained for Hellwarth[1] to predict the possibility of generating single giant pulses instead of the pulse trains. The theory of Hellwarth has been put into practice and is referred to as the technique of *Q-control*, or *Q-switching*.

GENERATION OF SINGLE PULSES

Referring to Wittke,[2] the amplification of an electromagnetic wave propagating through an emissive medium grows exponentially with distance until saturation effects make the system nonlinear. The medium acts as a travelling wave amplifier with a gain, G, given by:

$$G = \exp\,(\alpha L) \qquad (1)$$

Where: L is the length of the amplifier and α is the gain coefficient. The conventional equation for attenuation in a non-active medium is given by:

$$G = \exp\,(-\alpha L) \qquad (2)$$

Hence, the case for the active medium is sometimes referred to as the case of negative absorption. The coefficient α is related to the parameters of the laser material by:

$$\alpha = \frac{4\omega\mu^2\Delta N}{hc\Delta\nu} \qquad (3)$$

Where: ω is the circular frequency of the radiation, μ is the dipole moment associated with the transition between the ground and excited states, c is the velocity of light. ΔN is the number of atoms in the "excess" upper state population, $\Delta\nu = \Delta\omega/2\pi$ is the spectral width associated with the response of the atom to the field, and h is Planck's constant. Instead of using dipole moments μ, the Einstein coefficient of spontaneous emission A, which is related to the lifetime of the excited state, can be introduced[5]:

$$A = \frac{64\,\pi^4\nu^3\mu^2}{hc^3} \qquad (4)$$

The coefficient α can then be expressed as:

$$\alpha = \frac{A\Delta N c^2}{8\pi^3\nu^2\Delta\nu} \qquad (5)$$

The condition for oscillation of the laser is[6]:

$$Gr = \exp\,(\alpha l)\,r = \exp\left(\frac{A\Delta N c^2 l}{8\pi^3\nu^2\Delta\nu}\right) r \cong 1 \qquad (6)$$

Where: r is some effective reflectivity for the end plate. That is, a wave that has traveled through the crystal once and reflected once has changed in energy content by the factor $\exp\,(\alpha l)\,r$. Suppose that l, A, ν, and $\Delta\nu$ are constant; then, if r is small, ΔN can become large, but not so large that the condition for oscillation is reached. That is:

$$\exp\left(\frac{A\Delta N c^2 l}{8\pi^3\nu^2\Delta\nu}\right) r_{small} < 1 \qquad (7)$$

Now if suddenly r becomes large, ΔN still is large and:

$$\exp\left(\frac{A\Delta N c^2 l}{8\pi^3\nu^2\Delta\nu}\right) r_{large} >> 1 \qquad (8)$$

Here, the system is essentially overdriven and must relax by producing a burst of energy. By careful control of the pump sources and the timing, a substantial part of the energy can be made to be emitted in a single, short, high-peak-power pulse.

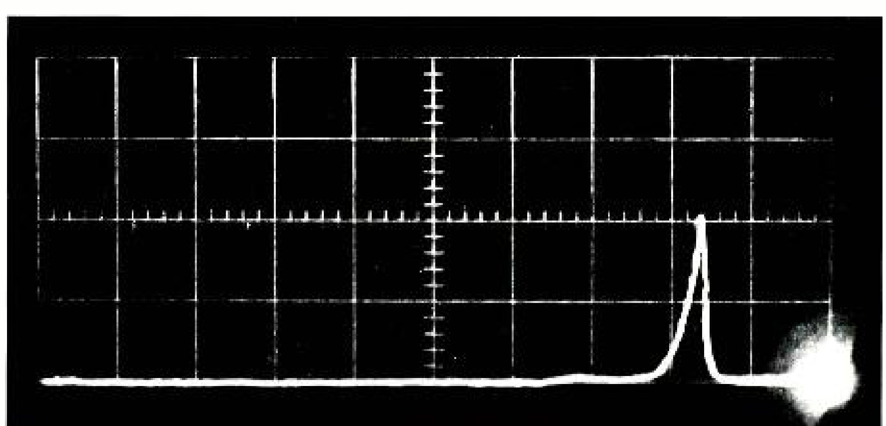
TECHNIQUE FOR CONTROLLING CAVITY Q

In the example given above, by changing the reflectivity of the end plates the laser output could be controlled. By examining Equation 6, other terms could be varied in order to achieve the same effect. For example, ΔN could suddenly be made large by pumping with a high-power, very-short-duration pump, such as a plasma pinch discharge. Also, $\Delta\nu$ could be made large by 1) application of an inhomogeneous magnetic field, 2) pumping the system to a large ΔN, but below threshold, and 3) then suddenly turning off the magnetic field.

The effective change in the reflectivity r, or the transmissivity $(1 - r)$, can be brought about in many different ways. One technique is to utilize the polarization of the stimulated emission in conjunction with an electro-optical modulator such as a Kerr cell or a Pockels effect cell. A very simple system using a rotating mirror or prism (demonstrated by R. C. Benson and M. R. Mirarchi at the Signal Corps in Ft. Monmouth, N. J.) can also be used. It is this latter approach which is presently being exploited at ACCD-Burlington.

The system is shown schematically in Fig. 2. A porro prism is used so that alignment and bearing tolerances are not critical, the prism being essentially

Fig. 4—Ranging laser pulse. Sweep speed, 0.1 μsec/cm (starting from left); power, 0.5 Mw/cm; material, pink ruby (3" x ¼"); prism speed 54,000 rpm; energy input, 160 μf at 2000 volts. (Because of low contrast on the original photo, the trace has been retouched.)



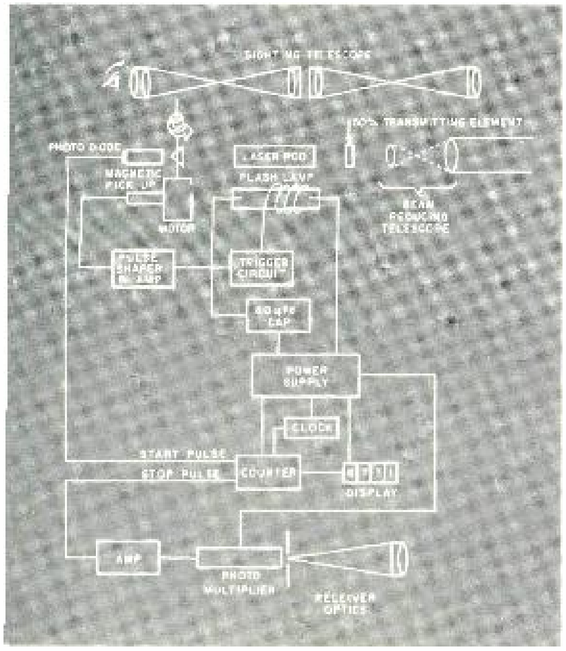


Fig. 5—Laser ranging system.

a retrodirective reflector in one dimension. The prism in the laboratory equipment is capable of rotating from 3000 rpm up to 60,000 rpm. A trigger circuit is provided so that the flash lamp is fired about 500 μsec before the prism face becomes parallel to the output face (Fig. 3). Two types of trigger circuits are used: One is a magnetic pickup in the motor housing and an iron slug buried in the motor shaft; the other uses autocollimating principle — that is, a parallel beam of light is reflected off the rotating member and detected by a 921A photomultiplier. The signal from the multiplier then triggers the flash lamp. Another autocollimator is also used to monitor the exact position of the mirror when the pulse is emitted.

Both magnetic and optical trigger units have been built. The magnetic system is used for field models because of its inherent ruggedness; the optical approach is used for laboratory experimentation because of its greater accuracy and versatility.

A sample pulse shape is shown in Fig. 4. (The pulse has been inked over for reproduction purposes.) In this case, the energy input to the flash lamp (a model XE1-3 line tube made by PEK Labs, Palo Alto, California) was 320 joules—2000 volts and 160 μf. The time scale shown is 10^{-7} seconds per centimeter division. The detector used was an RCA 925 phototube with a 100-ohm resistor and a 100-ohm cable termination feeding into a type 53/54L preamplifier on a Tektronix 545 oscilloscope. It is felt that the electronics are now limiting when these pulses are displayed. When the rotating prism speed will be increased beyond the present 60,000 rpm, a travelling-wave oscilloscope will have to be utilized. The photodetector was calibrated in a conventional manner using a standard lamp with a measured interference filter. The abscissa in Fig. 4 was then determined to have a scale of 0.5 megawatt per centimeter division.

RANGING SYSTEM CONSIDERATIONS

A typical ranging system is shown schematically in Fig. 5. A motor drives the Porro prism at a uniform rate. A magnetic pickup is mounted on the housing which generates the pulse to trigger the flash lamp. The 90° angle of the Porro prism has a very small chamfer so that when the laser beam is emitted through the 50-percent-transmitting end, a small amount of energy leaks backwards and provides a signal to a photodiode which generates a counter *start* pulse. The return pulse is received by optics boresighted with the transmitter. A *stop* pulse is generated and sent to the counter which has been counting the number of cycles elapsed on the clock. The counter is then interrogated and the result is displayed in decimal form. A beam reducing telescope can be used in front of the laser system to change the emitted beam (which may be as wide as 3 milliradians) to about 1 milliradian.

When packaged, the sighting telescope, the transmitter, and the receiver optics form a triangle about 5 inches on a side, and are all mechanically boresighted. Depending on the specific application and repetition rate, a typical laser ranging unit could be a portable man-pack unit, a vehicle-mounted unit, an airborne compact unit, or a unit operating in conjunction with some other sensor, such as an aquisition radar.

The laser rod itself is mounted inside an elliptical right cylinder at one focus with the flash lamp at the other. The elliptical right cylinder is essentially a very-high-speed optical system. Thus, the alignment of the flash lamp and laser rod has to be done with some care. Special techniques have been developed for the sagging of glass ellipses and the application of extremely durable, high-reflecting coatings. A typical cavity is shown in Fig. 6. The rotating reflector assembly uses a synchronous hysteresis motor directly coupled to the housing containing the reflector. Fig. 7 shows the relationship between the elliptical cavity and the rotating assembly.

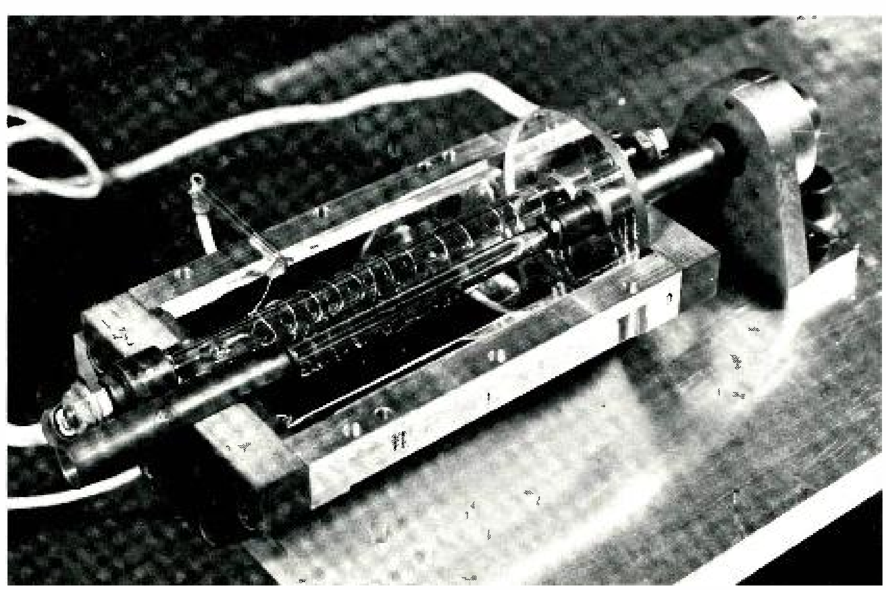
Various parametric studies have been undertaken to determine such variables as: the optimum spacing between spinning reflector and output face of the laser rod; the variation of output peak power as a function of reflector rotation speed; the effect of polishing or rough grinding the sides of the laser rod; the optimum coating for the transmitting end of the material; the power output as a function of inductance in the flash lamp circuit; the effect of temperature on power input and output; the effect of alignment errors on performance; and many others required to successfully design and manufacture a piece of operational, ruggedized military equipment.

RANGE EQUATION

The range equation developed for a laser ranging system is completely analogous to radar systems.

The power transmitted by a laser de-

Fig. 6—Optical coupling of lamp and three-inch laser rod in an elliptical cavity.



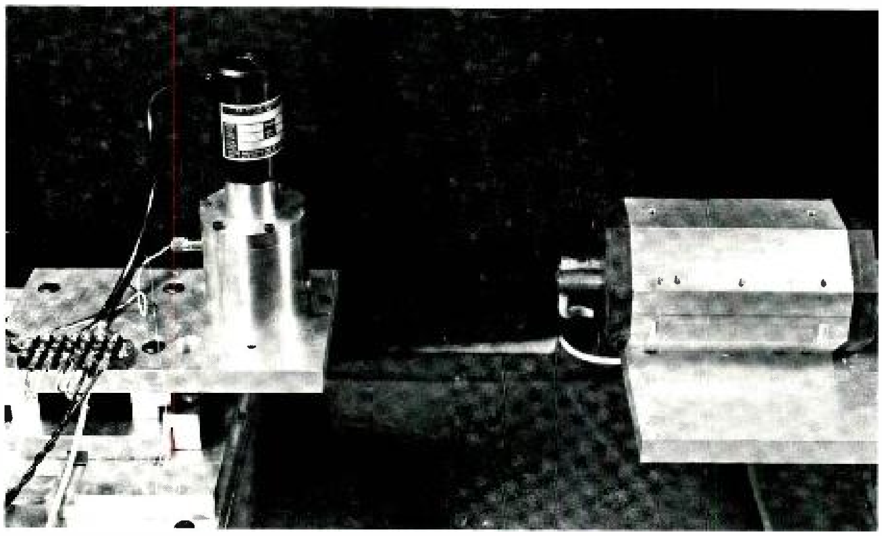


Fig. 7—Relationship of elliptical cavity of Fig. 6 and rotating assembly.

vice, I_T in watts/steradian, is:

$$I_T = \frac{I_L K_T}{\Omega_T} \qquad (9)$$

where: I_L = power output of the laser, watts; K_T = efficiency of the transmitting optics, and Ω_T = transmitter beamwidth, steradians.

The radiation I_o in watts incident on an object of cross-section area A_o at a range R with an atmospheric attenuation factor T is:

$$I_o = I_T \frac{A_o T}{R^2} \qquad (10)$$

If the object is assumed to be a Lambertian radiator normal to the radar line of sight, the reflected radiation I_R, in watts/steradian, is:

$$I_R = \frac{I_o \epsilon}{\pi} \qquad (11)$$

where: ϵ = reflectivity.

The power I, in watts, at the receiver is then:

$$I = I_R \Omega_R K_R T \qquad (12)$$

Where: $\Omega_R = A_R/R^2$, solid angle subtended by the receiver as seen from the target; A_R = receiver aperture area, and K_R = efficiency of receiving optics.

Combining Equations 9, 10, 11, and 12, we have, in watts:

$$I = \frac{I_L K_T K_R A_o A_R \epsilon T^2}{\pi R^4 \Omega_T} \qquad (13)$$

To get the signal-to-noise power ratio, the number of received photoelectrons, n must be calculated:

$$S{:}N = n = I\eta\bar{n}\tau \qquad (14)$$

Where: η = detector quantum efficiency, $\bar{n}$ = number of photons/watt-sec., and τ = pulse length. Substituting Equation 13 in 14:

$$S{:}N = \frac{I_L K_T K_R A_o A_R \epsilon \eta \bar{n} \tau T^2}{\pi R^4 \Omega_T} \qquad (15)$$

Background radiation from the sun and electronic noise must also be taken into account, but these have little effect unless the receiver happens to be pointed directly at the sun. A typical calculation for a ground ranging system shows why. For such a system let:

$K_T = K_R = 0.8$
A_o = 10 square meters
A_R = 50 square centimeters = 0.005 square meters
$\epsilon = 0.1$
$\eta = .02$
$\bar{n} = 3.5 \times 10^{18}$ photons/watt-second
$\tau = 5 \times 10^{-8}$ seconds
$\Omega_T = 10^{-6}$ steradians
$I_L = 10^6$ watts
$R = 10^4$ meters
$T = 0.7$

Then the signal to noise ratio back at the detector is 175, or 22.5 db.

One great advantage of the laser as a ranging device is the narrow beam which, when equal to the size of the target, as is possible in many ground applications, reduces Equation 15 to one containing R^2 instead of R^4. Then:

$$\Omega_T = \frac{A_o}{R^2} \qquad (16)$$

And, for the same parameters as the example above, a signal to noise ratio of 1750, or 32.5 db, is obtained.

To compute the effect of sunlight, consider that the power, I_o, in watts incident on the target from the laser is:

$$I_o = I_T \frac{A_o T}{R^2} \qquad (17)$$

Using values from the previous example, or, about 5.6×10^4 watts. The power from sunlight incident on the same target is about 0.144 watts/sq. meter/angstrom in a band centered about 0.7 microns, the spectral region of the ruby laser. Assuming use of a 10-angstrom-wide filter on the receiver, the sunlight incident on the 10-square-meter target which will be accepted by the receiver will be about 14.4 watts—compared to the laser, 5,600 watts.

In space applications, with no atmospheric attenuation or shimmer, the beamwidth can be reduced to $\Omega_T = 10^{-8}$ steradians with a range in the order of hundreds of kilometers for signal-to-noise ratios the order of 15 db.

CONCLUSION

In conclusion, one of the characteristics of a laser device has been exploited and has led to the manufacture of a militarized ranging system only two and a half years after the demonstration of the first ruby laser. The fact that such units will be in the hands of the field commands in this remarkably short time attests to the vitality and potential of this new field. If such a rate of progress is maintained, perhaps even the most imaginative applications thought of today will be obsolete tomorrow.

ACKNOWLEDGEMENT

The work briefly reported here is a result of efforts by many. Special references should be made to the contributions of W. Ahearn, B. R. Clay, T. Haddad, A. Lubin, T. Nolan, H. J. Okoomian, S. Sharenson, R. Smith and W. Strickland on the overall systems, mechanical and optical portions, and H. Halma, J. Klein, J. Lefebvre, K. Miller and J. Strom on the packaging and electronics.

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DAINIS KARLSONS received his BSEE in 1958 from Drexel Institute of Technology, and his MSEE in 1960 from the University of Pennsylvania. Mr. Karlsons' experience includes work on early solid-state microwave cavity masers. He has also participated in the design and evaluation of comb and meander-line traveling-wave microwave masers using ruby and rutile as the active materials. In addition, he developed a 5.8-Gc comb traveling-wave maser to be used in conjunction with a low-noise satellite tracking system. Mr. Karlsons was instrumental in the development of an early working model of a ruby laser. Shortly afterward he designed and operated a ruby laser which utilized an elliptical reflector to increase pumping efficiency. He also participated in a study evaluating the possibility of the laser's application to ranging and communications. Presently, he is developing a Q-controlled laser operating at liquid nitrogen temperature which should be capable of producing nanosecond pulses in the infra-red. Mr. Karlsons is a member of Tau Beta Pi and Phi Kappa Phi.

The authors, Donald J. Parker (l.) and Dainis Karlsons (r.), and a Q-controlled laser.

DONALD J. PARKER received a BS in Optics from the Institute of Optics, University of Rochester, N. Y. in June, 1950. He joined the Applied Research Section of RCA at that time, becoming supervisor of the Optics group in 1954 and Manager of Applied Physics in July 1957. His group is engaged in applied research and development in electrooptics, superconductivity, molecular resonance, and plasma physics. Current projects include microwave masers, laser tracking radar, laser fiber computer research, high field superconducting magnets, research in microwave generation by means of the Cerenkov effect and beam-plasma interactions. Mr. Parker's experience includes the development of many types of military and commercial optical systems such as optics for the first color TV cameras, color film broadcast equipment, large screen TV and color kinescope recording; airborne infrared search and tracking equipment, IR bomber defense systems, radar-IR seekers, television gunsights; high resolution scanning of color separations, TV aerial reconnaissance and radar recording. Mr. Parker is a member of the Optical Society of America, the Society of Photographic Engineers, the Society of Motion Picture and Television Engineers, and is a Senior Member of the IEEE.

LASER CHARACTERISTICS AND SOME POTENTIAL APPLICATIONS

While the laser can operate as an amplifier, it is more immediately attractive as a power generator capable of very high peak outputs, as well as attractive average power levels. Discussed herein are both favorable and unfavorable laser characteristics that importantly affect applications, and some current trends and needs in laser equipment development—where military applications seem to offer the most immediate promise.

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Both the laser and the maser are quantum mechanical amplifiers.[1,2] The laser (often called *optical maser*) operates at light frequencies, roughly a million times higher than the frequencies of the *microwave maser*. As a common physical principle, pumping power for either a laser or a maser is supplied to aggregates of ions located in a crystal structure, where this incoherent power is made available for coherent amplification of an electromagnetic signal propagating in the crystal structure.

But in actual operation and physical form, the similarity ends, for the vast difference in frequency between the laser and the maser causes the pumping and amplifier structures to differ in general appearance, and results in device characteristics that are very different.

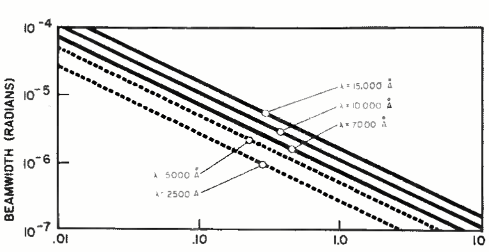

Fig. 1—Beamwidth vs. aperture diameter.

SIGNIFICANT LASER CHARACTERISTICS

The laser may be either an oscillator or an amplifier (just as the microwave maser), but so much energy is associated with individual photons supplied by the ions in the laser that its prime attribute is performance as a power generator capable of supplying very high peak powers and quite attractive average power. The microwave maser, on the other hand, is most attractive as an amplifier because the low energies associated with individual photons make low-noise amplification possible.

The ability to realize very small angular beamwidths in the optical region is the prime parameter of almost every application proposed for the laser, giving it the property of extremely efficient power transfer over long ranges. Emitter or receiver beamwidth as a function of wavelength and aperture diameter is shown in Fig. 1. The small wavelengths associated with lasers also means that small oscillating structures can serve as relatively high-gain antennas for radiating their energy, since even quite small structures are many wavelengths in diameter.

Laser systems suffer from very high noise temperatures relative to microwave systems, but the advantage of greatly decreased beamwidths more than obviates this deterrent. Other prime factors of interest, favorable and unfavorable, for the general application of lasers, are:

1. The very narrow spectral emission, which allows ultranarrowband filtering and, consequently, a high

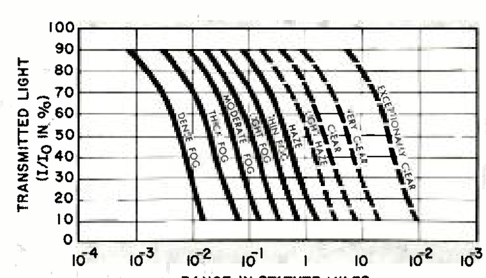

Fig. 2—Effect of scattering of light transmission at sea level.

degree of freedom from background noise sources.

2. Very difficult propagation characteristics for application through the earth's atmosphere. Mostly due to scattering, this effect is a problem on clear days and an impenetrable curtain on misty or foggy days. Figs. 2 and 3 illustrate the attenuation due to scattering from a collimated beam transmitted through the atmosphere in horizontal and vertical paths, respectively.
3. The very high frequency in the optical region, approximately 500 million megacycles (500 teracycles). This implies tremendous doppler frequency shifts and, hence, precise measurement of small range rate changes. Also, the tremendous bandwidth available in the optical frequency band is a great challenge to the communications engineer.
4. Rather poor efficiencies (1-percent prime power to transmitted power) with present laser transmitters. Until this situation changes it must be a basic systems consideration.

SIGNAL GRANULARITY

As mentioned previously, a high equivalent noise temperature is associated with the optical frequency band. The noise is due to the very high energy per photon ($h\nu$). For a given amount of transmitter power, there are approximately a million times fewer photons at the optical frequency than at RF. Thus, the signal is extremely granular. When received by a photon detector, the statistical fluctuations in the emission of the photons, and hence photoelectrons, corresponds to a physical noise-power source; the *shot noise* long observed in phototube use. Equated to an equivalent thermal noise (KTB), this noise source produces the equivalent noise temperature shown in Fig. 4 for a wide range of the electromagnetic spectrum. Equivalent noise power versus frequency is shown in Fig. 5, the noise power again being due to photon granularity.

DETECTORS

The noise temperature of the optical frequency region is so high, then, that low-thermal-noise amplifiers (even if there were any) do not offer improved performance over such conventional devices as the well-known multiplier phototube. The important characteristics that laser detectors must have are 1) *high quantum efficiency*, 2) *high frequency response* and *large bandwidth*, and 3) *good power-conversion characteristics*.

1.) *High quantum efficiency:* Since

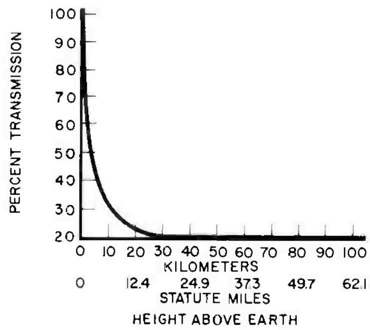

Fig. 3—Effect of scattering and altitude on light transmission (idealized curve for a clear day).

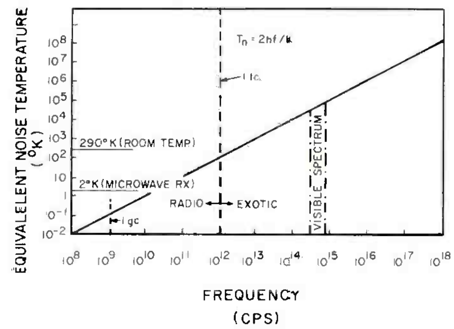

Fig. 4—Equivalent photon statistics noise temperature.

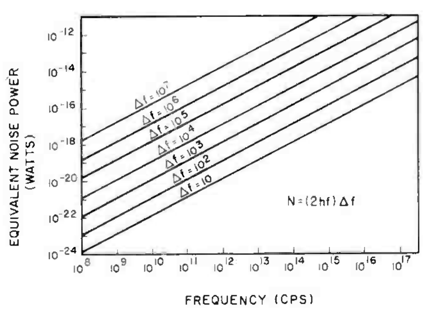

Fig. 5—Equivalent photon statistics noise power.

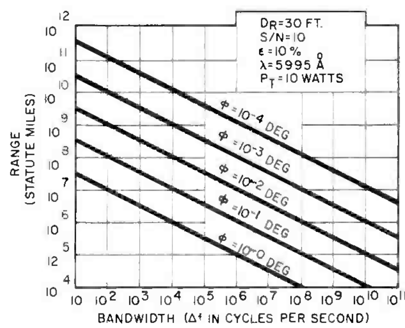

Fig. 6—Light range vs. bandwidth, varied beamwidth.

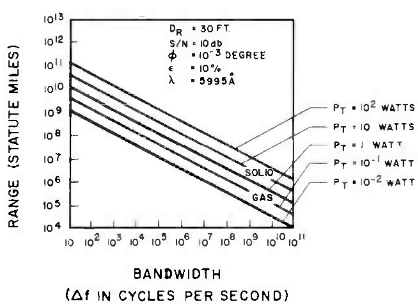

Fig. 7—Light range vs. bandwidth, varied transmitter power.

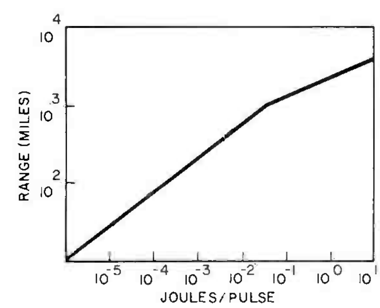

Fig. 8—Radar range vs. pulse energy.

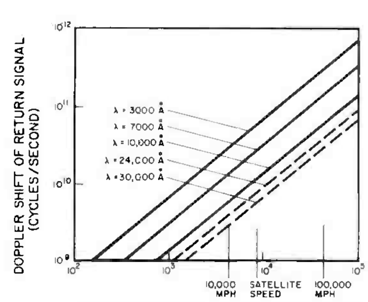

Fig. 9—Doppler shift of return radiation.

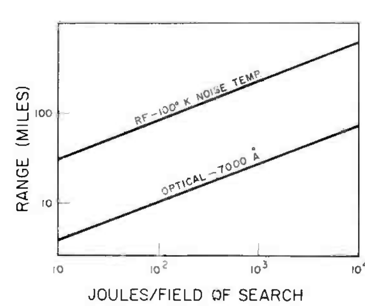

Fig. 10—Optical search power.

the main noise source is granularity due to the few photons per joule, the detector must convert as many of the incident photons as possible into photo-electrons. The multiplier phototube is capable of 40-percent quantum efficiency at its peak in the blue region of the visible spectrum. Beyond 1-micron wavelength, however, the quantum efficiency becomes so low that other detectors must be considered; such as semiconductor photodiodes, which exhibit response up to 10-micron wavelengths.

2.) *High frequency response* and *large bandwidth:* For heterodyned radar and communications applications the IF center frequency may be in the gigacycle range, requiring hundreds of megacycles of bandwidth. This type of operation stretches the conventional photomultiplier tube beyond its limits. Considerable work is being done on the combination of photoemissive cathodes with wide-band traveling-wave amplification techniques. Semiconductor detectors are also of interest because of their potentially high frequency response.

3.) *Good power conversion characteristics:* the output signal generated in the external circuit by the photodetector must be of sufficient magnitude to successfully compete with the thermal noise in the following amplifier stages. In the multiplier phototube, this is accomplished by secondary emission multiplication, an amplification factor of a million or more being achieved with negligible additional noise introduction. Amplification in the photo-emissive traveling-wave tube is in the range of hundreds—which is not sufficient for most low-level (long-range) applications. In the semiconductor photodiode, the situation is even worse in that there is no significant signal power amplification. However, in heterodyning, wherein the return signal is mixed with laser output, the signal at the difference frequency is proportional to the signal amplitude and also to the local-oscillator amplitude. Hence, by simply making the local oscillator much stronger than the signal, it seems possible to achieve the required power amplification. This raises some practical uncertainties as to requirements for amplitude and frequency stability of the local oscillator. However, this type of operation is extremely attractive since it would allow accurate, wide-band as width, infrared-sensitive systems.

COMMUNICATIONS APPLICATIONS

Fig. 1 showed that very narrow beamwidths are achievable with moderate aperture sizes. This gives the power transfer efficiency which is the attraction of the laser, and results in the ranges illustrated in Fig. 6. As Fig. 7 shows even for very low transmitter powers, the low-bandwidth systems can achieve phenomenal ranges. Figs. 6 and 7 have been determined from vacuum path calculations and, hence, relate to outer space applications.

Since the communication ranges shown in Figs. 6 and 7 are realized primarily from very narrow transmitted beamwidths, actual systems applications will require such precise tracking and pointing accuracy that the system trade-off of pointing requirements versus range may be somewhat disappointing. Undoubtedly, however, losses will be offset appreciably by increased power output and laser efficiency.

RADAR APPLICATIONS

The comparative range of optical *tracking* radar is much greater than RF. Fig. 8 shows the theoretically attainable radar range versus pulse energy. The break in the curve occurs at the point where the target size is the same as the beamwidth. The available range, computed for free-space conditions, is enormous.

The very narrow beamwidth, of course, also allows accurate measurement of angular position. Precision tracking pedestals and angle measurement equipment will be needed to realize the full potential of optical radar.

Optical radar offers the theoretical possibility of obtaining *very precise* range-rate information. Fig. 9 indicates the doppler shift for various wavelengths and target velocities; the large value of the doppler shift indicates the precise measurement possibilities. Many difficult transmitter and receiver problems must be solved in order to utilize this doppler shift. The return signal must be optically mixed with either the transmitted frequency or a stable local oscillator at a slightly different optical frequency. There is need for continued research and development in: single-mode CW laser pulse-amplifiers; wide-bandwidth high-sensitivity detectors; frequency translators; and tunable, stable local oscillators to realize the promise of optical doppler radar.

There has been considerable discussion about the prospect for optical *search* radars. Since the attractiveness of optical radars is attributable to the very narrow beamwidth, they are not competitive with RF for searching equivalent solid angles. Fig. 10 illustrates the relative powers required.

However, there will undoubtedly be many applications where RF search and acquisition is used to augment optical radars in order to gain the superior range rate and angular-position information. Also, optical radars will be used for small solid-angle search functions, such as accurate satellite orbital measurements where prediction information is available.

RCA ACHIEVEMENTS IN LASERS

The range of laser activity at RCA reflects the expected broad application of this device. It forms an important portion of the research work of the RCA Laboratories, the Electron Tube Division Advanced Development Laboratory, the Semiconductor and Materials Division, and many of the major operating divisions of Defense Electronic Products. Highlights of this work include:

1) *Material research resulting in the lowest threshold solid-state laser material yet discovered.* RCA Laboratories, Princeton, N. J. Air Force Contract.
2) *One of the first laser equipment contracts — optical ranging units.* DEP Aerospace Communication and Controls Division, Burlington, Massachusetts. U. S. Army Contract (see paper by Kornstein, *this issue*).
3) *First laser computer contract—the use of active optical fibers to do logic at nanosecond rates.* DEP Applied Research, Camden, N. J. Air Force sponsored.
4) *Optical tracking radar experimental program.* Joint effort by DEP Applied Research, Camden, and DEP Missile and Surface Radar Division, Moorestown, N. J.
5) *Several classified laser R & D contracts in DEP.*

Technical competition in the research and development of lasers is truly incredible. Since the first experimental results were announced in 1960 there have been more than 150 papers reporting experimental work in national technical journals. Most of this work has been paralleled many times by other workers in the field. It is to the credit of RCA's thoroughness in research in the 1950 era that the Corporation has been able to compete effectively for research and equipment recognition. The big payoff will undoubtedly be in major military systems; RCA's success in this area will depend on the creativity and skills applied in research and development over the next 2 to 3 years.

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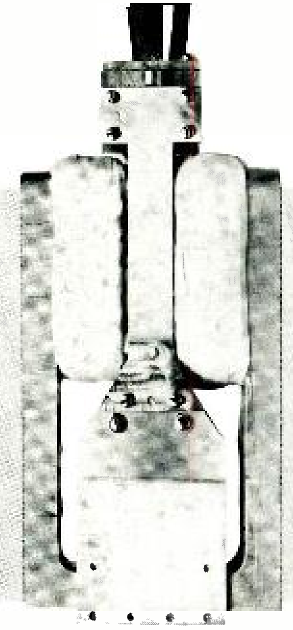

Fig. 1—Superconducting magnet clamped to a 2.5-inch-long maser developed at DEP Applied Research.

MASERS IN SYSTEM APPLICATIONS

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Maser development has reached the point where the masers are applicable to a wide variety of electronic systems. Further advancements will depend greatly on how cleverly the system designer can utilize the maser—especially in analyses of the sources contributing to system noise, and designing to minimize the noise contribution of the system components.

The maser is the most sensitive microwave amplifier available today. It is capable of amplifying signals that are close to the limits established by quantum energy ($h\nu$). The traveling wave version is extremely stable in both amplitude and phase. Since they now require negligible pump energy, maser amplifiers have reached the point of practicality for many military and commercial applications. The complete microwave and millimeter wave portion of the spectrum can be covered, since maser action does not depend on distributive circuit elements or electron flow. (Previous literature[1] describes the physics of maser operation.)

Until recently, the size of the maser package has subtracted from its desirability for many system applications. The magnets required to supply the electromagnetic fields weighed several hundred pounds. To get the maser temperature down to an operational level (liquid-nitrogen or liquid-helium temperatures), large dewar flasks or liquid refrigerators were necessary. Now recent advances in the supporting fields of superconductivity have provided a magnet weighing only a few pounds.[2] Higher-temperature operation with new maser materials and the development of closed-cycle refrigerators have greatly reduced the bulk of the cooling equipment. Maser packages small enough for satellite-borne systems are now approaching realization.

Maser noise temperatures are extremely low. This advantage cannot be fully realized, however, unless the effective noise temperature of the input is also low. It becomes important to consider the sources contributing noise to the maser input, such as system components and environmental sources. Some of these sources are briefly examined in this paper.

TABLE I—Noise Temperatures of Typical Losses at 300°K

Loss, db	*Gain (fractional)*	*Radiated Temperature °K*
0.05	0.99	3
0.50	0.89	33
0.70	0.85	45
1.00	0.80	60

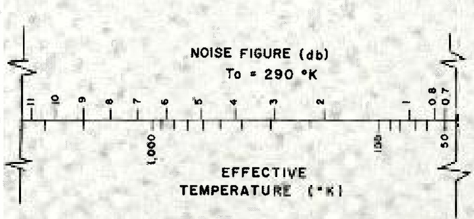


Fig. 2—Nomogram of noise temperature vs. effective temperature.

RECENT MASER DEVELOPMENTS

Until a few years ago the maser, in a cavity form, was a laboratory phenomenon with little application outside the field of radio astronomy. Later, workers at Bell Laboratories developed a traveling-wave version which placed the maser in the forefront of low-noise-receiver technology. This amplifier had a 15°K noise temperature and maintenance gain stabilities of 0.1 db for hours of operation. However, even at this point, maser amplifiers were clumsy. The amplifying structures themselves were 6 inches long, and they required heavy magnets and large refrigeration systems. Then, engineers at RCA-DEP Applied Research, Camden, N. J., working with a new material developed by the RCA laboratories (*rutile*[3]) and a novel slow-wave circuit, constructed an extremely small compact maser. It is only 2.5 inches long and requires less than 30 mw of pump power, compared with 100 mw for the older traveling-wave masers. This reduced pump power requirement combined with a reduction in magnetic field requirements has permitted the design of a compact superconducting magnet. Fig. 1 shows this superconducting magnet clamped to the maser.

Another area of investigation at DEP Applied Research is the development of millimeter-wave masers (operation at 35 Gc). Since the energy separation of rutile allows amplification in this region, workers are investigating the coupling of this maser material to a slow-wave circuit[5], such as the ladder structure. The high dielectric constant of rutile combined with the high ladder line slowly is expected to yield gains of 30 db in a structure length of less than 1 inch. A small superconducting magnet has been designed for this structure. The complete maser-magnet assembly will weigh less than 5 pounds. It will operate over a 10-percent tunable bandwidth.

A significant maser development this year took place at the RCA laboratories where scientists used a properly doped rutile crystal to achieve amplification at liquid-neon temperatures. Since the principle of low-noise amplification is not basically dependent on the amplifier bath temperature, this higher temperature operation can lead to the development of a maser refrigeration system that will weigh less than 30 pounds.

With these advances, the application of masers to communications, radiometry, and radar is inevitable. Already the maser is being employed in such applications as Telstar and Echo. The application of masers to deep space communications is another obvious step. The latest Venus probe experiment will utilize masers in the ground stations.

Masers need not be limited even to these applications. With a major design effort, airborne or spaceborne maser receivers may be developed. A maser operating above the atmosphere at 90 Gc and working in conjunction with a backward-wave oscillator delivering 50 watts of cw power would significantly improve a space-to-space communication system. A small antenna with high gain and directivity would then be possible.

MASER NOISE TEMPERATURE

An ideal receiver or amplifier does not add noise to that delivered from the signal source; thus, the output noise of an ideal amplifier is the input noise multiplied by the amplification factor. The *noise factor*, *F*, of an amplifier is defined as the ratio of the actual output noise

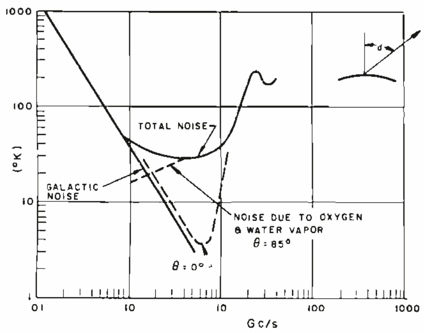


Fig. 3—Sky noise temperoture due to atmospheric absarption and galactic sources.

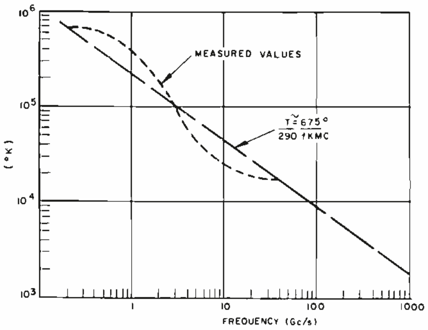


Fig. 4—Noise temperature of the sun.

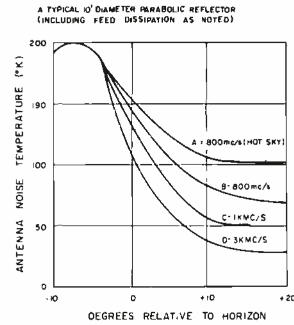


Fig. 5—Antenna main lobe elevation vs. antenna noise temperature.

power available to that which would be available if only the thermal noise of the source were amplified. The *noise figure* is this power ratio expressed in decibels. The input noise temperature of a network, in terms of its noise factor, can be represented as:

$$T_{eff} = T_o(F\text{-}1) \qquad (1)$$

where: T_o is a reference temperature. Since a majority of sources are at or near room temperature, T_o is assumed to be 290°K.

For n cascaded networks, the noise factor of the cascade is:

$$F_{1 \ldots n} = F_1 + \frac{F_2\text{-}1}{G_1} + \frac{F_3\text{-}1}{G_1G_2} \ldots + \frac{F_n\text{-}1}{G_1G_2 \ldots G_{n-1}} \qquad (2)$$

where G_n is the gain of the individual networks. This expression is simplified by introducing the effective input temperature:

$$T_{1 \ldots n} = T_1 + \frac{T_2}{G_1} + \frac{T_3}{G_1G_2} \ldots + \frac{T_n}{G_1G_2 \ldots G_{n-1}} \qquad (3)$$

The nomogram in Fig. 2 shows the relationship between noise figure and noise temperature. Components such as antenna feed lines, rotary joints, and waveguide runs have losses that contribute to the effective noise temperature of the system. The noise factor for networks with a power loss can be represented as:

$$F = 1 + \frac{(1\text{-}G)}{G}\left(\frac{T}{T_o}\right) \qquad (4)$$

Table I shows some of the noise temperatures of typical losses occurring at 300°K ambient. The noise temperature of a traveling-wave maser is:

$$T_a = \frac{G\text{-}1}{G}\left[\frac{\alpha_m}{\alpha_m\text{-}\alpha_o} T_s + \frac{\alpha_o}{\alpha_m\text{-}\alpha_o} T_o\right] \qquad (5)$$

Where: T_s is the maser negative spin temperature; T_o is the maser bath temperature; α_o is the attenuation constant describing the total loss experienced by the forward wave, including ohmic losses of the slow-wave circuit, reverse isolator losses, etc.; and α_m is the negative attenuation coefficient of the maser spin system.

For a good traveling-wave maser with small ohmic losses, and with power gain $G >> 1$, Equation 5 reduces to:

$$T_a = T_s + \left(\frac{L_o}{G}\right) T_o \qquad (6)$$

where: L_o is the forward loss associated with the maser's input and output, as well as the forward ferrite losses.

Presently measured maser noise temperatures agree quite well with theory. Masers are, in fact, the *lowest noise amplifiers known*. Noise temperatures of 15°K to 19°K have been measured for traveling-wave masers at RCA, Bell Labs, A.I.L., and Hughes.

ANTENNA NOISE TEMPERATURE

The noise power received by an antenna from its surroundings is defined as the integral of the product of antenna gain and the noise power radiated to the antenna from its surroundings. The antenna noise temperature T_a is related to the antenna noise power P_a as:

$$T_a = P_a/kB$$

where: k is Boltzmann's constant and B is bandwidth.

This noise temperature is the result of such sources as galactic and extra galactic noise, radio star noise, and noise resulting from atmospheric absorption produced by water vapor and oxygen. Within the microwave region, certain "windows" occur, wherein low antenna noise temperatures are possible. Fig. 3 shows a curve for one such window between 2 and 10 Gc.

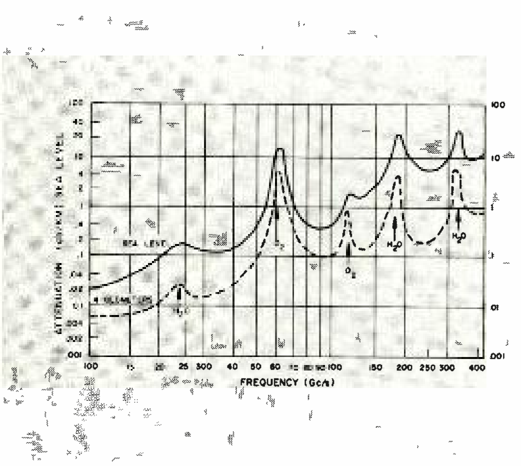


Fig. 6—Oxygen and water-vapor absorption in the atmosphere.

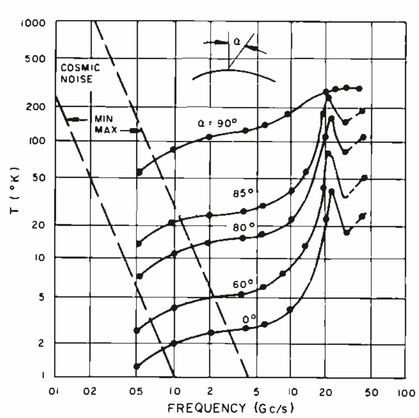


Fig. 7—Antenna temperature vs. antenna angle.

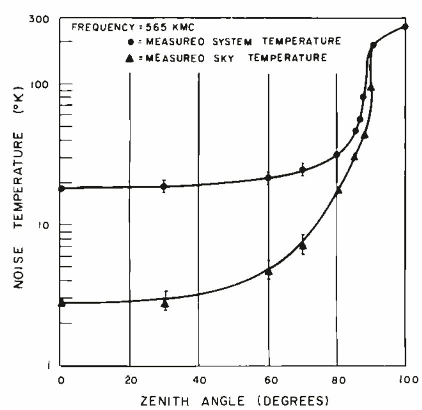


Fig. 8—Measured antenna temperature vs. zenith angle.

Such factors as back lobes, side lobes, and spill-over also contribute to antenna noise temperature. In some orientations, the sun may fall within the side lobes and back lobes, thereby raising the noise temperature considerably. Fig. 4 shows the relationship of the noise temperature of the sun as a function of frequency. Fig. 5 shows the effect on antenna noise temperature of moving the antenna main lobe away from the horizon.

ENVIRONMENTAL NOISE SOURCES

A number of phenomena within the atmosphere contribute noise to the system. Among these are aurora, the ionosphere, ozone, atmospheric oxygen, and water vapor. Fig. 6 shows the attenuation in the atmosphere due to the presence of oxygen and water vapor, as determined by Tolbert and Straiton at the University of Texas. Investigators at Bell Laboratories, R. W. DeGrasse, *et al*, have made measurements of the system noise temperature of a low-noise horn antenna operating into a traveling-wave maser. The results are shown in Figs. 7, 8, and 9. Note from these curves that, as the angle of the antena approaches the horizon, the noise temperature approaches 290°, or 3 db.

Galactic noise is the general background of radio waves existing in the aerospace environment and having the appearance of random noise. This is in addition to the radiation from discrete radio stars. Fig. 10 shows the intensity of galactic noise radiation, both in the direction of and perpendicular to the galactic plane.

MASER COOLING SYSTEMS

Amplification or oscillation is attained with a maser by the exchange of energy between externally applied electromagnetic fields and the bound electrons in the maser material. Within the crystalline maser structure, electron spin orientation is confined to a series of very narrow bands or lines corresponding to specific energy levels. At room temperature, ion distribution is approximately equal at the various energy levels. The differences between these fine-structure energy levels correspond to microwave resonant frequencies by the well-known relation, $E_i - E_j = hf_{ij}$, where h is Planck's constant.

To utilize the maser for amplification, it is first necessary to cool the material to a very low temperature. The electron spin orientation then approaches an equilibrium condition with greater ion population corresponding to the lower energy levels. A pump frequency is applied to attain an effective *negative temperature*, wherein more ions are forced into a higher energy level, making these levels more heavily populated than the lower levels. If this inversion involves three successive levels, then two of the levels are available to amplify a very weak signal field of corresponding microwave frequency. While this discussion is obviously oversimplified, it shows the need for extreme cooling of the maser material.

In early system applications of masers, the required cooling was obtained through the use of open-cycle dewars, by simply placing liquid helium in vacuum dewars and operating the masers for 12 to 18 hours, depending on the liquid-helium reservoir. This technique is feasible and still represents the most economical approach for certain applications. However, for most applications much more extended operation is involved, and either a more efficient cooling system is required, or maser operation must continue at higher temperatures, or both.

To meet the requirements of maser systems, a number of manufacturers have produced reliable closed-cycle refrigerators capable of thousands of hours of operation under severe operational environments. Some typical operating characteristics are 1,000 to 2,000 hours of continuous operation at 2.5°K. Such units are capable of supplying one watt of cooling capacity, and weigh in the order of 75 to 130 pounds. Additional supporting equipment such as motors, safety equipment, and temperature sensing equipment, generally located remotely from the cryostat, bring the total weight of the cooling system to some 200 pounds.

Fig. 11 shows a schematic flow diagram for a helium refrigerator, and Fig. 12 shows a complete helium refrigeration system. Fig. 13 shows the cryostat and helium refrigerator portion and the opening for the maser assembly. These are all products of Air Products Co. The major designers of liquefiers are Air Products, Garrett Corporation, and Arthur D. Little; all of these are working toward the development of lighter and more efficient helium systems.

FUTURE DEVELOPMENT

Maser development has reached the point where the masers are applicable to a wide variety of electronic systems. Recent experiments with laser (optical maser) pumping indicate that masers will some day be capable of operating at room temperature while retaining all their low noise properties.

With the advances in maser materials,

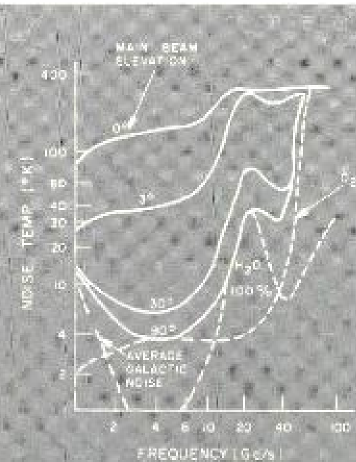
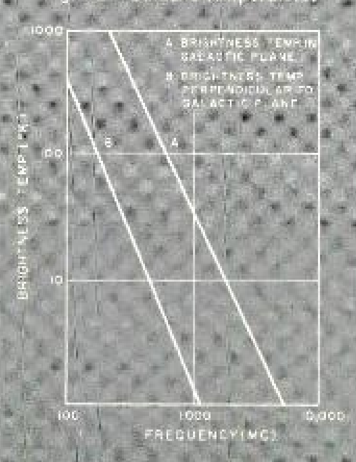


Fig. 9—Effect of main beam elevation on noise temperature.

Fig. 10—Galactic temperature.



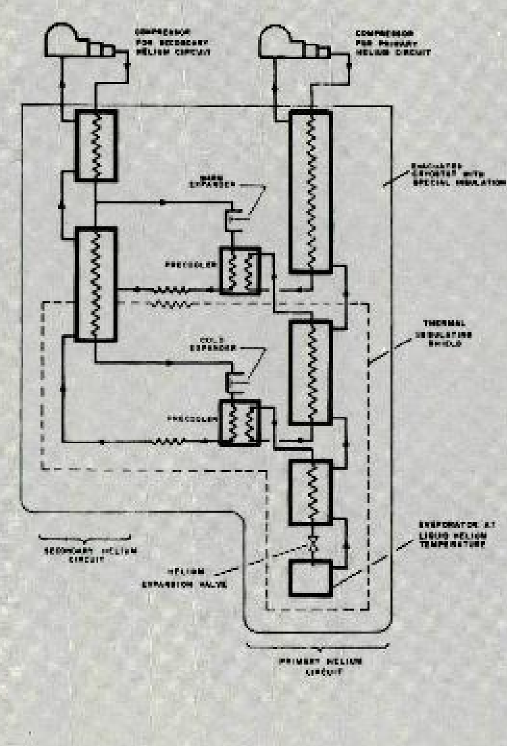


Fig. 11—Helium refrigerator.

the portion of the spectrum between 100 and 500 Gc will be spanned with sensitive receivers. The maser will be cheaper, smaller, and more reliable and less complex than any other existing low noise amplifier. Finally, the maser's inherent wide bandwidth, high gain, and stability will continue to improve.

Further advancement in this field depends to a great extent on the creative imagination of the electronic system designer and his ability to effectively utilize this new tool. Broadly stated, the areas requiring attention are analyses of the sources contributing to system noise, and designing to minimize the noise contribution of the system components.

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L. C. MORRIS was graduated from LaSalle College with a BA degree in Physics and Mathematics. He has completed the course requirements for the MS degree in Physics at the University of Pennsylvania. Mr. Morris was first associated with RCA in the Airborne Systems Activity, where he performed pulse and digital circuit design for video processing. His work included the transistorization of IF amplifiers, gating circuitry, and timing circuits, power supplies, and computer circuit design. At present, in DEP Applied Research, Mr. Morris is an engineering leader of several maser programs. He has performed studies in the area of generation, propagation, and detection of millimeter waves. He participated in the development of RCA's first solid state cavity maser and is presently participating in the development of traveling wave masers and other maser structures of a more advanced nature. He has performed studies and experiments with solid-state materials for masers and has helped in the evaluation of other low-noise amplification devices. Mr. Morris is a member of the IEEE and the Professional Technical Groups on Microwaves and Electron Devices.

The author, L. C. Morris, holds a travelling-wave maser; to the immediate right is the dewar.

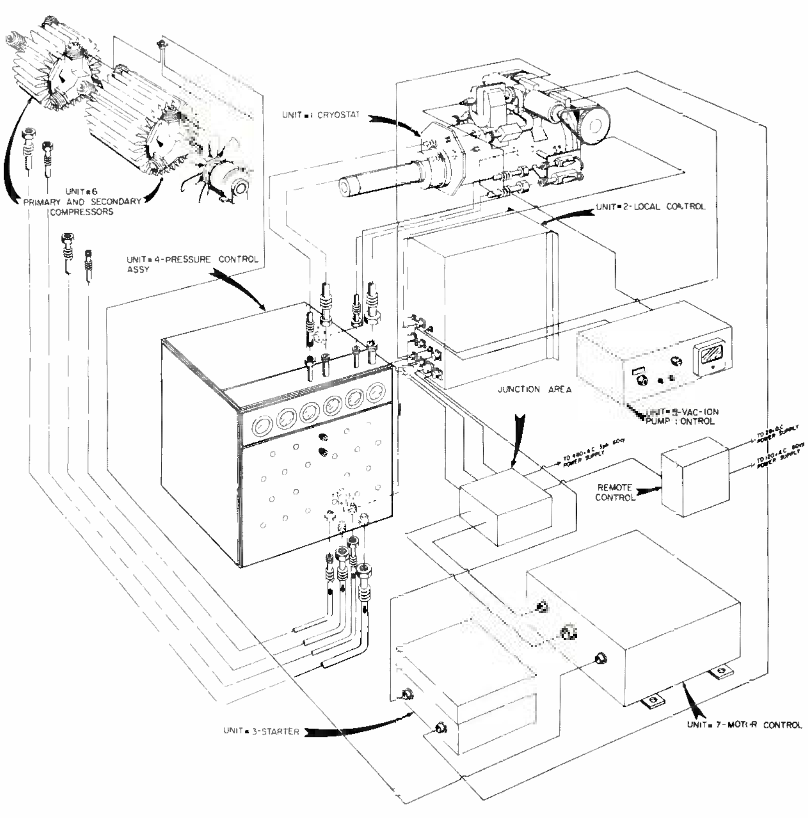

Fig. 12—Liquid-helium refrigeration system.

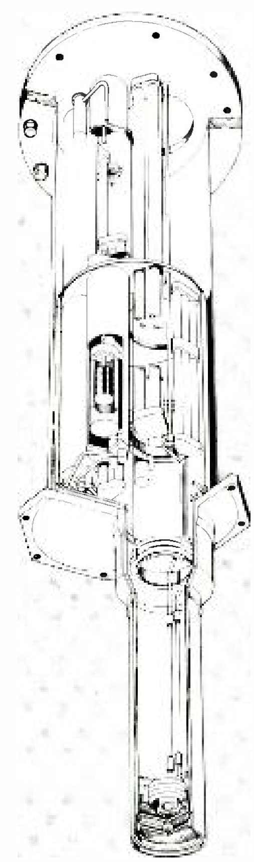

Fig. 13—Liquid-helium cryostat, 2.5°K operation.

REALIZING PARAMP POTENTIAL

. . . UHF, L-Band, and C-Band Parametric Amplifiers in Operational Radar Systems

In modern radar systems, ever-greater operating range and better target resolution are necessary to cope with the ballistic missile threat, and to detect and track space vehicles. These objectives can be met either by improving receiver sensitivity, or by increasing transmitter power. In recent years the development of the parametric amplifier and other advanced low-noise devices has opened the door to a great improvement in radar receiver sensitivity. The three systems described in this paper and their associated design problems cover a broad radio-frequency spectrum including UHF, L-band, and C-band, from 400 to 6000 Mc. No material is presented here on the theory of parametric amplifiers, since this aspect is available in other published papers.[1] Instead, the emphasis intended here is to show how research laboratory results only a few years old have been successfully translated into current operational radar-system hardware, making available the latest technological advances for better radar performance.

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DEP, Moorestown, N. J.

THE TREMENDOUS POTENTIAL of parametric amplifiers in the radar field was first realized when preliminary research laboratory results in 1957 verified their low-noise capabilities. These preliminary results, however, were obtained with laboratory units, a far cry from the practical circuits needed for modern radar system applications. Therefore, the microwave radar engineers job was to design and build parametric amplifiers providing this same low-noise performance, but without affecting any other radar functions unfavorably (e.g., tracking accuracy, moving target indicators, pulse compression, integration techniques, and range and velocity accuracy). At the same time, final amplifiers were to be capable of operation in extremes of environment, and with a reliability consistent with field use. This paper concerns three recent RCA radar system applications of parametric amplifiers and describes the amplifier-design solution developed for each system.

SYSTEM CONSIDERATIONS

The most attractive feature of the parametric amplifier for radar applications is, of course, its low-noise capability. The range of a radar system can be expressed as:

$$R = \sqrt[4]{\frac{P_t G^2 \lambda^2 A}{BF}} \qquad (1)$$

Where: P_t = peak transmitted power, watts; G = antenna gain factor (ratio); λ = transmitted wavelength, cm; A = effective target area, sq. meters; B = receiver bandwidth, cps; F = receiver noise figure (ratio); and R = range, nautical miles.

It can be shown that the range equation is consistent with the units given, assuming that the minimum detectable signal is just equal to the ambient noise level. This radar range equation can be used to demonstrate the importance of receiver noise figure in determining ultimate radar sensitivity, or range. *Reducing* the receiver noise figure 6 db (a ratio of four) has the same effect on the radar range as *increasing* the transmitted power by a factor of four. This noise-figure improvement can be effected in many existing systems simply by the use of a parametric amplifier at a relatively low cost—whereas a four-to-one power increase in a high-power radar transmitter to accomplish the same result would involve a tremendous increase in system size and cost. Therefore, it seems inevitable that parametric amplifiers or other low noise devices will be incorporated into most radar receivers; however, each system application will have its own peculiar problems to be considered in the design of the parametric amplifier. First, the type of amplifier to be used (i.e., upconverter, regenerative, traveling wave, degenerate) must be selected on the basis of system requirements.

Theoretical considerations show that the gain of a parametric amplifier is of the following form:

$$Power\ Gain = \frac{f_o}{f_i}\frac{K}{(1-\alpha)^2} \qquad (2)$$

Where: f_o is the output frequency, f_i the input frequency, K an efficiency factor (≤ 1), and α a regeneration factor related to pump power. The gain is dependent on two factors, frequency conversion and regeneration.

The relative stability depends chiefly on the regeneration factor—as α approaches unity, the gain increases rapidly, and a slight change in pump power results in a large change in gain. Therefore, for good stability, α should be as small as possible to minimize gain changes with pump power. This indicates that the stability of up-converters (with $f_o > f_i$) should be superior to that of regenerate amplifiers ($f_o = f_i$) with the same total gain. Optimum stability also dictates up-converting the input signal to a frequency as high as is practical, consistent with other system considerations, so that the required regeneration is as low as possible.

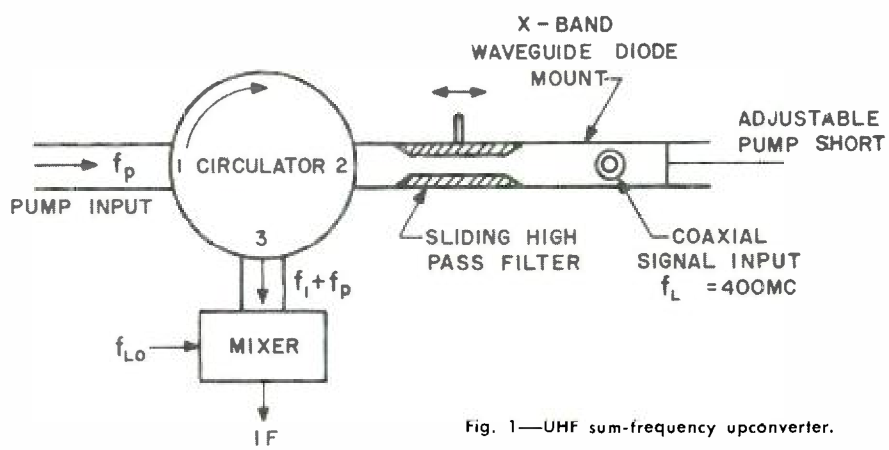


Fig. 1—UHF sum-frequency upconverter.

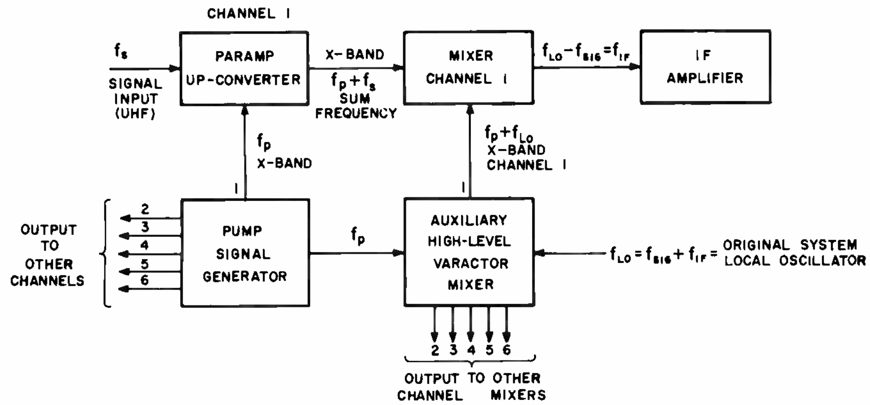


Fig. 2—Frequency-stable pumping scheme.

UHF PARAMETRIC AMPLIFIER

One defense application required the design of a system of parametric amplifiers for a high-power UHF monopulse tracking radar operating in the vicinity of 400 Mc; the system has six separate receiver channels, each requiring a parametric amplifier. Since the monopulse radar requires extremely stable receivers, the parametric amplifiers are designed to remain stable within ±3° in phase and ±0.5 db in gain over long periods of time. The desired noise figure for this application is less than 2 db, compared with a 5-db noise figure for the best operational vacuum-tube units.

The stability requirement suggested the use of an upconverter parametric amplifier. The final unit developed for this application uses an x-band circulator to separate the pump- and sum-frequency output signals, as shown in Fig. 1. A single varactor diode used in an x-band diode waveguide mount is matched for both pump and sum frequencies. The output signal is coupled to one arm of the circulator through a movable cutoff-type high-pass filter by which the difference frequency is reflected to provide internal regeneration.[2,3,4] The input signal is introduced coaxially to the diode, and tuning is done by means of a coaxial double-stub tuner.

The differential stability of the set of six upconverter amplifiers is assured by using a common pumping source for the amplifiers. To prevent an intermediate-frequency shift resulting from a possible change in pump frequency, an auxiliary x-band local-oscillator signal is derived by means of the circuit shown in Fig. 2. This arrangement utilizes an auxiliary high-level varactor mixer to generate the x-band local oscillator signal by combining a portion of the x-band pump with the basic UHF system local-oscillator signal. Thus, any change in pump frequency will result in an equal change in up-converted parametric amplifier output frequency and auxiliary local-oscillator frequency, so that the intermediate-frequency output will be unaffected. An extremely stable pumping source is used to maintain the ultra-stable phase and gain characteristics necessary for this application. The pumping source utilizes a feedback control system in conjunction with an x-band klystron so that the output frequency is held constant with respect to a tuned reference cavity.

The parametric-amplifier performance achieved in this system is shown in Table I. The design of a similar set of amplifiers was completed for use in a four-channel UHF monopulse tracking radar which is operating in conjunction with our national early-warning defense system.

L-BAND PARAMETRIC AMPLIFIER

The modification of an existing radar system to improve its receiver noise properties, without producing any detrimental effect on other radar functions, required the development of an L-band parametric amplifier. The L-band radar system is a tunable search radar operating in the vicinity of 1300 Mc; a unique one-knob tuning feature allows the operator to change transmitter frequency over a 100-Mc range while maintaining normal radar operation.

The parametric amplifier designed for this application is easily tunable so that it may be locked to the existing tuning functions. Other requirements satisfied were: an 8-Mc bandwidth, a low noise figure, sufficient gain to overcome noise contributions from succeeding stages, and a good short-term stability to preserve MTI operation.

A tunable narrowband amplifier of this type has certain advantages over a

TABLE I—UHF Paramp Performance

Signal Frequency Tunable Range,	400-450 Mc
Power Gain,	16 db
Pump Frequency,	9000 Mc
Amplitude Stability,	±0.5 db
Signal Bandwidth,	5 Mc
Noise Figure,	2 db (system)
Phase Stability,	±3°
Pump Power,	20 mw per amplifier

Fig. 3a—A basic sum-frequency upconverter with resonant circuit indicated.

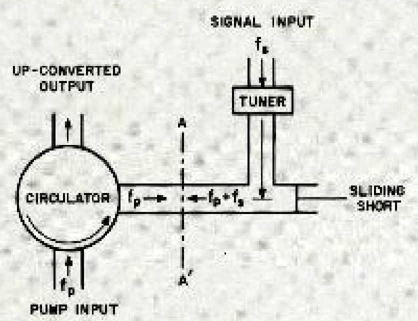


Fig. 3b—Band response of three-frequency upconverter.

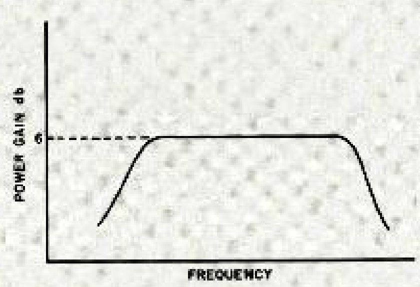
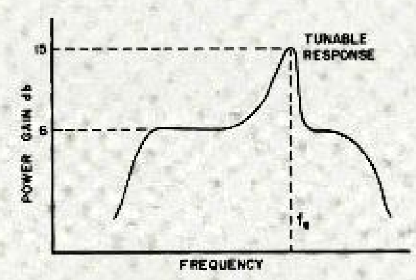


Fig. 3c—Band response of four-frequency upconverter.



wideband amplifier in radar system applications. An amplifier producing wideband gain over the full 100-Mc tuning range may be unattractive because of the image response present when used with a superheterodyne receiver; also, wideband amplifiers are more susceptible to interference problems from either friendly or unfriendly sources. The tunable parametric amplifier discussed below allows very rapid tuning of a narrowband receiver; when electronically controlled, the amplifier can be used in a frequency-diversity radar with pulse-to-pulse tuning accomplished in microseconds.

The sum-frequency upconverter amplifier developed for the search radar provides a very convenient means of tuning over extended bandwidths. Utilizing four-frequency operation with wideband upconversion gain plus tunable regenerative gain, a simple "one-knob" tunability isprovided and low-noise performance is maintained.

Conventional diode parametric upconverters involve the control of three frequencies: pump (f_p), signal (f_s), and difference ($f_p - f_s$) frequencies; or pump, signal, and sum ($f_p + f_s$) frequencies. The mode of operation chosen for this design is again sum-frequency upconversion (the output is taken at $f_p + f_s$) with loading at the difference frequency producing regeneration and resultant increased power gain.

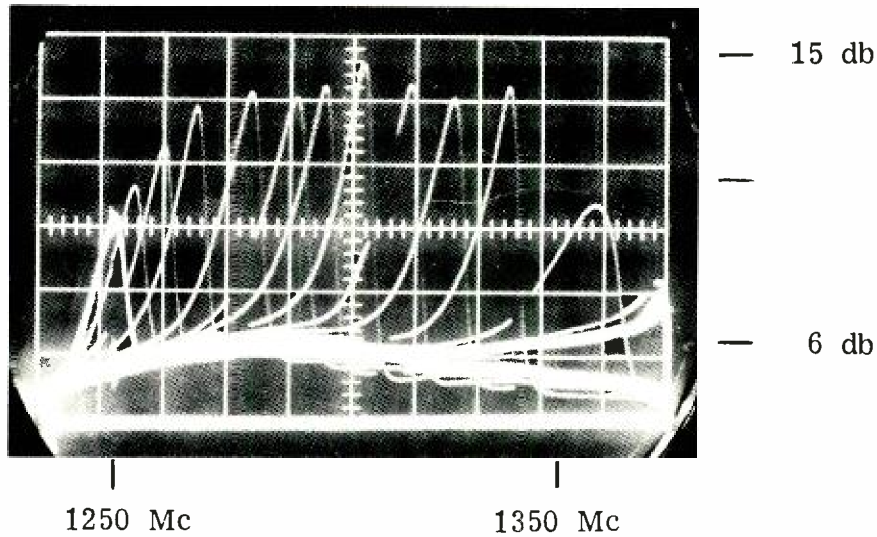
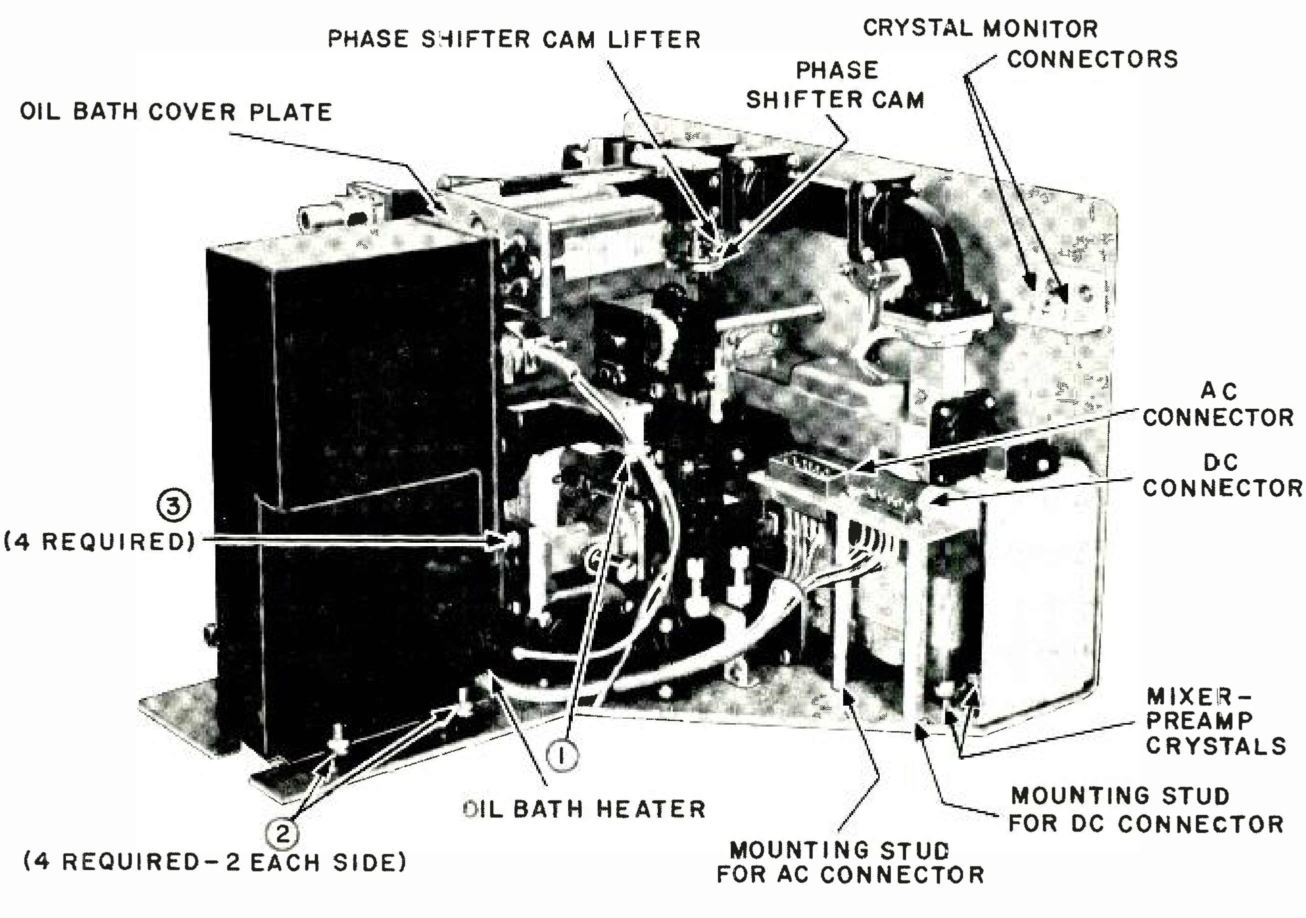


Fig. 4—An oscilligram showing the bandpass response for the L-band upconverter.

A sketch of the basic sum-frequency upconverter is shown in Fig. 3a with the various resonant circuits indicated. This amplifier is of the same general type as the UHF upconverter previously described; it can be tuned to obtain the response indicated in Fig. 3b. When a restricting section of waveguide (having a good RF match at both sum and pump frequencies but which is below cutoff at the difference frequency) is inserted in the amplifier, its position relative to the varactor diode can be adjusted for resonance at the idler (difference) frequency. Such an amplifier has the response shown in Fig. 3c. Resonating the difference frequency produces regenerative gain; this method for producing regeneration can be varied quite simply to tune the amplifier over a considerable bandwidth.

In the experimental amplifier, tunabil-

Fig. 5—Tunable L-band parametric amplifier.



ity is obtained by inserting a movable dielectric vane in the waveguide section between the diode and the highpass filter; a ferrite phase shifter may also be used in place of the dielectric vane to tune the amplifier electronically. Using a silicon varactor diode, the experimental amplifier operates from 1250 to 1350 Mc, with a fixed pump frequency of 8800 Mc. The amplifier produces the expected response, is tunable over the complete frequency range with an instantaneous bandwidth of 8 Mc, and has a noise figure less than 2.0 db. Fig. 4 shows bandpass response oscilloscope traces for this amplifier, illustrating various tuning conditions. Packaged models of the amplifier such as that shown in Fig. 5 have been incorporated in several radar systems for evaluation. Range increases commensurate with the noise figure improvement have been realized; no adverse effects on other radar functions have been observed.

C-BAND PARAMETRIC AMPLIFIER

A system of three C-band parametric amplifiers was designed for an RCA monopulse tracking radar application. These amplifiers were required to exhibit excellent differential gain and phase stability, provide a 20-Mc bandwidth, be tunable between 5.4 Gc and 5.9 Gc, and have a minimum noise figure. The amplifiers were to be mounted directly behind the antenna in a restricted space where they must withstand severe shock and large ambient temperature variations. Continuous tracking capability in case of amplifier failure during tracking operations was another requirement of this application.

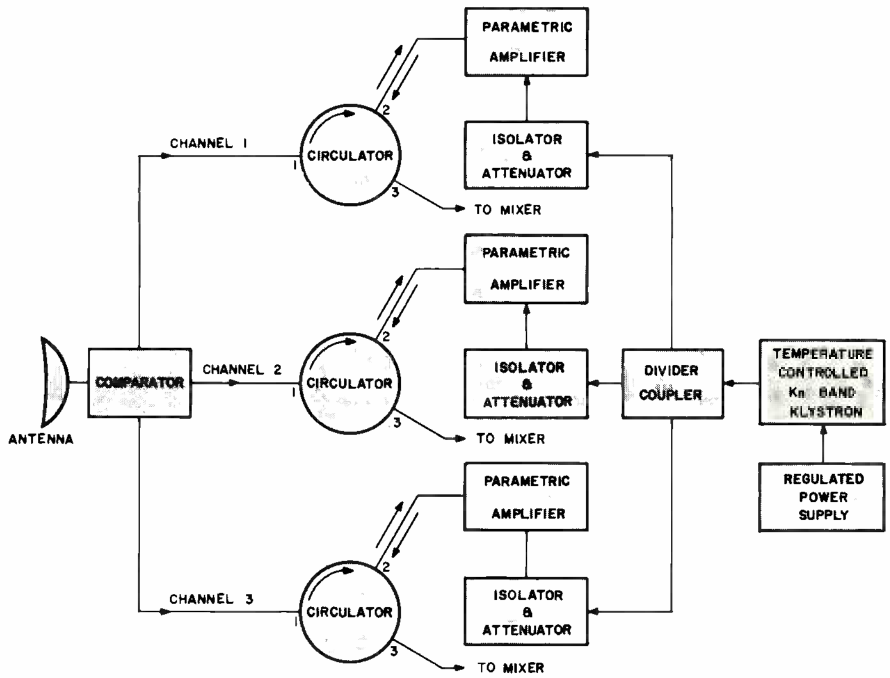


Fig. 6—Three-channel C-band parametric amplifier.

The regenerative type of parametric amplifier proved to be the best suited to meet this long list of design objectives: the regenerative amplifier uses a minimum number of components, and is thus compatible with the small amount of available space. To achieve the required differential phase and gain stability, a 250-mw pump supply provides the common pump source for the set of three amplifiers. The optimum theoretical pump frequency for low-noise performance was calculated to be between 16 and 19 Gc, which led to the use of a KU-band pump klystron. Since a feedback system for control of pump frequency was discarded because of space considerations, pump stability had to be assured by careful regulation of both the klystron temperature and the klystron supply voltages. A block diagram of the system is shown in Fig. 6. In the event of pump or diode failure, the amplifier reflects the incoming signal into the mixer. Substantial bracketing as well as mechanical locks on all variable adjustments were provided in order to maintain stable electrical operation of the parametric amplifiers during the severe shock and vibration conditions encountered in radar operation.

The C-band parametric amplifier itself consists of the waveguide tee junction shown in Fig. 7. Waveguide sections forming the tee junction include a modified KU-band section used to couple pump power at 16 Gc into the diode, yet isolate the pump circuit from the idler or difference frequency. The half-height X-band waveguide section, which forms the idler cavity, provides the necessary resonant circuit at the difference frequency (10.1 to 10.6 Gc); this allows regenerative amplification at the signal frequency. The signal is fed coaxially into the diode, which is mounted at the center of the junction. As shown in Fig. 7, tuning is provided for pump, idler, and signal frequencies. A circulator is provided to separate the input signal from the amplified output signal.

A typical set of electrical characteristics obtained with this amplifier is listed in Table II.

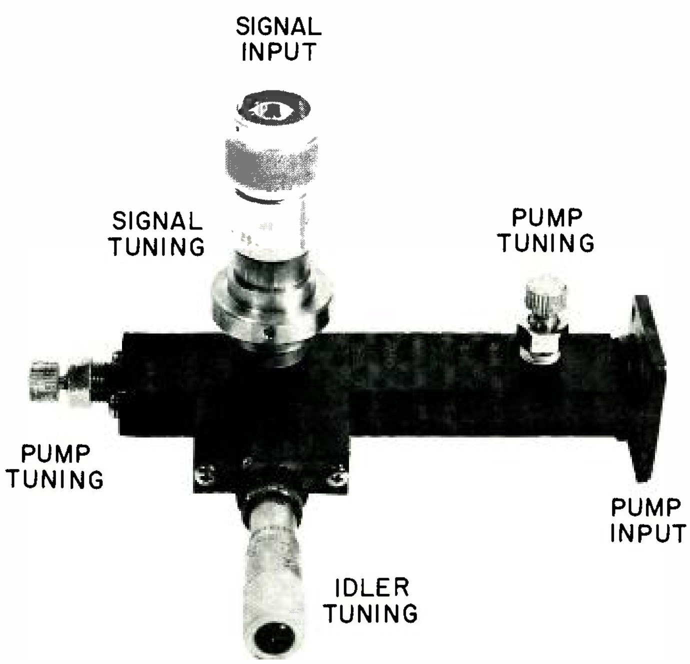


Fig. 7—C-band parametric amplifier.

TABLE II—C-Band Paramp Characteristics

Frequency, tunable 5.4-5.9 Gc
Instantaneous Bandwidth, 20-50 Mc
Gain, 16-20 db
Receiver Noise Figure, 3.5 db (system)
Phase Stability (differential), ±3° (long and short term)
Amplitude Stability (differential), ±0.5 db (long and short term)

Field tests showed that an increase of approximately 60 percent in range resulted when parametric amplifiers were inserted ahead of the conventional crystal mixers previously used in the receiver of a typical instrumentation-type radar. Stable operation of the parametric amplifier without readjustment was obtained in this system for a period of two months. A completed system installation is shown in Fig. 8.

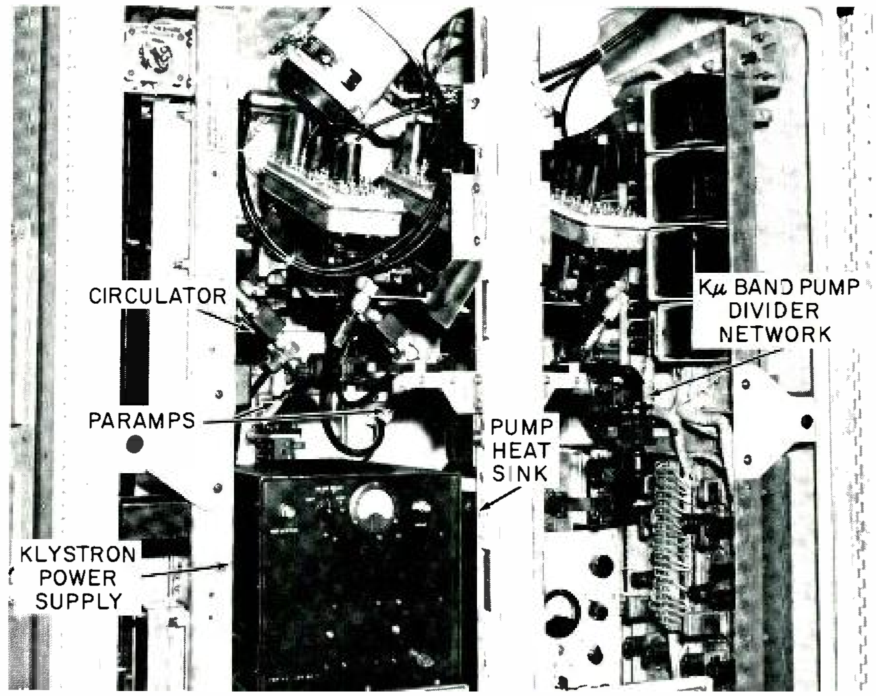


Fig. 8—A typical parametric amplifier installation.

CONCLUSION

These examples demonstrate that parametric amplifiers have been successfully incorporated into operational radar systems and their low-noise potential effectively realized. The parametric amplifier has, in effect, made a timely transition from the research lab to the radar system.

ACKNOWLEDGEMENT

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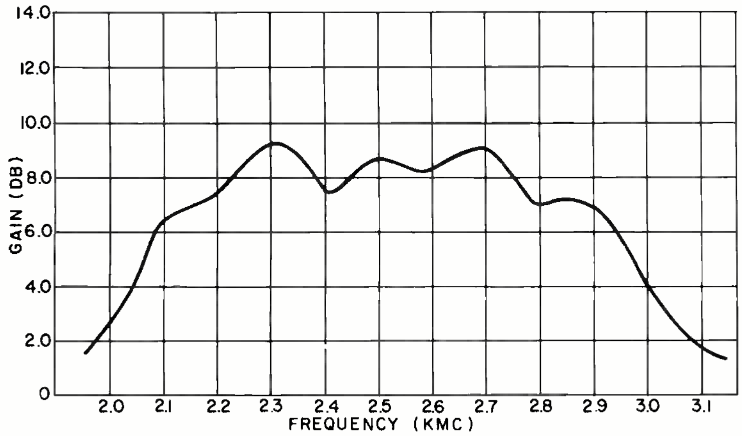

Fig. 1—Gain response of broadband parametric amplifier.

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BROADBAND PARAMETRIC AMPLIFIERS BY SIMPLE EXPERIMENTAL TECHNIQUES

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While a rigorous analytical approach should be utilized wherever possible in achieving broadband parametric amplification, it is often advantageous to use experimental methods. Described here is a method for building broadband parametric amplifiers of either the reflection or upconvertor type having large gain-bandwidth products and a minimum of additional network elements and circuitry. While not a fundamentally new approach to parametric device theory, it allows more efficient use of the components of a typical paramp. Main features are: maximum utilization of parasitic elements of the varactor diode; waveguide techniques that give an exceptionally broadband idler circuit; and simple diagnostic techniques for rapid circuit evaluation.

THE VARIABLE-CAPACITANCE parametric amplifier represents an important step forward in the field of low-noise UHF and microwave communications. Since parametric-amplifier operation does not involve any noise-generating mechanisms, very low receiver noise figures can be and in fact have been achieved even at high microwave frequencies.

Residual noise in a practical parametric amplifier is contributed primarily by the semiconductor diode spreading resistance R_s, and by unavoidable ohmic losses in the amplifier structure. By proper hardware design using diodes of high-cutoff frequency, these noise contributions can be minimized to a point where measured noise figures approach those predicted by the theory.

Relatively good noise performances were achieved soon after the solid-state parametric amplifier was reduced to practice; yet, it proved much harder to obtain bandwidths broader than a few percent because of the reactive nature of the device, and the fact that resonant circuits at the signal, idler, and pump frequencies are essential for proper functioning of the amplifier.

NETWORK FILTER APPROACHES

Since it is possible to represent the pumped-variable-capacitance diode by a linear equivalent circuit, a logical step toward broadbanding the device is to apply the well-developed techniques of modern network theory.

This has been done and articles have appeared in the literature which analyze and propose various filter-type approaches for achieving broadband parametric amplification.[1-12] Resultant networks have yielded excellent theoretical gain-bandwidth products; in many cases, predicted performance has been realized in practice.[7]

Good correlation between theory and practice occurs when the networks designed on paper are realizable using hardware applicable to frequencies of interest; this is not easy at microwave frequencies where inductances and capacitances must be simulated by distributed elements. Moreover, sections of the transmission line coupling the individual network elements to the varactor diode have frequency sensitivities of their own which must be considered in transforming a lumped element design into a distributed microwave equivalent, especially when broad (20- to 50-percent) bandwidths are involved.

VALUE OF EXPERIMENTAL APPROACH

While the rigorous analytical approach should be considered wherever applic-

* Credit is due R. Pettai, formerly with RCA and now with Micro-State Electronics Corp., who contributed significantly to the writing of this paper.—*B. B. Bossard.*

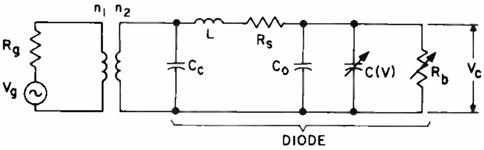

Fig. 2—Varactor diode terminating the input line. (See Equation 1, text.)

able, it is often advantageous to use various experimental methods; such criteria have been developed for building broadband parametric amplifiers of either the reflection or upconverter type having large gain-bandwidth products and a minimum of additional network elements and circuitry.[12] The design method described herein does not constitute a fundamentally new approach to the theory of parametric devices, but it does lead to a more efficient use of the existing building blocks of a typical parametric amplifier. The main features of the technique could be summarized as follows:

1) Maximum utilization of parasitic elements (R_s, L, C_o, C_c) of the varactor diode;
2) Suitable waveguide obstacles in the plane of the diode result in an exceptionally broadband idler circuit;
3) Simple experimental techniques permit a rapid evaluation of given circuitry.

To illustrate the value of this method, a reflection-type parametric amplifier has been developed with the characteristics listed in Table 1, and having the typical gain response shown in Fig. 1.

EQUIVALENT PARAMETRIC AMPLIFIER NETWORK

The experimental technique is best explained by discussing separately the input and the output circuits of a typical parametric amplifier. Thus, a variable capacitance diode terminating a transmission line (for example, the co-axial input line of a parametric amplifier) can be represented by the equivalent circuit of Fig. 2, where a transformer has been added to adjust the impedance level. The assumption has been made here that the signal frequency cannot propagate in the pump and the idler circuits; such circuits are commonly found in waveguides, and since signal frequency is usually much lower than the other two frequencies, the above assumption is justified. One may thus ignore any first-order loading effects of these circuits on the equivalent circuit of Fig. 2.

The quantity of interest is now the ratio $|V_c/V_g|$ as a function of frequency and of the diode parameters. This ratio can be looked upon as a measure of transmission of the input voltage to the variable capacitance $C(v)$, and can be expressed by:

$$\left|\frac{V_c}{V_g}\right| = \frac{\left(\frac{n_2}{n_1}\right)}{\left[1-\left(\frac{\omega}{\omega_o}\right)^2-\omega^2 R'_g R_s C_o C_c\right]^2+\left(\omega C_o\right)^2\left[R_s+R'_g+R'_g\frac{C_c}{C_o}\left\{1-\left(\frac{\omega}{\omega_o}\right)^2\right\}\right]^2} \qquad (1)$$

Where: the parameters of special interest are the transformer ratio n_2/n_1, and the self-resonant frequency of the diode given by:

$$\omega_o^2 = \frac{1}{L\,C_o} \qquad (2a)$$

also:

$$R'_g = \left(\frac{n_2}{n_1}\right)^2 R_g \qquad (2b)$$

Fig. 3 shows the results of some calculations based on Eq. 1 using typical values for the diode parameters. Two different transformer ratios were tried and as expected the dependence of $|V_c/V_g|$ on frequency and loading (i.e. on the ratio n_2/n_1) is that of a tuned RLC circuit. It is evident from Fig. 3 that by choosing the self-resonant frequency ω_o and the transformer ratio n_2/n_1, one is able to control the transmission of the input voltage to the variable capacitance. Given a certain band of frequencies, it is thus possible to obtain the most efficient transmission over the given band to ensure a maximum interaction between the signal and the pump frequency.

BROADBANDING CRITERIA

Actually the importance of above conclusion lies in the fact that it suggests an experimental procedure for testing diodes and input circuits, and leads directly to the first of the two broadbanding criteria described below.

As seen from Fig. 2, the voltage $|V_c|$ which represents the useful signal voltage developed across the variable capacitance, is also across the non-linear barrier resistance R_b. By measuring the rectified output of the diode while sweeping the input signal over a band of frequencies, it is possible to ascertain the frequencies at which the input circuit and the diode yield the best transmission of the input voltage to the variable capacitance $C(v)$. This is so since a maximum rectification implies a maximum voltage across C_o and indicates varactor series resonance.

Fig. 4 shows some experimental results obtained by the above technique. As seen from the plot, the measured curves do resemble those obtained from Eq. 1. This technique rapidly checks the diodes at hand to find the unit best suited for an application. Further bandwidth adjustments are then made by altering the transformer ratio n_2/n_1 and the bias voltage applied to the diode. Since, as a second-order effect, the geometry of the mount does have some influence on the voltage distribution across the diode at the signal frequency, changing the mechanical design slightly may also help.

The first criterion for broadbanding is now simply that the diode in its mount should show good rectification efficiency throughout the proposed signal band—i.e., the varactor diode should series-resonate at the signal frequency.

A somewhat different situation exists in an upconverter idler circuit which also becomes the output circuit of the device. Here, a common configuration is a varactor diode shunt-mounted in waveguide; the mount (a short section of waveguide) thus constitutes a two-port structure at the idler frequency. The coaxial signal input line usually contains either a lowpass filter or a choke to block the idler and is therefore not seen by the idler frequency.

When one port of the mount is terminated in a matched load, the diode-mount combination exhibits a two-port transmission characteristic of the bandpass type; this characteristic can be easily determined by plotting the insertion loss or the VSWR of the diode-mount combination (to expedite the work, a sweep generator was used). The extent and the quality of the passband is determined not only by the diode parameters, but also by the type of mount or obstacle holding the diode. Fig. 5 shows some typical results obtained with an MA460E cartridge diode mounted on different obstacles. Since the application of a DC bias changes the C_o of the diode, it is not surprising that the passbands are, to a lesser extent, also functions of the bias voltage (Fig. 6).

This phenomenon has also been observed by DeLoach[13] who showed that a

TABLE I—Characteristics for Experimental, Reflection-Type Parametric Amplifier.

Signal Band	2000-3000 Mc
Gain	9.0 db
3-db Bandwidth	870 Mc
Noise Figure	2.5 db
Pump Frequency	12.400 Mc

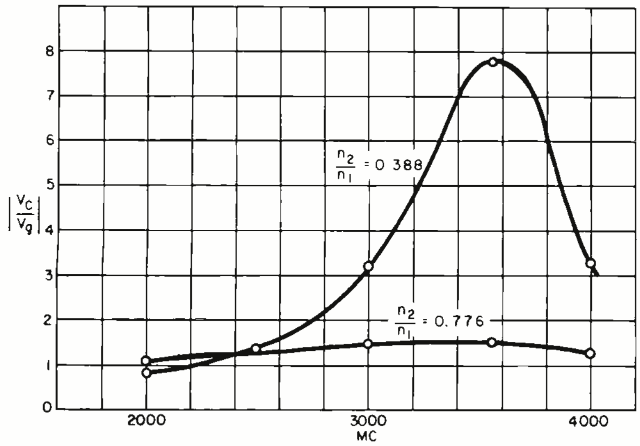

Fig. 3—$|V_c/V_g|$ as a function of transformer ratio.

Sharpless type diode properly mounted on a transverse capacitive ridge in WR90 waveguide exhibits a bandpass characteristic. In his experiment the bias was also used to adjust the exact position of the passband.

The second broadband criterion now requires that for broadband performance the idler frequency must be placed within the observed passband —i.e., the varactor should shunt-resonate at the idle frequency. By satisfying both requirements, i.e. maximizing the $|V_c/V_g|$ ratio and by operating within a passband at the idler frequency, it has been possible to develop single-channel voltage gain-bandwidths approaching 3300 Mc with only one diode.

To show why the above two requirements lead to a good performance, certain assumptions are made regarding the equivalent circuit of the diode-obstacle combination. *First*, a maximum $|V_c|$ across the variable capacitance means that the amplitude of the idler current $|I_2|$ generated by the pumping action will also be a maximum, since from the four-terminal equivalent circuit of a pumped varactor diode:

$$|I_2| = \omega_2\left(\frac{\Delta C}{2}\right)V_c \qquad (3)$$

Secondly, since the self-resonant frequency is much lower than the idler frequency, the diode at the latter frequency appears inductive. (It will be recalled that in the present method, the self-resonant frequency of the diode is placed at the input signal frequency.) Because the diode-obstacle combination shows a passband, a rough equivalent circuit of the waveguide is as shown in Fig. 7a, where L denotes the lead inductance, C_o the average bias point capacitance and C_x the equivalent capacitance of the obstacle. From Fig. 7a, it is easy to see that for the admittance of this shunt branch to vanish, then:

$$B_o\,(1 - X_L B_x) + B_x = 0 \qquad (4)$$

In actual operation, the circuit shown in Fig. 7b applies where the voltage capacitance $C(v)$ across C_o becomes effectively a generator of the idler frequency, whose output is given by Eq. 3. The load G_L represents the actual output load in case of a parametric upconverter. In case of a reflection-type parametric amplifier, G_L can be considered as the loading present in the idler circuit.

It is of interest to calculate the power P reaching G_L:

$$P = \frac{|I_2|^2 G_L}{G_L^2[1 - B_o X_L]^2 + [B_o(1 - X_L B_x) + B_x]^2} \qquad (5)$$

which is indeed maximized if the condition (4) holds.

It is true that C_x could be replaced by an idler tuning plunger as is commonly done. Some designs use a waveguide cross or a tee with adjustable shorting plungers in the side arms to tune the idler frequency. While this is satisfactory for narrowband applications, the frequency sensitivity of short-circuited lines used to simulate the required B_x makes is difficult to achieve high gain-bandwidth performance. The advantage of the diode-obstacle combination is that the frequency sensitivity of such a compact resonator is not dependent on any critical line length. On the other hand the performance of the configuration can be easily checked by simple *VSWR* or insertion loss measurement techniques.

In some upconverter designs, a high-pass filter (section of waveguide below cutoff at f_i) is commonly placed between the diode and the pump input line to keep the idler out of the pump circuitry. At the idler frequency the filter acts effectively as a shorting plunger and does contribute some frequency sensitivity to the system. However, since the main tuning of the idler circuit is now accomplished by C_x, the short-circuit plane of the highpass filter is no longer needed for this purpose. As a result, the short circuit plane can be placed exactly $\lambda_g/4$ away from the diode, thus minimizing the variation with frequency of the Y_{short} as seen by the diode.

Furthermore, if the pump frequency can be placed outside the observed passband, the leakage of pump power into the idler circuit can be reduced. This inherent filter action is useful in applications where direct leakage of pump power into the output circuitry (as in a parametric upconverter) is undesirable. An extra blocking filter may thus be necessary.

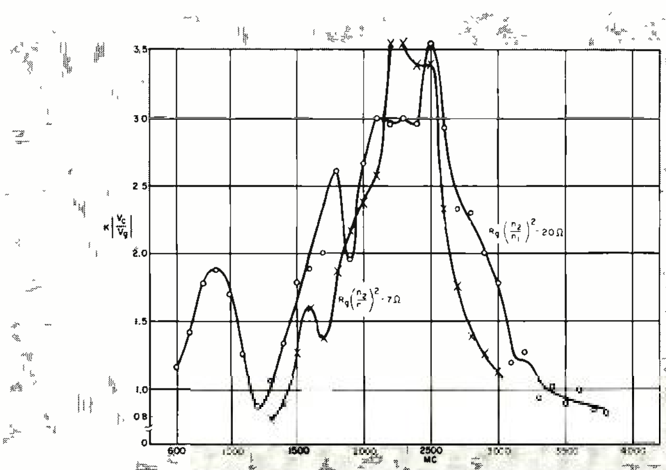

Fig. 4—Experimental results for $|V_o/V_g|$.

PERFORMANCE RESULT

In addition to the results shown in Fig. 1, this technique has yielded the performance shown in Table II.

With reduced gain (2 to 5 db), single-channel amplification bandwidths of 2000 Mc have been achieved at 3 Gc.

A brief discussion of the noise performance is of interest. In all of our experiments with reflection-type amplifiers, a controlled amount of extra loss was used in the idler circuit to lower the idler

TABLE II—Performance of the Experimental Paramp.

Signal Frequency	2200 Mc	2500 Mc
Gain at Midband	20 db	11.2 db
3-db Bandwidth	330 Mc	650 Mc
Noise Figure	(Not Measured)	

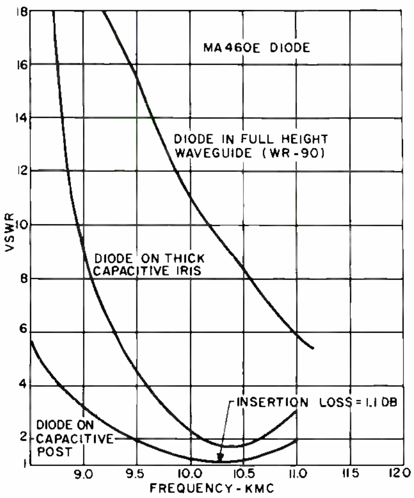

Fig. 5—Effect of obstacle on diode passband.

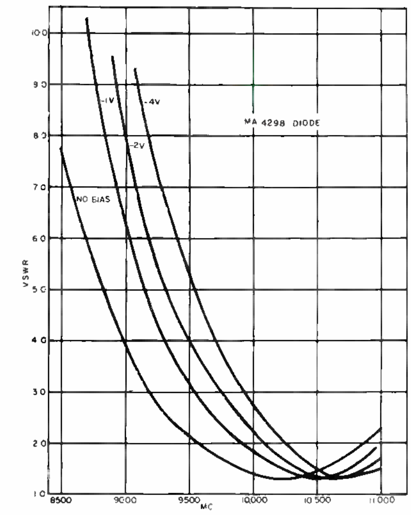

Fig. 6—Effect of bias voltage on diode passband.

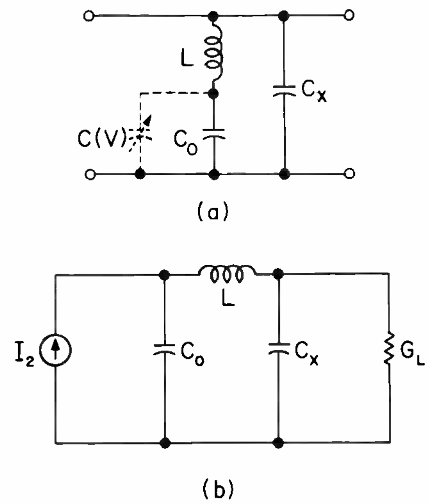

Fig. 7—Idle equivalent circuit.

circuit Q for broadbanding purposes. In practice, this was done by terminating the idler circuit in a waveguide load and using tuners between the load and the diode to transform the G_a to a value (typically 0.3 G_a or less) required for the optimum gain-bandwidth product. The loading conductance does, of course, contribute some noise and will degrade the noise figure of the amplifier; however, appreciable broadbanding can be obtained before serious degradation in noise figure occurs. It is difficult to establish a firm criterion here, since much depends on the intended application of the device. If the lowest possible noise figure is required, the spreading resistance R_s of the diode should clearly constitute the only loading. (The concept of a diode-obstacle resonance is still applicable and the combination should be tested accordingly; this ensures that the bandwidth remains the maximum available under this low loading condition.) The broadband concepts described above have been utilized in the recent development of a 3-Gc paramp with a 1.5-db noise factor and a single tuned bandwidth of 2 percent.

On the other hand, there are applications where broad bandwidths are of prime importance. In such cases, a slight degradation in noise figure may well be a tolerable compromise.

The concept of series resonating the varactor diode at the signal frequency in a coaxial transmission line while shunt resonating the same diode at the idling frequency by means of a waveguide obstacle is extremely useful in the development of tunable parametric amplifiers. For example, a fixed-pump-frequency, single-knob, mechanically tuned paramp has been demonstrated with a 16-db gain and a tuning range of 900 Mc. This device uses the broadband input circuit in conjunction with a variable asymmetric capacitive window, C_x (see Fig. 7a). The broadband paramp can also be tuned over a 1000-Mc frequency range by simply changing the pump frequency. Constant instantaneous single-tuned bandwidths of 60 Mc were obtained for amplifier gain levels of 16-db.

CONCLUSION

The broadband criteria just described should help the designer of a parametric amplifier make a better evaluation of the various structures and diodes at hand. The technique is simple to use and wherever applicable (i.e. wherever the specifications allow some freedom in selecting the structure and fixing the critical operating frequencies) it leads to a good gain-bandwidth performance. Since the gain of a parametric amplifier is a function of many variables, such as ΔC, circuit impedance (source and loading impedances as seen by the diode), ohmic losses in the circuit, etc., the existence of passbands at the operating frequencies is not a sufficient condition for optimum performance. Their existence does, however, appear as a necessary condition for a large gain-bandwidth.

ACKNOWLEDGEMENT

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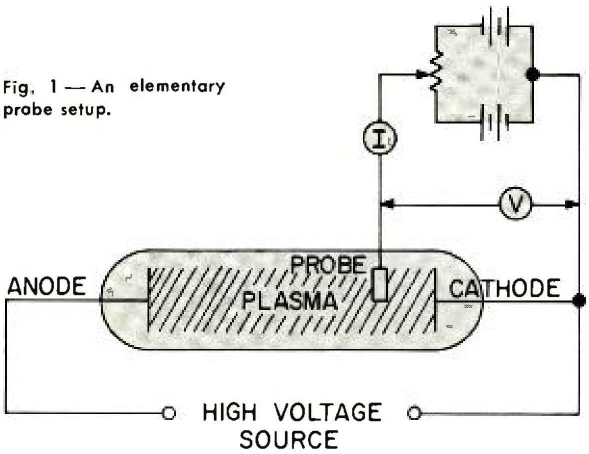

Fig. 1 — An elementary probe setup.

MICROWAVE PROPERTIES OF PLASMA

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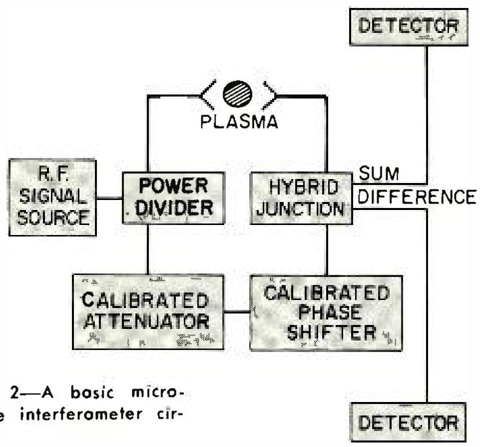

Fig. 2—A basic microwave interferometer circuit.

This review is slanted toward the effects of plasma on the radar-return, beacon, or telemetry signals from an object in an environment containing plasmas. Discussed are plasma measurements, microwave propagation in plasmas, plasma re-entry effects, and microwave plasma devices. A bibliography of basic references is included.

For our purposes, a plasma is generally considered to be an electrically neutral cloud, or a collection of very numerous ionized particles. Interest in the microwave properties of such plasmas arises from two somewhat different sets of considerations: 1) the application of plasmas to microwave devices, and 2) the effects of plasmas as an environment on a microwave signal.

The use of plasmas in microwave devices is of long standing, in TR tubes and noise tubes, for example. However, recent interest centers around use of the properties of plasmas as a dispersive medium with controllable parameters. Similarly, the effects of the ionosphere on high-frequency propagation have been studied for some thirty years, but in this paper our interest is in the effects of plasmas on the radar-return, beacon, or telemetry signals from an object in an environment containing plasmas.

BRIEF HISTORY OF PLASMA PHYSICS

The study of plasmas in recent years stems from the studies of the phenomena of electrical discharges carried out in the decade 1923 to 1933, by I. Langmuir[1] in association with H. Mott-Smith,[2] and L. Tonks,[3] by J. J. Thomson and G. P. Thomson,[4] and by many others. The classic paper on the phenomena of plasma resonance and oscillation is that of Langmuir and Tonks.[3] In that paper, a theory was proposed which accounted for most of the anomalous behavior discovered in the study of electric discharges up to that time. This same idea of plasma resonance and oscillation is one of the fundamental concepts underlying the explanation of the propagation of electromagnetic waves in an ionized medium. This is the starting point of modern plasma physics.

The study of electric discharges has continued with significant results in the development of modern high-power vacuum devices, noise tubes, etc. used in radio and radar. As radio developed back in the early days, the effects of the ionosphere on HF transmission were observed and studied. Theoretical work on this subject dates back to Kennelly and Heaviside in 1902, but the development of present-day theory stems mainly from the work begun by Larmor in 1924 and by Appleton and others subsequently,[5] continuing to the present day. The importance of this work from our point of view is the emergence of the *magneto-ionic* theory of propagation of electromagnetic waves through the ionosphere in the presence of the earth's magnetic field. This is another of the fundamental concepts underlying the study of the microwave properties of plasmas.

Work in many other fields has contributed to the present state of plasma knowledge, chief among them being magnetohydrodynamic power generation, and hypersonic flow of gases about a vehicle. In both these fields, microwaves are utilized to study the plasma. In such diagnostic techniques, microwaves are sent through a plasma and are affected by it; from measurement, of these effects, some properties of the plasma may be deduced.

Somewhat closer to home, there has been the development of ferrites. Both plasmas and ferrites, when immersed in a magnetic field, are gyromagnetic media, i.e., media whose properties of propagation are significantly affected by changes in an applied magnetic field.[6] In a ferrite, the tensor permeability $[\mu]$ is affected; in a plasma the tensor dielectric constant $[\epsilon]$ is affected; but the treatments of propagation in plasmas and ferrites are quite analogous, mathematically.

PLASMA MEASUREMENTS

Consider a volume of ionized gas whose properties must be measured. Two questions arise: *What is to be measured?*

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How is it to be measured? Since a plasma is a collection of positive, negative, and neutral particles, it may be desirable to know the following: 1) how many of each particle, 2) what are these particles, 3) some information on the energy of these particles, 4) their speed, or 5) particle density distribution. Some of these properties may be measured by considering a plasma to be an ordinary gas, especially when only a very small percentage of the gas atoms are ionized. At other times, useful information can be obtained by considering the plasma to be a gas composed only of the free electrons. The latter assumption is significant when the electron moves around more freely, i.e., has high mobility. Occasionally, as in hydromagnetic waves, ionic considerations are most important. Now, to measure these plasma properties, which can be represented by such different models, requires new measurement techniques.

A plasma gives off light, heat, and radio noise and may be characterized by a special shape or form. These properties and many others may be used to deduce the make-up of the plasma. One of the oldest methods, still used today, is to stick one or more probes into the plasma to find out what is going on inside. When the probe is held at a definite potential relative to the surrounding discharge, it will attract either positive or negative charges, forming a *sheath* about it. This sheath will be a controlling factor in the currents collected at the electrode. It can be shown[2] that the currents to the probe are a function of the electron velocity distribution as the potential of the probe is changed; from this, the plasma density may be indicated. If a Maxwellian velocity distribution can be assumed, then the mean temperatures of the electrons and the plasma density may be calculated in a straightforward manner. Probe techniques work quite well over a wide range of plasma temperatures and densities; however, they sample only the region of the plasma right near their position. Here are some additional problems with probes: The presence of the probe perturbs the plasma; the shape and size of the probe affects the measurements; the probes are not readily moved about; and they can contaminate the plasma. Nevertheless, probe techniques continue to be developed and used to advantage in certain phases of plasma work (see Fig. 1).

A family of techniques called *microwave diagnostics* has been available in recent years for the study of plasma by means of microwave instrumentation.[7,8] Such methods have been quite successful as energy levels necessary for dependable instrumentation are often low enough to minimize perturbation of the plasma while the measurement is taken. One of the diagnostic methods is that of the *microwave interferometer;* Fig. 2 illustrates the basic microwave circuit.

The plasma is inserted in one branch of the interferometer and a null is obtained. This is compared to the initial null and the phase shift and attenuation introduced by the plasma is known. These quantities may be related to the plasma parameters of electron density and collision frequency. An example is given in the following discussion of a particular plasma column in waveguide and how the plasma parameters were found from the measured interferometer data.

MICROWAVE PROPAGATION IN PLASMAS

The thing that is different about a plasma as a propagation medium is the presence of charged particles. The differences between the propagation characteristics of an ionized gas as a medium and a gas without charged particles must be explained by the behavior of those charged particles in the presence of electric and magnetic fields. A simple way is to consider first the effect of constant electric

Fig. 3a—A charged particle in an electric field.

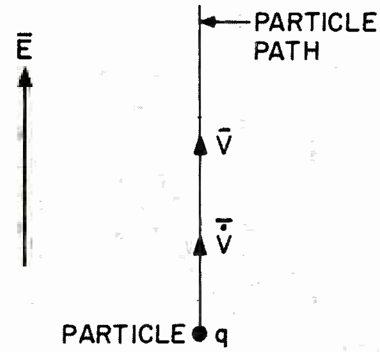
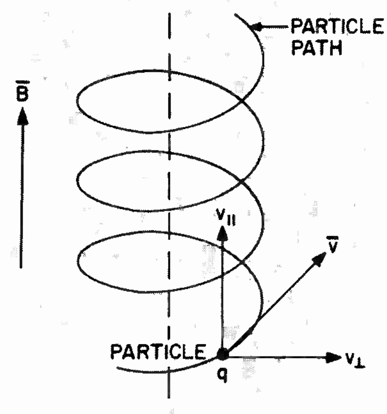


Fig. 3b—A charged particle in a magnetic field.



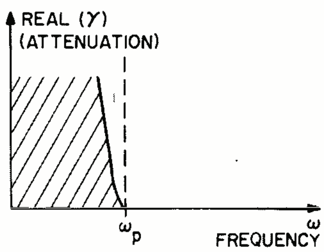


Fig. 4a—A collisionless plasma transmitted wave.

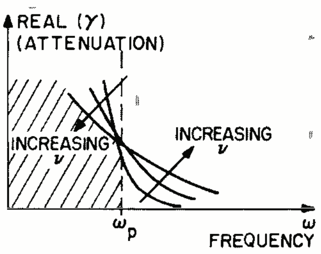


Fig. 4b—Effect of collisions on transmitted wave.

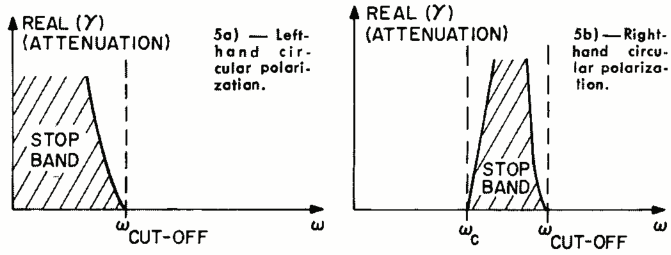


Fig. 5—Propagation parallel to B_o external magnetic field, transmitted wave ($y = O$).

and magnetic fields on charged particles. For an electric field alone, the equations of motion are:

$$\bar{F} = q\bar{E} = m\dot{\bar{v}} \qquad (1)$$

For a magnetic field alone:

$$\bar{F} = q\bar{v} \times \bar{B} = m\dot{\bar{v}} \qquad (2)$$

Where $\bar{F}$ is force, $\bar{E}$ and $\bar{B}$ are electric and magnetic fields, q is charge, m is mass, $\bar{v}$ is velocity and $\dot{\bar{v}}$ is acceleration, and MKS units are used throughout. From this, it is seen that an electric field accelerates a charged particle parallel to itself (Fig. 3a) and a magnetic field accelerates a moving charged particle (Fig. 3b) so that in general, it will spiral about the magnetic field lines while moving with velocity component v_{11} along them.

From Equation 2 and the laws of dynamics:

$$\omega_c = \frac{qB}{m}$$

Where: ω_c is the angular frequency in radians per second of the spiralling motion, and is called the *cyclotron frequency*.

A plasma might also be considered a "jelly" with electrons embedded in it. This is not as bad a picture as one might think, as the neutrals and ions are so much heavier than the electrons. Then, if the electrons are all displaced and the resultant electric field and restoring forces arising from it are considered,[3] oscillations of electrons in the plasma are produced with an angular frequency:

$$\omega_p = \left(\frac{ne^2}{m\varepsilon_o}\right)^{1/2}$$

Where: ω_p is the angular frequency in radians per second of the oscillatory motion and is called the *plasma frequency*, and n is the electron density, e is the electron charge and ε_o is the permittivity of free space. It turns out that the propagation of electromagnetic waves in a plasma is very much a function of ω_c and ω_p, which in turn are functions of the magnetic field $\bar{B}$ and the electron density n, respectively. Up to this point, no account has been taken of the fact that the plasma is a gas whose particles are always colliding. For less than 1-percent ionization, it can be shown that the collisions can be taken account of in a simple theory by considering only the frequency of collisions between electrons and neutral atoms, this frequency being noted as ν.

This is a very brief explanation of how the three principal parameters of microwave propagation in a plasma arise. A more complete derivation of the problem of microwave propagation in a plasma would start with Maxwell's equations and the Boltzmann equation and may be found in the literature.[9] For our purposes, in terms of the parameters discussed above, the treatment can be much simplified.[10]

Consider a uniform plasma, very large in extent, with a uniform magnetic field B_o everywhere. Then, microwaves in this plasma must propagate according to 1) Maxwell's equations:

$$\nabla \times \bar{E} + \frac{\partial \bar{B}}{\partial t} = 0$$

$$\nabla \times \bar{H} - \left(\frac{\partial \bar{D}}{\partial t} + \bar{J}\right) = 0$$

2) The equation of continuity:

$$\frac{\partial en}{\partial t} + \nabla \cdot \bar{J} = 0$$

and 3) the Lorentz force equation:

$$nm\dot{\bar{v}} + nm\nu\bar{v} = ne(\bar{E} + \bar{v} \times \bar{B}_o)$$

Where $\bar{H}$ is the magnetic field intensity, $\bar{D}$ is the electric displacement, and $\bar{J}$ is the current density. The term $\bar{B}_o$ is an external field, constant everywhere. First consider the case where $\bar{B}_o = 0$, (no external magnetic or electric fields). Solving the above system of equations, propagation in a plasma is like propagation in a lossy dielectric with a relative dielectric constant:

$$\varepsilon = \varepsilon' + j\varepsilon''$$

$$= \left[1 - \frac{\omega_p^2}{\nu^2 + \omega^2} - j\frac{\nu}{\omega}\left(\frac{\omega_p^2}{\nu^2 + \omega^2}\right)\right]$$

Where: $\omega = 2\pi \times$ (frequency of microwaves), and ε'' is the term that produces attenuation. If the collisions are neglected ($\nu = 0$), then:

$$\varepsilon = \varepsilon' = 1 - \frac{\omega_p^2}{\omega^2}$$

In optical terms, the index of refraction η is given by:

$$\eta = \sqrt{\varepsilon} \cong \sqrt{1 - \frac{\omega_p^2}{\omega^2}}$$

And, the optical path R is given by:

$$R = \int \eta\, ds$$

$$= \int \left(1 - \frac{\omega_p^2}{\omega^2}\right)^{1/2} ds$$

For ω less than ω_p, then η is imaginary and there is reflection. Note that ε' will be less than unity. For ω greater than ω_p, then ε' is greater than zero and the microwaves will propagate in the plasma (Fig. 4a).

The effects of ν are two-fold. There will be loss (attenuation) at all frequencies. When ω is less than ω_p, there will be some penetration into the plasma by the microwaves at all frequencies, and cutoff is not so sharp (Fig. 4b).

If the case where external magnetic field $\bar{B}_o$ is applied is considered, some complications arise. The equivalent dielectric is now a tensor:

$$\varepsilon = \begin{bmatrix} \varepsilon_{11} & j\varepsilon_{12} & 0 \\ -j\varepsilon_{12} & \varepsilon_{11} & 0 \\ 0 & 0 & \varepsilon_{33} \end{bmatrix}$$

Here, $\bar{B}_o$ is parallel to the 3-axis. It now turns out that the polarization of the microwaves has a distinct effect on their propagation. These effects can be classified by considering the cases[9] where propagation is parallel to the external magnetic field or perpendicular to it. In the first case, where microwaves are propagated parallel to the magnetic field, it can be shown that the plasma behaves like a high-pass filter to left-hand circularly polarized waves, where the cut-off frequency is a function of ω_c, ω_p, and ν. For a right-hand circularly polarized wave the plasma behaves like a band stop filter, with a stop-band between ω_c and the upper cut-off frequency

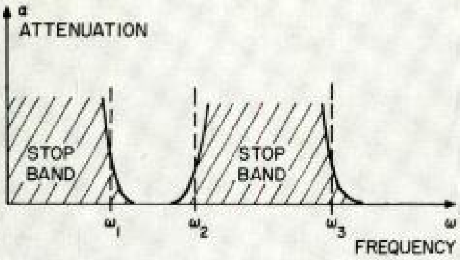


Fig. 6—Propagation for $\bar{E}$ perpendicular to $\bar{B}$.

which is a function of ω_c, ω_p, and ν (Fig. 5).

When propagation is perpendicular to B_o, then the electric field vector will have a component in the same plane as $\overline{B}_o$. Then if the electric field vector $\overline{E}$ is resolved into its components $E_{//}$, parallel to $\overline{B}_o$ and $E_{\perp}$, perpendicular to $\overline{B}_o$, the following applies: The component $E_{//}$ is unaffected by the presence of the magnetic field B_o. This is logical, since any velocity imparted to particles by $E_{//}$ will be $v_{//}$ and:

$$\overline{v}_{//} \times \overline{B}_o = v_{//} B_o \sin\theta = 0$$

That is, B_o has no effect. On the other hand, $E_{\perp}$ is strongly affected. Now propagation is possible in a passband and at higher frequencies (see Fig. 6) and again ω_1, ω_2, ω_3 are functions of ω_c, ω_p, and ν. In all cases, ν acts like a lossy element in any filter; as $\nu/\omega_p \rightarrow 0$, cutoff is sharper; as ν/ω_p becomes larger, attenuation rises in passbands and cutoff is less marked.

For the $E_{\perp}$ case, for the collisionless case ($\nu/\omega_p = 0$), the equivalent relative dielectric constant takes on reasonably simple forms in the passband, i.e.:

$$\varepsilon = \varepsilon' = 1 - \frac{\left(\frac{\omega_p}{\omega}\right)^2 (\omega_p^2 - \omega^2)}{\omega_c^2 + \omega_p^2 - \omega^2}$$

When the plasma is not uniform, the dielectric constant is a function of position and the propagation properties become quite complex. In some cases solutions are known for given variations of electron density in space. Maxwell's equations and the wave equation must be satisfied in terms of the microscopic dielectric constant for the assumed variation of electron density and collision frequency. It would be difficult to make general statements concerning propagation under these conditions, but some comments can be made. When the variation of electron density is such that the critical density is exceeded for some layer normal to the direction of propagation, it does not follow that total reflection occurs. The energy will be partially reflected but there will be evidence of penetration. This is the case of a more or less gradual transition into the critical region and under these conditions, the mismatch is less severe, resulting in less than total reflection and a definite depth of penetration. In this discussion, it is assumed that the variation in dielectric constant is over a distance comparable to a wavelength. Variations that occur gradually over many wavelengths can be analyzed using the techniques developed for study of the ionosphere in which the layers or strata may be considered as uniform layers of constant electron density. Microwave optical techniques may then be used to analyze the propagation effects.

PLASMA RE-ENTRY EFFECTS

Consider the effects of the plasma sheath and wake created by a body re-entering the earth's atmosphere. The sheath is a layer of plasma that surrounds the body. The electron density in the sheath may be high enough to prevent penetration by microwave signals and so disrupt communication with the ground. The wake is a plasma trail following the body which causes distortion of the radar return pulse so that interpretation is difficult.

The DAMP projects area has devoted considerable effort to the analysis and interpretation of the radar return from the re-entry vehicle and its wake. Analysis of the reflection from an inhomogeneous plasma was required and since very little work had been done in this field a laboratory measurement program was instituted. The purpose of the program was to investigate the scattering properties of various gradients in electron density for fairly simple models.

A first approximation to the problem was a plasma column with a variation or gradient in electron density along the axial dimension. Experimentally, the column could have been studied by means of various microwave diagnostic techniques, using either a pair of horns in one branch of an interferometer, or placing the column in waveguide. The horn technique was discarded as it required rather large lengths of column at reasonable measurement frequencies. Small gradients in electron density would not be seen by that method. However, placing the column in reduced height waveguide, with its axis in the E-plane for TE_{10} propagation mode, would permit determination of average electron density over a very small length of column. Taking a large number of measurements along the column would permit plotting the gradient along a glow discharge tube. As a result of these considerations the column-in-waveguide was chosen for the experimental model.

THE COLUMN-IN-WAVEGUIDE EXPERIMENT

A microwave interferometer was constructed at X-band using direct-reading variable attenuators and phase shifters. A special low-height waveguide section was constructed whose height was reduced to 0.05 inch. This permitted measurement of the average value of electron density along the column at small intervals in order to construct the gradient. However, the half-inch-diameter column partially filled the waveguide, and the measured values of phase shift and attenuation produced by the plasma had to be translated into the equivalent plasma parameters. An expression had to be derived that related the dielectric constant of a given dimension of column in waveguide to the attenuation and phase shift produced. (The attenuation as given by the interferometer is deceptive in that it just gives the transmission loss and thus is not in general an indication of absorption but rather of reflection for this case).

From the work of Marcuvitz[11] a column of dielectric mounted in waveguide in the E-plane may be described in terms of the equivalent "tee" of Fig. 7. The scattering matrix for this tee is given as:

$$S = \begin{bmatrix} S_{11} & S_{12} \\ S_{21} & S_{22} \end{bmatrix}$$

The matrix is symmetric ($S_{12} = S_{21}$) for the matched case, and S_{12} is the desired transmission coefficient:

$$S_{12} = \frac{2(A - B)}{2A\left(1 + 2B + \frac{B}{A}\right) - 1}$$

Where: $A = [(Z_a/Z_o) + (Z_b/2Z_o)]$ and $B = (Z_b/2Z_o)$.

Graphs were plotted that permitted direct indication of the equivalent dielectric constant from the measured transmission coefficient. Fig. 8 shows a typical plot for a half-inch-diameter plasma column. The equivalent dielectric constant is given as:

$$\varepsilon = 1 - \frac{\omega_p^2}{\omega^2 + \nu^2}$$

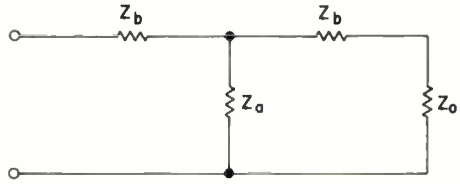


Fig. 7—Equivalent-tee configuration for column of dielectric mounted in waveguide in E-plane.

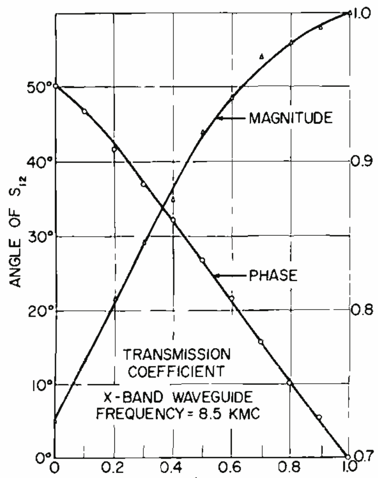


Fig. 8—Typical plot of transmission coefficient for half-inch diameter plasma column.

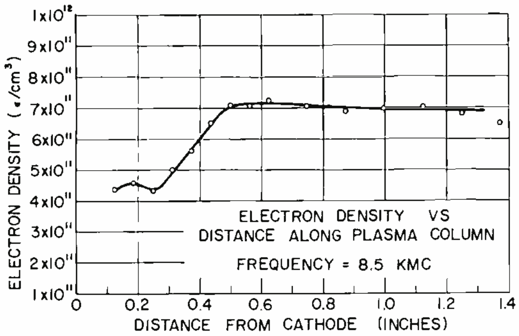


Fig. 9—Curve showing electron density along plasma column versus the distance from cathode.

Since the values of ω are known, measurements should be taken at two frequencies in order to specify electron density and collision frequency. The collision frequency was negligible for the gas pressures used so that practically speaking only one measurement frequency was required. However, the two frequency methods proved to be a means of estimating measurement error. A typical plot of electron density along the column is shown in Fig. 9. A definite gradient was found in a dark space region near the cathode end.

The expressions for the scattering coefficients enabled estimates to be made of the effect of a gradient on scattered

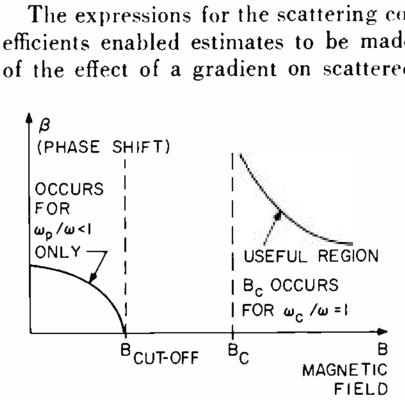


Fig. 10—Effect of controlling phase shift by varying external magnetic field.

microwave energy. As the terminal current of the plasma column was varied, smooth variation of phase shift was produced. The argon tube used permitted a continuous variation of phase shift of 25° at x-band (10 Gc). This could be easily extended by a serpentine column arrangement that would permit many columns to be inserted along a waveguide section.

MICROWAVE PLASMA DEVICES

From the study of microwave propagation in plasmas it was seen that even a simplified viewpoint gives rise to a complicated picture of propagation, especially in the presence of a magnetic field. However, the picture can be given structure and form by keeping in mind the sort of dependence that arises in terms of plasma frequency, ω_p, cyclotron frequency, ω_c, and collision frequency, ν, and the simplified picture of how these themselves arise as described earlier. It was seen that cut-off frequencies, i.e., attenuation by reflection are a function of ω_p and ω_c, and attenuation by absorption a function of ν, principally.

Examining the expressions for dielectric constant, it is seen that they are a function of frequency. This means that plasma is a dispersive medium. Further, the dependence of the dielectric constant ϵ on ω_p, ω_c, and ν means that the discharge current, external magnetic field, and pressure all affect ϵ and hence one can control ϵ by appropriate variation of these parameters. For example, one might make a simple phase shifter by putting a discharge tube in a waveguide and varying the discharge current. That is, the experiment of the preceding section was, among other things, a crude phase shifter.

When microwaves are propagated in a bounded medium, such as a waveguide or coaxial cable which is filled or partially filled with plasma, then in addition to everything else, the microwaves have to obey the particular by-laws, more formally called boundary conditions, of the physical environment. It turns out that for certain cases there are different phase velocities for left- and right-hand circularly polarized waves in a waveguide.[12,13] Since a linearly polarized wave can be resolved into a left- and right-hand circularly polarized wave, then the effect of transmitting a linearly polarized wave in a circular waveguide partly filled with plasma, with an axial magnetic field, will be to rotate the plane of polarization, for the correct choices of ω_c/ω_p, ω_p/ω, and ω_c/ω. This is the Faraday effect, the same gyromagnetic effect as is used in certain ferrite isolators, and can be applied to make plasma isolators.

Again consider a coaxial cable, with the space between the inner and outer conductors filled with plasma and an axial magnetic field. It has been demonstrated experimentally,[14] as well as theoretically, that a phase shifter of this configuration is possible, and that phase shift may be controlled over significant ranges as a function of ω_c/ω, i.e., by varying the external magnetic field (see Fig. 10). Still another approach to phase shifting using plasmas may be found in the case of the partially filled waveguide.[15] Here, analysis and experiment have shown the existence of waves of slow phase velocities which can be exploited in phase shifting.

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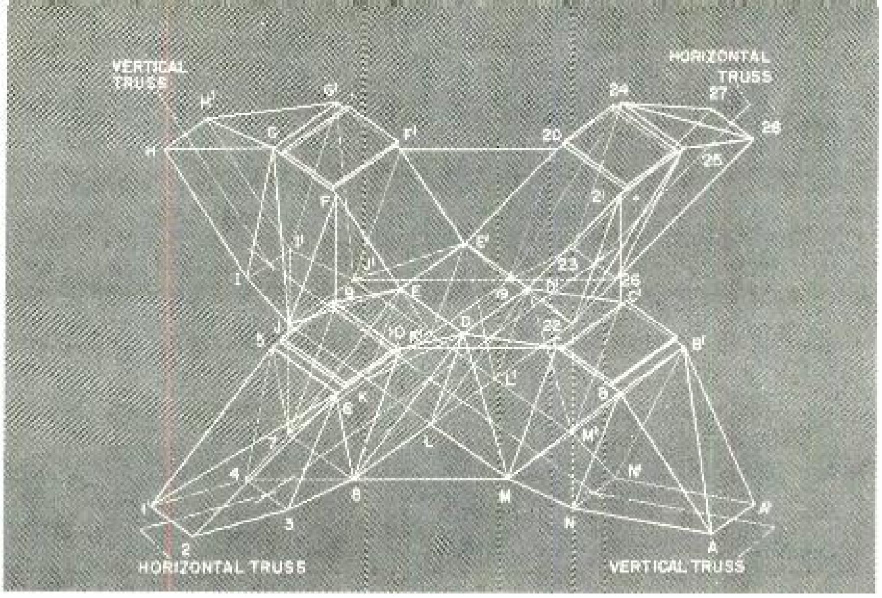

Fig. 1—Vector diagram showing how horizontal and vertical trusses are joined by interconnecting members.

ANALYSIS OF A TRI-DIMENSIONAL FEEDHORN SUPPORT STRUCTURE

A correlation coefficient, developed from experimental stress analysis and performed on a full-scale model under partial loading in the laboratory, was used to predict the performance of the mechanical structure under actual loading. The design and test analysis of a space structure involving results from over 100 tests are described.

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A TRACKING RADAR transmits and receives with an antenna having a highly directional pattern. The radar beam is compressed in both height and width, resembling a searchlight beam. A radiation pattern of this sort is obtained from a reflector whose surface is a paraboloid of revolution. The reflector feed system is located at its focal point; when the reflector focal distance is very long, the mechanical structure supporting the feed system, or simply speaking, the feedhorn becomes very complicated.

BORESIGHT ERROR

A factor called *boresight error* determines the final accuracy of the radar system. Precise location of the target requires a zero angle between the axis of the reflector and the axis of the feedhorn. Any deviation from this condition outside the tolerable limits will cause an error in locating the range or direction of the target. The reflector, feedhorn structure, and the supporting legs are structural elements subject to deflections due to their own weight as well as any mass they are designed to support. The reflector deflection at the point of attachment of the supporting legs causes a rotation of the feedhorn face from its zero position. This deviation can be reduced to a minimum by the proper structural design of the supporting legs and of the feedhorn structure.

MECHANICAL DESIGN APPROACH

Since boresight error has such a decisive effect on performance, the selection and numerical value of the design parameters and numerical values for the feedhorn support structure must be carefully chosen. Design of the tri-dimensional feedhorn support structure is based on the following parameters:

1) Maximum weight of support, 650 pounds.
2) Safety factor of dynamic load at least 1.5.
3) Angular deflection of the feedhorn axis with respect to vertex of reflector within 0.020 mils.

Based on previous experience, the dynamic load was assumed to be 10 times the static load. This figure represents the combined upper limit for all the loads caused by acceleration, deceleration and buffering of the antenna reflector and its components. With this factor of safety and the parametic specifications as design goals, the feedhorn support was developed as a statically indeterminate space structure consisting of two tri-dimensional trusses: a vertical truss and a horizontal truss intersecting at right angles. Both horizontal and the vertical trusses consist of two parallel-plane trusses, joined together by interconnecting members (Fig. 1). Structural members consist of solid and hollow bars having rectangular cross sections and flat plates; the complete structure is fabricated from aluminum 6061 and has over-all dimensions of 12 by 12 feet.

The feedhorn support is rigidly connected to the reflector by four legs pinned at the reflector; thus the reflector, the legs, and the feedhorn support structure form a single rigid structural system.

The load analysis of an indeterminate structure requires two solutions: a *determinate*, and an *indeterminate*. This approach evaluates forces in redundant members, and predicts the effect on the other members. However, the combinations of bars and plates required in the feedhorn support represent a very difficult problem to solve in determining loads in the members and deflections at critical points. The indeterminate solution is of little help, since certain basic assumptions must be introduced. Some of these assumptions are:

1) *Construction of joints and use of gussets.* Gussets at welded joints are used throughout as a fabrication technique. Since high bending stresses were expected in many members, the differences between gusseted and pinned (i.e., calculated) joints were given serious consideration.
2) *The effect of the plates in a vertical truss.* The method of analyzing the joints could not make allowances for the effect of the plates on an over-all load distribution.
3) *The stress distribution in the horizontal truss.* The vertical truss is assumed to carry all the applied and reactive load; this assumption needs confirmation which an analytical stress analysis could not provide.

Since none of the analytical methods give a reasonable guarantee that the design approach satisfies all design specifications, the experimental stress analysis along with the determinate solution seemed to be the best method. Thus the strain gage technique not only replaced the indeterminate solution, but also could provide a source of information and a guide for future design of this type of structures.

To use the determinate solution properly with the strain-gage technique, a *correlation coefficient* was needed to relate experimental and analytical data; this was imperative since the system of forces acting on the feedhorn support could not be reproduced in the laboratory. The function of the correlation coefficient is defined as follows: *When the* *correlation coefficient for a structural member (obtained from a laboratory test) is independent from the magnitude of the applied load, it can be used to predict the performance of the structure under an actual set of loads.* This performance result is obtained by multiplying the load in each member (calculated from a determinate solution) by the correlation coefficient for that member. The concept of the correlation coefficient is developed in next paragraph. The order used during the analysis of the feedhorn support design was as follows:

1) A set of loads representing the resultants of all the forces acting at the two interfaces between the vertical pair of the supporting legs and the feedhorn was applied; the load distribution was then found in every member of the vertical structure by the tri-dimensional method of joints. These loads were: 3,580 pounds axial force; 200 pounds shear force; and 160,000 inch-pounds moment. Plates were not taken into consideration.
2) Cross sections of loaded members were determined using a safety factor of at least 2 (based on a yield point of Al 6061 T6) for loaded members in a structure. Since most members consist of extruded hollow bars, three cross-sections were used: 1.7 in^2 for the most loaded members; 1.2 in^2 for the medium-loaded members; and, 1.00 in^2 for the least-loaded members. For redundant members whose loads were unknown, 1.2-in^2 cross-sections were assigned.
3) The cross sections in the corresponding members of the horizontal truss were the same as in the vertical.
4) The total structure weight was computed with allowances for weldments and gussets. Since structure weight fell within limits prescribed by design specifications, no changes in the design were necessary.
5) Using Castigliano formula, the angular deflection of the feedhorn face was computed by assuming the feedhorn to be a rigid body whose angular deflection (under a set of loads) is directly dependent upon the rotation of structural members to which it is connected.

Table I lists some of the important members of the vertical structure along with their loads and stress level under dynamic loading.

Using the described method, the angular deflection of the feedhorn axis was computed to be 0.026 mils. However, the specification permitted a maximum deflection of 0.020 mils; it will be shown later that the actual deflection as a result of laboratory tests amounted to 0.014 mils.

DERIVATION OF CORRELATION COEFFICIENT

Since the reproduction of actual loading was impractical under laboratory conditions, an applied load similar to the actual was selected. The load was a variable moment applied to the mounting sections of the vertical truss. The concept of the correlation coefficient was based on the following principle:

The axial force F_{ij} in a strained member $_{ij}$ can be expressed as a function of the applied load F_o (if all applied loads are the same) and a distribution factor α_{ij}:

$$F_{ij} = \alpha_{ij} F_o \qquad (1)$$

But, since the applied load F_o is caused by the moment M_o applied to two mounting frames, then:

$$F_o = \frac{M_o}{2L_o} \qquad (2)$$

Where: L_o is a distance between pads in the mounting frame. Substituting Eq. 2 into Eq. 1:

$$F_{ij} = \alpha_{ij}\frac{M_o}{2L_o} \qquad (3)$$

From Hooke's Law, we have:

$$\epsilon_{ij} = \frac{\sigma_{ij}}{E} = \frac{F_{ij}}{EA_{ij}}$$

Where: A_{ij} is a cross section of the $_{ij}$ member, from which:

$$F_{ij} = \epsilon_{ij} EA_{ij} \qquad (4)$$

Substituting Eq. 4 into Eq. 3:

$$\epsilon_{ij} EA_{ij} = \alpha_{ij}\frac{M_o}{2L_o} \qquad (5)$$

Substituting 31 inches $= L_o$, Eq. 5 can be shown in final form as:

$$\epsilon_{ij} = \frac{\alpha_{ij} \times 161 \times 10^{-4} M_o}{EA_{ij}} \qquad (6)$$

In this way, the strain in the $_{ij}$ member of the truss can be expressed as a function of applied moment. Consequently, the theoretical stress analysis for the laboratory loading must precede any

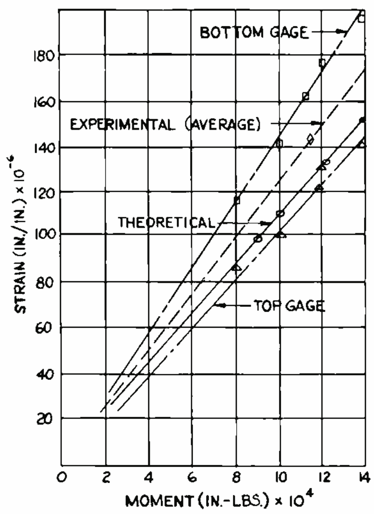

Fig. 2—Graph showing a typical representation of strain in loaded member (BN of Fig. 1).

laboratory measurements in order to derive ϵ_{ij} for Eq. 6.

The ratio between experimental strain ϵ_{0ij} read from instrumentation and ϵ_{ij} from Eq. 6 constitutes a correlation coefficient,

$$\frac{\epsilon_{0ij}}{\epsilon_{ij}} = f_{ij} \qquad (7)$$

Since the correlation coefficient f_{ij} was found constant for the whole range

TABLE I — Loads and Stress Levels

Member	*Cross Section In.^{2}*	*Force Lbs.*	*Static Stress Lbs./In.^{2}*	*Dynamic Stress Lbs./In.^{2}*
AB	1.7	+1,950	+1,150	+11,500
AB'	1.2	0000	000	000
AA'	1.0	+ 214	+ 214	+ 2,140
AN	1.7	+ 265	+ 150	+ 1,500
NN	1.0	− 508	− 508	− 5,080
BN	1.7	−3,070	−1,805	−18,050
BM	1.2	+1,678	+1,400	+14,000
CM	1.2	− 811	− 675	− 6,750
DC	1.2	+1,132	+ 945	+ 9,450
DM	1.2	− 852	− 710	− 7,100
DL	1.2	0000	000	000
DE	1.2	− 368	− 305	− 3,050

Table II — Correlation Coefficients and Actual Stress Levels

Member	*Cross Section*	*Correlation Coefficient*	*Adjusted Force Lbs.*	*Stress Lbs./In.^{2} Adjusted Dyn.*
AB	1.7	0.848	+1,650	+ 9,700
AB'	1.2	1.000	0000	000
AA'	1.0	6.397	+1,370	+13,700
AN	1.7	5.936	+1,570	+ 9,250
NN'	1.0	2.054	−1,040	−10,400
BN	1.7	1.126	−3,460	−20,300
BM	1.2	1.072	+1,800	+15,000
CM	1.2	0.712	− 576	− 4,800
DC	1.2	0.642	+ 728	+ 6,060
DM	1.2	2.000	−1,700	−14,200
DL	1.2	1.000	0000	0000
DE	1.2	1.945	− 715	− 5,950

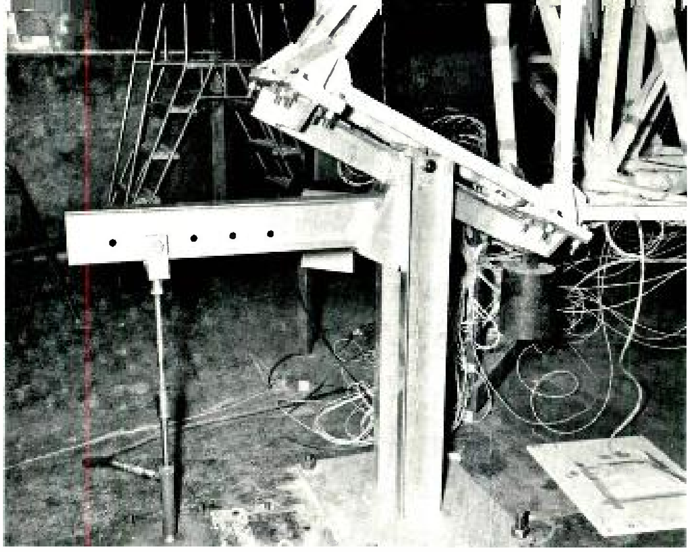

Fig. 3—Hydraulically operated fixture used for variable moment application.

of applied loads, it was used to predict the stress level in members of the truss under field conditions for a given system of applied forces.

Multiplying the values for a calculated load (see Table I) by the correlation coefficient gives the actual stress level; these results are shown in Table II.

LABORATORY TEST

The experimental stress analysis (strain-gage technique) was performed on the full-size structure. The test program comprised about 150 tests during which over 5000 readings were taken. Hydraulically operated fixtures were used as the loading equipment. Four moments ranging from 80,000 in-lb to 140,000 in-lb were applied to the vertical structure. Fig. 2 shows test data plotted and analyzed for each member of the structure, and used for the computation of the correlation coefficient in the structure.

The next objective of the test was to determine the deflection in the feedhorn face. Two mirrors and autocollimetors were used; the mirrors were attached to the feedhorn face and to a reference bar. Angular deflections of the mirrors under a specified load yielded data used for the computation of angular deflection of the feedhorn axis with the respect to vertex of reflector.

CONCLUSIONS

Test results show that the determinate solution along with the strain gage technique can be successfully used in the analysis of statically indeterminate space structures and in the prediction of their behavior under actual load conditions. The main advantage of this method is that the correlation coefficient can be derived from the system of forces used for the test which is different from the actual system of forces; however, the correlation coefficient must be constant for the whole range of forces used for the test. This method is also desirable from an economical standpoint, since the design and fabrication of loading equipment to reproduce actual forces would run into prohibitive expense.

The test results which provided the answers for earlier assumptions can be listed as follows:

1) *Low stress level in the horizontal truss.* It was assumed that the horizontal truss does not carry any forces. The presence of low value forces can be attributed mainly to redundant members which connect transition sections of the vertical truss to the horizontal ones.
2) *Plates had very little effect on distribution of axial forces in the members of the vertical truss.* Plates were not taken into account in calculations; their negligible effect confirms the correctness of these assumptions in the preliminary analytical work.

Derivation of the correlation coefficient for the members of the vertical truss was one of the design goals. This coefficient ranged from 0.6 to 2.00 except for some members located at the interface, where a correlation coefficient as high as 6.3 was noted. As shown in Table II, the correlation coefficient for the highly loaded members like *BN* was 1.126. The corrected stress level for dynamic load gave a safety factor of better than 1.5 (based on the yield point of Al 6061-T6).

The over-all stress level in members of the vertical structure due to axial forces was in complete agreement with design specifications; therefore, no corrections or replacements in the structure were necessary. Generally, it can be assumed that the effect of welded-gusseted joints was the main reason for a correlation coefficient other than unity.

The high correlation factor for members located at the interface was investigated since it was felt that the deflection of the loading fixture might offset the axial force of the low magnitude. The calculations performed on the basis of this assumption came very close to experimental results.

The redundant members have shown a very low stress level. The exact figure of their axial force in the actual conditions can be found simply by applying the equation of statics to the joints where they belong.

The angular deflection of the feedhorn was investigated as another design objective using the method described in the previous paragraphs. It was found that the actual magnitude of deflection of the feedhorn axis was 0.014, instead of 0.026 mils, as obtained through purely analytical means. Therefore, the effect of the welded joints and gussets proved to be advantageous in bringing the third design parameter (angular deflection) down to specified limits.

ACKNOWLEDGMENT

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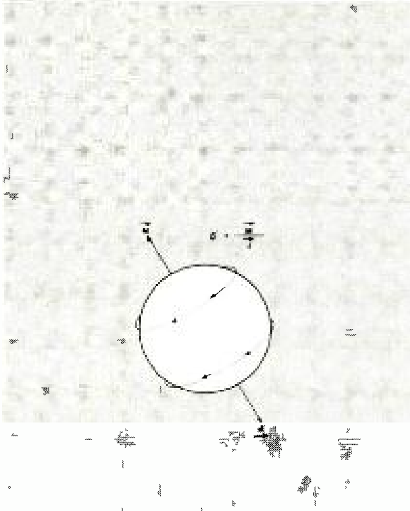

Fig. 1—A spinning electron with magnetic and angular moments.

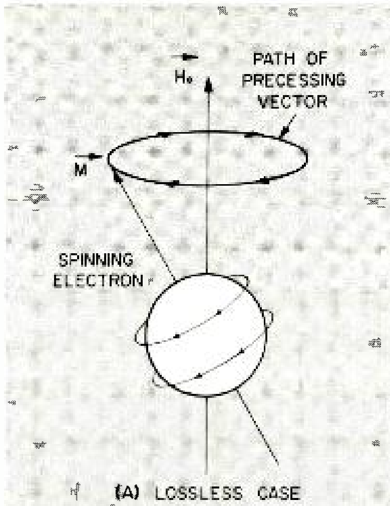

Fig. 2a—Electron motion in DC magnetic field—lossless case.

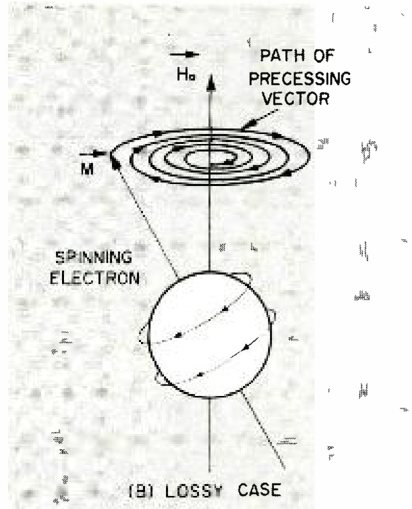

Fig. 2b—Electron motion in DC magnetic field—lossy case.

FERRITES AND THEIR APPLICATION TO MICROWAVE DEVICES

During the past decade, microwave ferrite devices have evolved from laboratory curiosities to major components in almost all radar systems. Growth in the use of these devices progressed as the understanding of propagation in ferromagnetic materials became more apparent. This paper reviews properties of ferrite materials, develops their microwave permeability relationships, and relates permeability to the behavior of the more popular microwave ferrite devices. A selected bibliography for further basic reading in this field is included.

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MANY FERROMAGNETIC materials are found in electronic equipment—the most common being the soft iron used in most low-frequency transformers or inductors. Such soft iron is characterized by high permeability and low resistivity, and from Maxwell's equations we may deduce that it is unsuitable for use at high frequencies because of its low resistivity. Measurements bear out such predictions and prove that good conductors make good reflectors. Ferrite materials, on the other hand, have high permeability and high resitivity, of the order of 10^{12} times that of soft iron—because ferrite materials are compounds rather than basic elements. The electrons responsible for the high magnetic moments associated with ferrites are not in the conduction band as they are in soft iron materials, but rather at some lower energy level.

The basic formula of a ferrite is $MeO \cdot Fe_2O_3$, where Me is a divalent metallic ion. Magnesium, manganese, and nickel are most commonly used for the metal ion, while copper, cobalt, zinc, and cadmium have also been studied. The crystalline structure of ferrite called the *spinel structure*, is characterized by cubic symmetry. The unique character of ferromagnetic materials, whether ferrite, soft iron, cobalt, or others, is the high degree of coupling or exchange forces that exist between the magnetic moments of neighboring electrons. Thermal energy tends to decouple or reduce these exchange forces; as a result, properties of ferromagnetic materials are temperature-sensitive. The temperature at which neighboring electron spins are completely decoupled is known as the *Curie temperature*. Above this temperature, the material loses its ferromagnetic properties and becomes a paramagnetic material. Typical Curie temperatures of microwave ferrites range from 100 to 300° C.

FERROMAGNETIC RESONANCE

Electron motion, a combination of revolution and spin, determines the magnitude of the magnetic moment of ferrites. Spin motion is the major contributor to the magnetic moment, and the revolution about the center of the atom may be neglected for most applications. Since the electron has mass, an angular momentum is associated with its spin motion (Fig. 1). The ratio of its magnetic moment to its angular momentum is known as the gyromagnetic ratio, γ, and has a nominal value of 2.8 Mc/oersted.

To examine the effect of externally applied magnetic fields on the motion of the spinning electron, consider first the application of a DC magnetic field to a ferrite material. There will be a torque acting on all the electrons in the material. If the DC field is given by $\vec{H}_o$, the torque on each electron is the cross product of its magnetic moment (M) and the applied DC field. From Newton's second law, torque is the time rate of change of angular momentum ($\vec{J}$). Therefore:

$$\text{Torque} = \frac{d\vec{J}}{dt} = \vec{M} \times \vec{H}_o \qquad (1)$$

Since the gyromagnetic ratio γ is known, Equation 1 may be expressed entirely in terms of magnetic moments and applied field:

$$\frac{d\vec{J}}{dt} = \frac{1}{\gamma}\frac{d\vec{M}}{dt} = \vec{M} \times \vec{H}_o \qquad (2a)$$

$$\frac{d\vec{M}}{dt} = (\vec{M} \times \vec{H}_o) \qquad (2b)$$

Solution of Equation 2b indicates many of the significant characteristics of microwave ferrite devices. It does not, however, account for observed losses. A more exact equation, relating applied fields to resultant magnetic moments and having a loss term, λ, was proposed by Landau and Lifshitz in 1935:

$$\frac{d\vec{M}}{dt} = \gamma(\vec{M} \times \vec{H}) - \lambda\left[\frac{(\vec{H} \cdot \vec{M})\vec{M}}{M^2} - \vec{H}\right] \qquad (3)$$

For simplicity, only Equation 2b will be solved herein, but throughout the discussion, ferrite behavior will include the loss mechanism. With the application of H_o, a spinning electron will precess about the axis of the applied field. In the absence of losses the electron would precess about the applied field indefinitely. Because of losses the electron precessional motion spirals into alignment with the applied field (Fig. 2). Next, consider the effect of applying an alternating magnetic field to the ferrite material in the plane perpendicular to the direction of the applied DC field. The total applied field is then given by:

$$\vec{H} = h_x i + h_y j + H_o k \qquad (4)$$

Lower-case letters represent AC components with ϵ^{jwt} time variation assumed. The resulting magnetic vector produced by the electron is:

$$\vec{M} = m_x i + m_y j + M_o k \qquad (5)$$

Next, $\vec{M}$ is determined as a function $\overleftarrow{H}$. Taking Equations 4 and 5 and performing the operations dictated by Equation 2b gives an equation which may be broken up into three parts, one for each of the three dimensions:

$$jwm_x = \gamma\,(H_o m_y - h_y M_o) \qquad (6a)$$
$$jwm_y = \gamma\,(h_x M_o - H_o m_x) \qquad (6b)$$
$$O = \gamma\,(h_y m_x - h_x m_y) \qquad (6c)$$

Solving for m_x and m_y, the RF magnetizations, as functions of the applied magnetic field gives:

$$m_x = \left(\frac{\gamma^2 H_o M_o}{\gamma^2 H_o^2 - w^2}\right) h_x - j\,\frac{\gamma w M_o}{\gamma^2 H_o^2 - w^2}\, h_y \qquad 7(a)$$

$$m_y = j\left(\frac{\gamma w M_o}{\gamma^2 H_o^2 - w^2}\right) h_x + \left(\frac{\gamma^2 H_o M_o}{\gamma^2 H_o^2 - w^2}\right) h_y \qquad 7(b)$$

The true character and unique properties of ferrite behavior at microwave frequencies may be obtained from careful consideration and study of Equations 7a and 7b. The most striking thing about these two expressions is that a magnetization or flux density vector may be produced in a given direction not only by an applied field in that direction, but by an applied field perpendicular to this given direction. This is indicated in Equation 7a by the presence of h_y in producing m_x, and in Equation 7b by the presence of h_x in producing m_y. It is also interesting to note that the component of magnetization produced by the perpendicularly applied field is in time quadrature with respect to the applied field, this is indicated by j. Another point of great significance is that the sign of the h_y factor in the m_x expression is negative whereas the sign of the h_x factor in the m_y expression is positive. Extension of this observation indicates that propagation in ferrite media may be *nonreciprocal*. Finally, note that the denominator of all of the coefficients in both Equations 7a and 7b indicate a resonance condition will exist when $\gamma H_o = w$. This implies that a DC field may be chosen for a given frequency so that very large values for these coefficients may occur.

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These coefficients are susceptability terms, since they relate applied field to resulting magnetizations. It is often convenient to express the susceptability of a ferrite material in the form of a matrix:

$$\overleftrightarrow{\chi} = \begin{bmatrix} \chi_{xx} & -j\chi_{xy} & 0 \\ j\chi_{xy} & \chi_{xx} & 0 \\ 0 & 0 & 0 \end{bmatrix} \qquad (8)$$

Because of the presence of the off-diagonal components ($j\chi_{xy}$ *and* $-j\chi_{xy}$) this expression is called a *tensor susceptability*. Relating Equation 8 to Equations 7a and 7b:

$$\chi_{xx} = \frac{\gamma^2 H_o M_o}{\gamma^2 H_o^2 - w^2} \qquad (9a)$$

$$\chi_{xy} = \frac{\gamma w M_o}{\gamma^2 H_o^2 - w^2} \qquad (9b)$$

For many applications, it is preferable to use permeability instead of susceptability. The transition from tensor susceptability to *tensor permeability* results in:

$$\overleftrightarrow{\mu} = \begin{bmatrix} \mu & -jK & o \\ jK & \mu & o \\ o & o & \mu_o \end{bmatrix} \qquad (10)$$

Where: $\mu = \mu_o\,(1 + \chi_{xx})$, and $K = \mu_o \chi_{xy}$. Fig. 3 is a pair of curves for μ and K as functions of H_o for a fixed frequency. Resonance occurs when H_o is at a value of w/γ.

Losses were neglected throughout the development of the tensor permeability. Fig. 4 is a curve of the diagonal component of the tensor permeability predicted by Equation 3. The dashed curve is the imaginary or loss component of the permeability. The low-field losses, not accounted for in Equation 3, are due to hysteresis; in order to avoid them, the material should be magnetized to a level above saturation.

The resonance loss term predicted by Equation 3 occurs from the coupling of energy to the precessing electrons. The exact resonance loss mechanism is fairly involved and is related to crystal imperfections, impurities, inhomogeneities, grain boundaries, etc. In order to minimize losses, the DC applied field should be adjusted for operation in either region 1 or region 2 (Fig. 4). As frequency is decreased, the resonance loss curve shifts to the left, thus reducing region 1. At some frequency, this region disappears and low-loss operation may be achieved only in region 2 or, above resonance. If frequency is still further reduced, the value of $\overleftrightarrow{\mu}$ may decrease to a level where the material becomes useless as a magnetic material. Thus, low

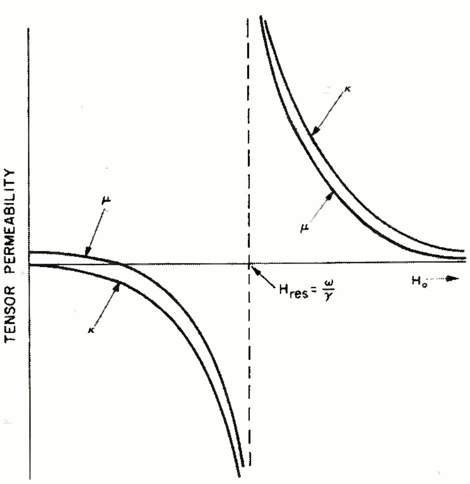

Fig. 3—Plot of μ and k versus applied field (lossless case).

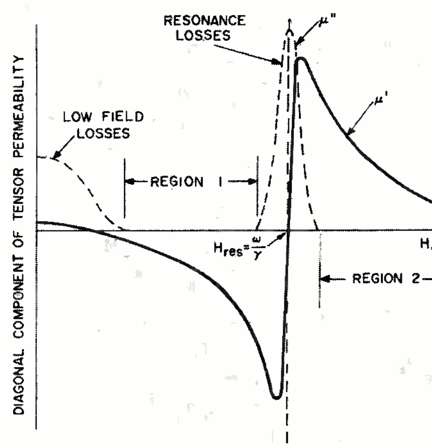

Fig. 4—Diagonal component of tensor permeability μ versus applied field (loss component included).

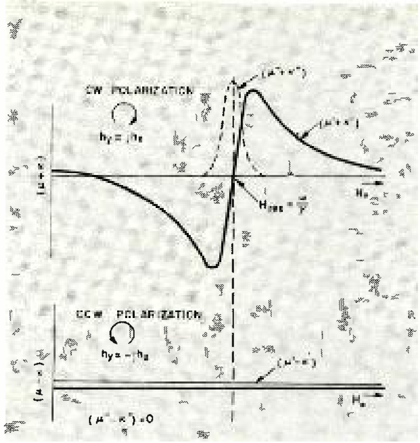

Fig. 5—Permeability of clockwise and counterclockwise polarized waves in ferrite media.

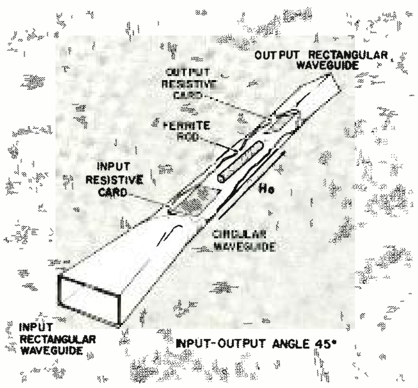

Fig. 6—Faraday rotation isolator.

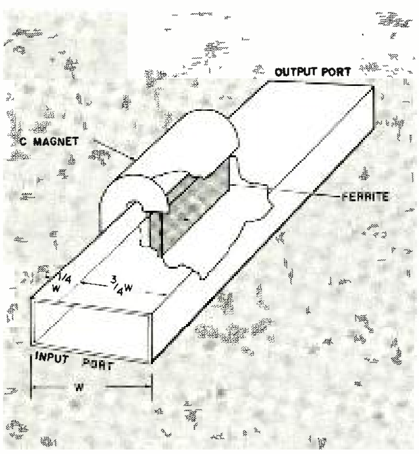

Fig. 7—Resonance absorption isolator.

field losses impose a lower frequency limit for the use of ferrite materials as a propagating medium. Materials have been developed for use at frequencies as low as 100 Mc.

The width of the resonance loss curve is referred to as *linewidth*, ΔH, and is given in oersteds. By definition, it is the 3-db width of the imaginary component of the tensor permeability. The factor λ of Equation 3 is related to the linewidth of a material by:

$$\lambda = \frac{\gamma M_o \Delta H}{2 H_{res}} \qquad (11)$$

As ΔH gets larger, maximum values of both μ' and μ'' decrease and their widths increase. When used in nonabsorption devices, the ferrite material generally becomes more lossy as linewidth increases.

PROPAGATION IN FERRITE MEDIA

To determine the propagation constant of electromagnetic waves in ferrite media, it is necessary to solve Maxwell's equations using the tensor permeability developed in the preceding section. This is both tedious and beyond the scope of this paper. Suffice it to say that the propagation constant is a function of μ and K (which are functions of the magnitude of H_o) and the angle between H_o and the direction of propagation. (A complete discussion of the development of the propagation constant for waves in ferrite media may be found on pages 97 through 103 of Reference 2.)

Before discussing the operation of microwave ferrite devices, one very significant phase of propagation in ferrite media should be considered: the effect of ferrite on circularly polarized waves or waves where $h_y = \pm jh_x$. That is, where the component of the magnetic field in one direction is equal in magnitude but 90° out of phase with the component at right angles to it.

The equation relating flux density and applied field is the familiar $\vec{b} = \overleftrightarrow{\mu}\,\vec{h}$ matrix. Where the material is ferrite, $\overleftrightarrow{\mu}$ is the tensor permeability defined by Equation 10. By expanding that matrix, the flux density in ferrite becomes:

$$b_x = \mu h_x - jKh_y \qquad (12a)$$

$$b_y = jKh_x + \mu h_y \qquad (12b)$$

$$b_z = \mu_o h_z \qquad (12c)$$

Consider first propagation of a clockwise polarized wave or where $h_y = j h_x$. Substituting this expression into Equations 12a and 12b results in:

$$b_x = \mu h_x + Kh_x = (\mu + K)\, h_x \qquad (13a)$$

$$b_y = Khy + \mu h_y = (\mu + K)\, h_y \qquad (13b)$$

Thus, the permeability for clockwise polarized waves propagating in ferrite media (in the direction of the DC bias field) is a scalar quantity of magnitude $(\mu + K)$.

When the same method is applied to counterclockwise polarized waves ($h_y = -j\, h_x$), the resulting permeability is again a scaler quantity of magnitude $(\mu - K)$. Fig. 5 plots both $(\mu + K)$ and $(\mu - K)$ vs. applied DC field. Both the real and imaginary components of $(\mu + K)$ and $(\mu - K)$ are shown, the imaginary component being that part of the permeability attributing to the losses of the material. (Low field losses are not shown.) From these curves, the operation of most rotating and resonance absorption ferrite devices may be explained without actually deriving propagation constants for these devices.

FERRITE ISOLATOR DEVICES

Faraday Rotation Isolator

The first ferrite isolator to receive widespread use is the Faraday *rotation isolator* (see Fig. 6). It consists of a rectangular waveguide input, a transition to circular guide, and a transition back to rectangular guide with the axes of the output guide at 45° to the input guide.

In the circular section of the device, two resistive absorption cards and a ferrite rod are positioned along the axis of the circular guide. One resistive card is placed close to the input rectangular guide with its surface normal to the E-fields of the input guide. Another resistive card (with its surface normal to E-fields of the output guide) is placed close to the output rectangular guide. Since the angle of the input-output guide is 45°, this same angle exists between the two resistive cards. A DC bias field is applied to the ferrite rod in the direction shown in Fig. 6. Since permeability for clockwise polarized waves is different from that of counterclockwise polarized waves, the direction of polarization of a linearly polarized wave (which consists of the sum of oppositely rotating circularly polarized waves) will rotate as it travels through a ferrite medium biased in the direction of propagation. This rotation occurs because the velocity of propagation of electromagnetic waves is inversely proportional to the square root of the permeability of the medium being traversed. Therefore, clockwise polarized waves travel through the ferrite at a velocity slower than that of counterclockwise waves. When the two waves leave the ferrite rod, they combine to form a linearly polarized wave with its direction of polarization at some angle other than the input angle. The magnitude of the DC field is adjusted so that the difference in $(\mu + K)$ and $(\mu - K)$ results in a rotation of 45° in the angle of polarization for the given length and diameter of the ferrite rod. The input resistive card offers zero attenuation to input waves since the card's lossy material is normal to the incident E fields. The output resistive card also offers zero attenuation since the angle of polarization has been rotated 45° by the ferrite so that the output E-fields are normal to the output resistive card. When, a reflection beyond the output port of the isolator occurs from a mismatch of some component further down the system, the reflected wave enters the output port of the isolator, passes the output resistive card with zero attenuation, is propagated through the ferrite rod. The wave thus rotates an additional

45° and is absorbed by the input resistive card, since the E fields of the reflected wave becomes parallel to the lossy material of the input card. Thus, we have a device capable of attenuating waves traveling in one direction yet offering no attenuation for waves traveling in the other direction.

Resonance Absorption Isolator

Resonance absorption isolators are by far the most common isolators on the market today. Their small size, power handling capacity, and relative ease of manufacture have led to widespread use throughout the microwave field. Although many variations of this type of isolator have been designed, the basic mode of operation is the same for all.

The device consists of a ferrite section positioned approximately one-fourth the way from the narrow wall of standard rectangular waveguide (Fig. 7). A C-type magnet is used to supply a transverse DC magnetic bias field to the ferrite. The primary questions related to the operation of this isolator are: 1) *Why is the ferrite placed where it is?* and 2) *What is the magnitude of the DC bias field?* The answer to the first question is related to the H-field configuration associated with the dominant (TE_{10}) mode in rectangular waveguide. In Fig. 8 the H-fields are transverse at the center of the guide and longitudinal at the side walls of the guide. As the fields propagate down the guide, at a position about a quarter of the way from the side wall, the direction of the H-fields is first transverse, then longitudinal, then transverse in the opposite direction and finally longitudinal in the opposite direction. In other words, the H-fields appear to be circularly polarized, rotating about an axis at this quarter point. The same phenomenon occurs at the three quarter position in waveguide but the circularly polarized waves rotate in the opposite sense. So, for waves propagating in the direction indicated in Fig. 8 at the one-quarter position, the H-fields appear to be counterclockwise circularly polarized; at the three-quarter position, the H-fields appear to be clockwise circularly polarized. For propagation in the opposite direction, the sense of polarization at these two positions reverses. If a ferrite section is placed at either of these positions and biased to resonance (see Fig. 5) by the C magnet, the device absorbs waves traveling in the other direction. This is exactly the type of unidirectional transmission required for isolator operation. Typical characteristics of this type of isolator are 30 db of isolation with 0.5-db insertion loss over a 10-percent bandwidth.

Other types of isolators have been built but none have achieved the widespread use of the resonance absorption isolator and the Faraday rotation isolator. (For detailed descriptions of some of these other isolators, see Reference 2.)

FERRITE CIRCULATOR DEVICES

Faraday Rotation Circulator

As in the case of the ferrite isolators, the first circulator to receive widespread use was the Faraday rotation circulator; construction and mode of operation are similar to its isolator predecessor with modifications made to the input and output ports. Fig. 9 is a sketch of this circulator which is a four-port device. The ferrite is magnetically biased to produce a 45° rotation, as in the isolator. Hence, waves entering port 1 pass through the ferrite section, rotate by 45° and leave the device at port 2. There is no coupling of energy from port 1 to either port 3 or port 4, since as the wave passes these ports, the E-field polarization is orthogonal to the propagating mode at both ports. Waves entering port 2, pass port 4 (for the same reason), rotate 45° on the way through the ferrite, and then exit via port 3—since the waves are now polarized for propagation to port 3. Energy does not propagate to port 1 because this port is now orthogonal to the propagating mode. The same procedure of analysis is used to show that waves entering port 3 will go to port 4 and waves entering port 4 will go to port 1. The characteristics of this type circulator are 25 db of isolation to uncoupled ports and 0.5 db of insertion loss to the coupled port with operation over a 10-percent bandwidth.

Differential Phase Shift Circulator

The differential phase shift circulator operates by the utilization of a non-reciprocal ferrite phase shifter in one arm of a balanced bridge system. Fig. 10 is a sketch of this type of circulator. The ferrite is placed at the plane of circularly polarized H-fields (Fig. 8). The magnitude of the magnetic field is adjusted so that the ferrite produces a phase shift of 180° more for waves traveling in one direction than in the other direction; in Fig. 5, this is accomplished by adjusting H_o to some value below or above resonance. With the ferrite biased to produce this differential phase shift of 180°, operation of the circulator is understood. Waves entering port 1 split in magnitude and propagate to the two co-linear arms of the first "magic" tee. As the wave in the right arm of the bridge goes through the ferrite section it is retarded by an angle 180° more than that of the wave traveling in the left arm of the bridge. When these waves meet at the second magic tee, they are therefore 180° out of phase resulting in propagation to port 2 which is the difference arm of the second magic tee. Waves entering port 2 split in magnitude and propagate out of the two co-linear arms of the second magic tee. Their phases, however, are 180° out of phase because of the nature of the magic tee. As the wave in the right arm passes the ferrite section, it is not retarded this time. When the waves meet at the first magic tee they are still 180° out of phase and therefore go out port 3, the difference arm of this magic tee. Further analysis will show that waves propagate from port 3 to port 4 and from port 4 to port 1. The characteristics of this circulator are similar to those of the Faraday rotation circulator but the power-

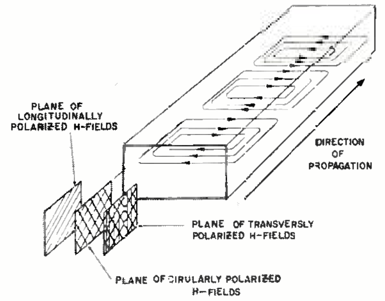

Fig. 8—H-field configuration of lowest order mode (TE_{10}) in rectangular waveguide.

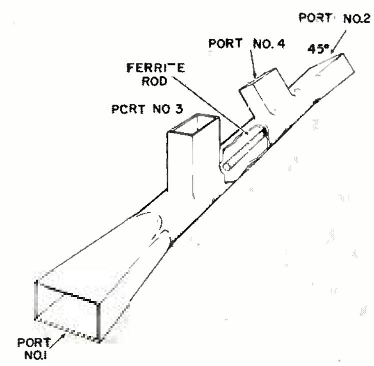

Fig. 9—Faraday rotation ferrite calculator.

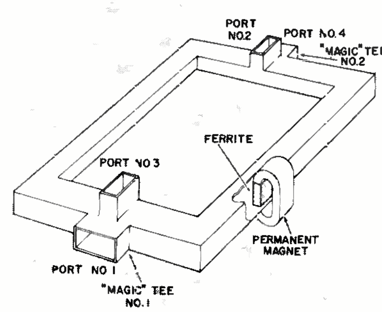

Fig. 10—Differential phase shift ferrite calculator.

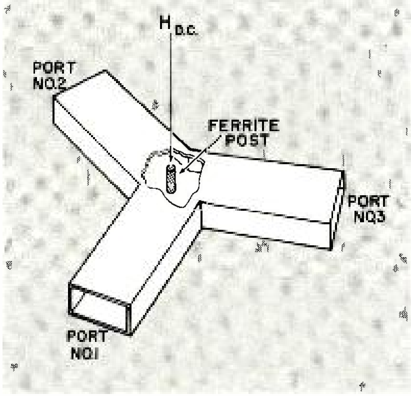

Fig. 11—Birefringent three-part ferrite circulator.

handling capability of the differential phase shift circulator is much higher.

Birefringent Ferrite Circulator

The birefringent ferrite circulator is the most popular circulator today. It is a three-port device constructed in either waveguide or stripline transmission line. Fig. 11 is a sketch of the waveguide model. A ferrite post is centrally placed in the *Y*-junction. A *C*-type permanent magnet is used to produce a transverse magnetic bias field. The wavefront, propagating in a transversely magnetized ferrite medium, rotates about the axis of the magnetizing field. This phenomenon is known as *birefringence*, hence the name birefringent circulator. The DC bias is adjusted to produce 60° of bending resulting in transmission from port 1 to port 2, port 2 to port 3. and port 3 to port 1 for a symmetrical *Y*-waveguide configuration. The stripline model operates in exactly the same manner, except for the construction of the device; the latter model contains a center conductor plate which divides the narrow dimension of the waveguide in half. Very low insertion loss may be achieved when using this model circulator (0.1 to 0.2 db) with isolation greater than 30 db. It is however a relatively low-power device.

FERRITE ATTENUATOR AND MODULATOR

Many ferrite devices are variations of circulators or isolators. For example, a microwave variable attenuator may be constructed by replacing the permanent magnet of the Faraday rotation isolator with an electromagnetic coil. The output of such a device is proportional to the cosine of the angle between the direction of the *E*-field polarization as it leaves the ferrite and the normal to the output resistive card. The *E*-field polarization is a function of the magnitude of H_o. By varying the current in the coil, H_o is varied so that variations are produced in the total angle of rotation of the *E*-fields as they propagate through the ferrite.

This same device may be used as a microwave modulator where the modulation information is applied to the coil. Modulation at audio frequencies has been easily accomplished. Extension to higher modulation frequencies is being studied: the limiting factor is the modulation coil inductance. An SPST ferrite switch may also be constructed out of the Faraday rotation isolator where the *on* condition occurs with a coil bias current providing an *E*-field orientation perpendicular to the output resistive card; the *off* condition occurs with a coil bias current which produces an *E*-field orientation parallel to the output resistive card. A SPDT ferrite switch may be constructed out of the Faraday rotation circulator or the birefringent circulator. In the latter device the transverse bias field for one "position" of the switch is adjusted for propagation from port 1 to port 2. For the other position of the SPDT switch, the bias field is reversed in direction but kept equal in magnitude. Propagation then occurs from port 1 to port 3.

FERRITE PHASE SHIFTER

Ferrites have also been used most effectively for microwave variable phase shifters. The most popular model is probably the longitudinally magnetized ferrite phase shifter (see Fig. 12). It consists of a section of waveguide with a ferrite rod centrally located along its axis. A variable longitudinal magnetic field is applied to the ferrite by means of an electromagnetic coil wrapped around the waveguide. The ferrite is operated in the region below saturation. In this region, the off-diagonal component K of the tensor permeability may be neglected. The large differential phase shift occurs because of the large initial increase in μ, the diagonal component of the tensor permeability, as the applied field is increased from zero. This initial rise is not predicted by Equations 2 or 3. As μ increases there is a tendency for the microwave fields to concentrate in the ferrite rod. With this concentration of fields, there is an associated decrease in guide wavelength; this decrease is manifested as an increase in phase delay. The phase differential increases as μ increases, and continues until saturation is reached; at this point, μ begins to decrease obeying the equations discussed earlier. Differential phase shift as high as 700° may be achieved by this device with little variation of insertion loss.

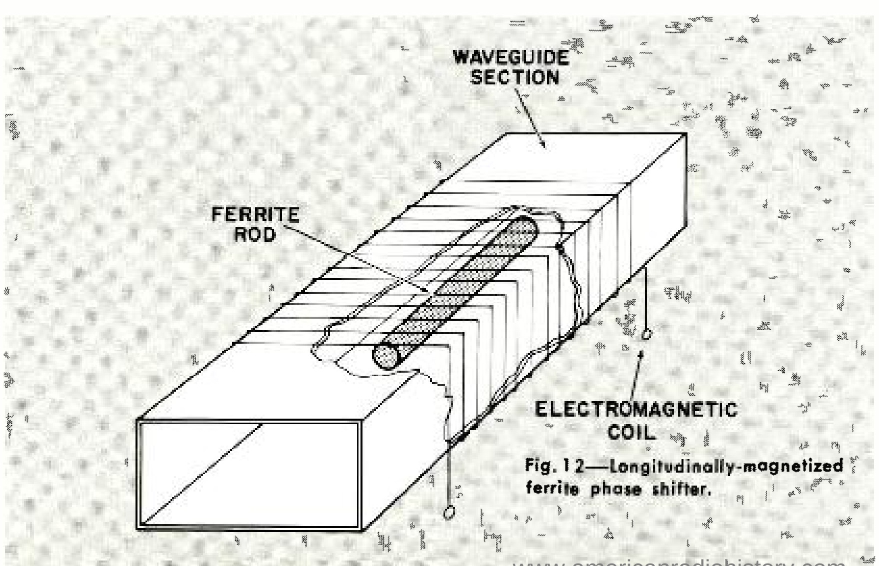

Fig. 12—Longitudinally-magnetized ferrite phase shifter.

CONCLUSION

There are many other ferrite devices and phenomena which space does not allow discussing here. This includes the entire area of non-linear effects in ferrite materials involving spin wave theory, limiting action and ferromagnetic amplification.

The field of microwave ferrites is still in its infant stages. Many areas ranging from material development to device development remain unexplored. The literature is filled with informative and sophisticated discussions of the subject; the following bibliography is a good starting point for the interested reader.

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GALLIUM ARSENIDE DEVICES—THE STATE OF THE ART

Gallium arsenide is a new semiconductor material with exciting possibilities for revolutionizing broad areas of the electron-device field. Because GaAs has a wider bandgap than silicon and higher electron mobility than germanium, it offers outstanding high-temperature, high-frequency performance, and can be expected to dominate in three specific device areas: high-frequency varactor diodes, solar cells, and tunnel diodes for microwave oscillator applications. Future developments in materials and device technology can be expected to lead to the use of GaAs in high-temperature transistors and rectifiers. This paper cites the reasons for the growing interest in GaAs and describes recent advances in devices using this material.

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GALLIUM ARSENIDE is a III-V compound semiconductor having a zinc-blende crystal structure identical to that of germanium and silicon, except that alternate sites are occupied by gallium and arsenic atoms (Fig. 1). This dual occupancy results in marked differences in the crystal properties, which are reflected in device technology.

SIGNIFICANT GaAs PROPERTIES

The opposite [III] crystal faces differ; one consists exclusively of gallium atoms (the [III] face); and the other of arsenic atoms (the $\overline{[III]}$ face). When these two faces are etched, the arsenic face reaches a high polish, and deep etch pits appear on the gallium face, which is more highly reactive.[1] Similarly, the rate of epitaxial growth on the gallium face is generally several times that on the arsenic face. Such differences must be recognized and means provided for wafer orientation early in the processing of the device. Polishing etches have proved effective for this purpose.

The energy-band structure for GaAs (Fig. 2) shows that GaAs has a wide energy gap which separates the valence and conduction bands (approximately 1.4 ev at room temperature); the values for silicon and germanium are 1.07 and 0.67 ev, respectively. The energy gap determines the temperature range over which a semiconductor can be used. The low electron effective mass of GaAs (0.072Mo) indicates that the electron mobility is unusually high. A comparison of effective carrier masses for germanium, silicon, and GaAs is shown inset in Fig. 2.

Further, the conduction-band minimum and the valence-band maximum of GaAs occur at the same point in k-space. In other words, an electron-hole pair can recombine directly across the energy gap without momentum transfer from the lattice (a phonon interaction). This mode of recombination differs greatly from that in germanium and silicon, in which recombination occurs by means of trapping centers. Later it will be shown that this feature has important implications for minority-carrier lifetime in GaAs crystals.

The combination of a large energy gap and a high electron mobility found in GaAs is quite unique, as shown in Fig. 3—the energy-gap and electron-mobility values for several semiconductors. Although there are some compound semiconductors which have wider bandgaps (such as gallium phosphide) and some with high electron mobilities (such as indium arsenide), GaAs has no peer among materials which operate at 100°C and higher.

The advantage of the high electron mobility of GaAs becomes even more striking as the doping level in the semiconductor is increased. As shown in Fig. 4, the presence of dopant impurity atoms in the crystal lattice reduces the carrier mobility. However, because of the partially ionized state of the GaAs lattice itself, the decrease of mobility with doping level is smaller in GaAs at doping levels of 10^{17} cu cm, which are common in transistors and diodes. The mobility values on the right edge of Fig. 4 apply

Fig. 1—Crystal Structure of Gallium Arsenide

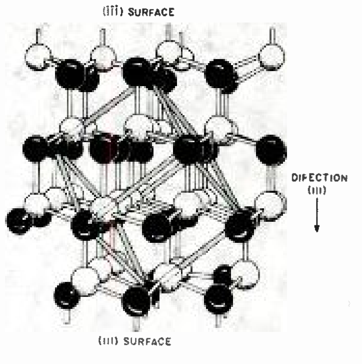
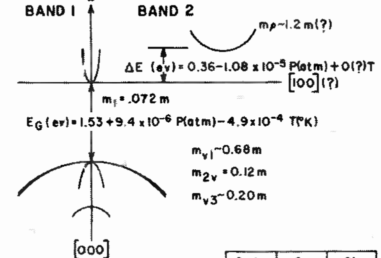
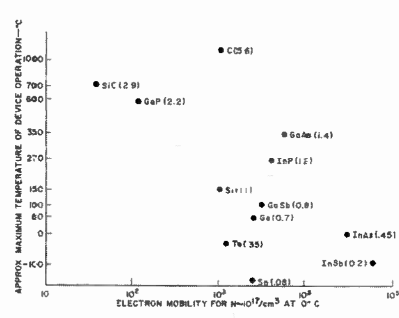


Fig. 2—Energy-band Structure of Gallium Arsenide



	GaAs	Ge	Si
EFFECTIVE ELECTRON MASS (Mo)	0.072	0.55	1.1
EFFECTIVE HOLE MASS (Mo)	0.5	0.37	0.59

Fig. 3—Energy-gap and Electron-mobility Values for Several Semiconductors (The energy-gap value is written adjacent to the point representing the semiconductor.)



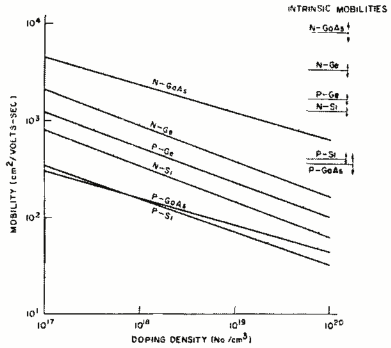


Fig. 4—Variation of Mobility as a Function of Impurity Concentration. The mobility values on the right hand side apply to very lightly doped intrinsic materials.

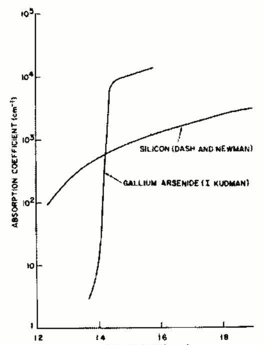


Fig. 5—Absorption Coefficient as a Function of Photon Energy for Silicon and Gallium Arsenide

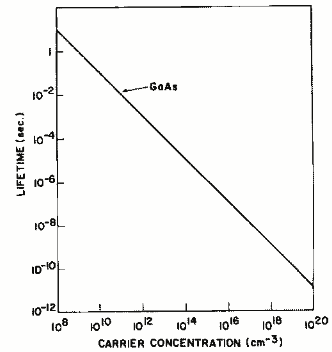


Fig. 6—Direct Recombination Lifetime in Gallium Arsenide

to very lightly doped (intrinsic) materials. The difficulties in achieving the limiting mobility value of 12,000 sq cm/ to gauge the progress still to be made in volt-sec for n-type GaAs can be used purification and crystal growth. Today. GaAs crystals having mobilities of 6000 sq cm/volt-sec at doping levels of 10^{16}/ cu cm are available, but silicon and germanium crystals with 10^{12}/cu cm dopant levels and mobilities very near the intrinsic values are common.

The final fundamental property to be considered is minority-carrier lifetime in GaAs crystals. Considerable basic information on carrier lifetime can be gained from examination of the optical absorption curves (Fig. 5) for GaAs and silicon. The abruptness of the absorption curve of GaAs implies that a hole-electron pair is created easily with energy even slightly in excess of the bandgap value. Detailed balance arguments can be used to show that the converse process or direct recombination also occurs easily. Kudman and Kinsel used the shape of the absorption edge in a relation by Van Koosbroeck and Shockley[2] to calculate the recombination rate for minority carriers. Their results. which set a *maximum* value for lifetime, are shown in Fig. 6. The lifetime is inversely proportional to the majority-carrier density; for example, at the 10^{17}/cu cm doping level. the lifetime for direct recombination is 20 nsec. Measured lifetimes in today's crystals are less than this value by at least an order of magnitude. It will be shown later. however. that it is possible to build some very useful high-frequency semiconductor devices using this material.

SOLAR CELLS

Recently, GaAs solar cells have been made available for sampling by the Semiconductor and Materials Division. The use of GaAs results in cells with a higher potential efficiency. considerably less fall-off of efficiency at high temperature, and. most important. outstanding radiation resistance. As a result. GaAs is. in many ways. better suited than silicon for use in solar cells.

The efficiency of a solar cell depends on the fraction of incident photons in solar energy that the cell can use, and on the open circuit voltage of the cell (this voltage is directly related to the forbidden-energy gap). Fig. 7 shows the energy distribution in the solar spectrum and indicates the portion of the spectrum which can be used by solar cells of various semiconductor materials. As shown. because silicon cells can use all photons with energies greater than 1.07 ev, 2.8 × 10^{17} hole electron pairs can be generated per sq cm/sec. In GaAs cells. however. only those photons having energies greater than 1.4 ev can be used and the total pair current is 1.8 × 10^{17} hole-electron-pairs/sq cm/sec. However, if the current is lower. the voltage across the solar cell is considerably higher for GaAs. These two factors and others relating to the saturation current of the forward biased junction were correlated by M. Wolf[3] in the curve shown in Fig. 8. which plots limiting solar-cell conversion efficiency as a function of the energy gap of the semiconductor material. This curve indicates an optimum energy gap very nearly equal to 1.4 ev for GaAs.

There are two other factors of importance for solar-cell conversion efficiency. First, it is desirable to collect all minority carriers generated by the incident light. Although the lifetime in GaAs crystal is very small, as previously discussed. this factor is counteracted by the very high absorption coefficient in GaAs for radiation greater than bandgap. As a result. all carriers are created in a very shallow region near the surface. Good collection efficiencies can be achieved by the use of a very shallow junction and a large built-in field from the impurity distribution. Second, it is important to minimize series-resistance losses in the cell. Solar cells of GaAs generally display very small resistive losses because of the possibility of doping GaAs p-type material very heavily and because of the higher impedance level of the cells (smaller currents at higher voltages) relative to silicon. Fig. 9 compares the static characteristic curves of typical silicon and GaAs cells. and illustrates the higher impedance level.

Although the GaAs cells presently offer efficiency comparable to that of silicon cells, the higher temperature performance capability and greater radia-

Fig. 7—Useful Portion of the Solar Spectrum

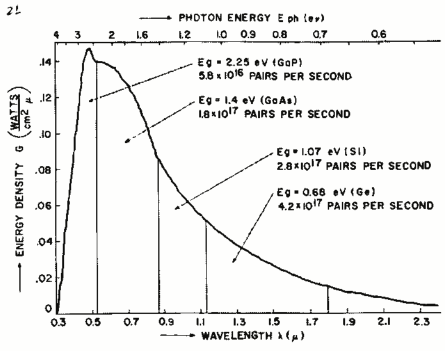


Fig. 8—Limit Conversion Efficiency as a Function of the energy-gap Width for Signal p-n Junction Solar Cells (Direct Transition type)

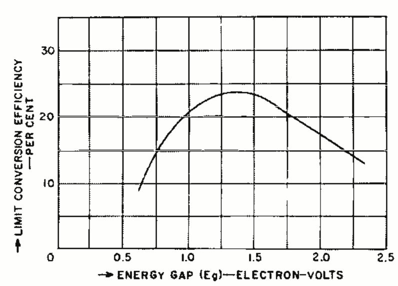


Fig. 9—Typical Solar Cell Characteristics

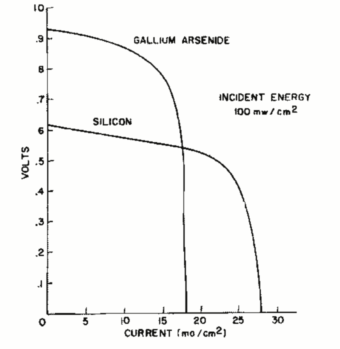


TABLE I — Solar-Cell Radiation Resistance*

(Flux Required to Reduce Cell Efficiency by 25 Percent)	
18-Mev protons:	
GaAs Cells	6 x 10^{12}/cm^2
Silicon n-on-p	
Silicon p-on-n cells	7 x 10^{11}/cm^2
800-kev electrons:	
GaAs Cells	1 x 10^{15}/cm^2
Silicon n-on-p Cells	1 x 10^{15}/cm^2
Silicon p-on-n Cells	4 x 10^{13}/cm^2

**Data supplied by J. Wysocki, RCA Labs.*

tion resistance of GaAs are of more importance to the user. Because of the wider bandgap, GaAs cells continue to show very respectable energy-conversion efficiencies even at 150°C, where silicon cells are cut to half their room-temperature efficiency. The experimental curves on GaAs and silicon solar cells shown in Fig. 10 illustrate this striking difference in temperature dependence. As a result, GaAs cells permit the use of reflectors for energy concentration in satellites without undue concern for overheating of the solar cells. Furthermore, the availability of GaAs cells simplifies the design of solar-cell power supplies for missions which go near the sun, such as the Venus Probe.

In addition, preliminary data indicate that GaAs solar cells are approximately order of magnitude more resistant to high-energy proton radiation damage than silicon cells. The significance of the results of recent radiation damage measurements (Table I) at RCA Laboratories is illustrated[4] in Fig. 11, which compares GaAs and silicon solar-cell degradation as a function of time in the Van Allen Belt. The resulting longer life offered by a GaAs solar-cell power supply is an attractive feature in such space applications as extended data-gathering missions in the Van Allen Belt or in communication satellites at that altitude.

VARACTOR DIODES

Where the solar cell takes advantage of the large band gap in GaAs, the varactor diode takes advantage of the high electron mobility. As shown in Fig. 12, the basic figure of merit for varactor diodes is determined by the electron mobility, which depends on close control of device geometry. The dielectric constant for GaAs is equal to that of silicon. As shown in Fig. 4, the advantage in mobility increases from a factor of four at low doping levels to a factor of seven at doping levels of 10^{18}. A second important property for varactors is the avalanche breakdown voltage, which is dependent on the doping level and the ionization properties of the material. Fig. 13 shows[5] that at very high dopings, breakdown voltage for GaAs levels off at 8 volts rather than at 4 volts as it does for silicon.

Today, varactor technology is undergoing intensive changes as increasing use is made of epitaxially grown semiconductors in which the material near the junction is lightly doped and the remainder of the wafer is very heavily doped. Evidence of the effectiveness of this technique is seen in today's silicon varactors, and it is expected that GaAs varactors will soon show similar advances. At the present time, some of the best low-voltage (less-than-30-volt) varactor diodes are made from GaAs, while the high-voltage, high-power varactor field is dominated by epitaxial silicon units. As GaAs epitaxial devices continue to develop, it can be anticipated that the voltage breakdown boundary, now at 30 volts, will shift upwards.

The Smith Chart plot (Fig. 14) shows the transformed impedance of a GaAs varactor diode having a cutoff frequency higher than 200 Gc. This value was obtained from 10-Gc data. At this time, units with cutoffs over 300 Gc are available for sampling. The use of these diodes results in exceptionally low-noise figures in X-band parametric amplifiers, as well as high-energy conversion efficiencies in harmonic generators. Table II shows some examples of the outstanding amplifier performance achieved with these GaAs varactor diodes.

TABLE II — Performance Data of Gallium Arsenide Varactor Diodes (Amplifier: Non-degenerate Negative-Resistance Type)

	Group I	*Group II*
Signal Frequency, Mc	9375	8700
Pump Frequency, Mc	35800	35000
Gain, db	15	10
Bandwidth, Mc	9.0	200
Noise Figure, db	2.8	2.9
Pump Power, mw	125	180

TUNNEL DIODES

One of the most important factors in tunnel-diode applications is the power-handling capabilities of the diode. Gallium arsenide offers an advantage over germanium in this respect. As shown in Fig. 15, the voltage swing for GaAs (and therefore the magnitude of the negative resistance) is twice that for Ge for identical peak currents. Alternatively, GaAs diodes have twice the current swing and twice the voltage swing of Ge diodes for the same impedance level. The power output of GaAs diodes is therefore four times that of Ge diodes. Despite this power advantage, and the improved frequency response resulting from high electron mobility, GaAs tunnel diodes have certain limitations.

It has been found that the peak current of very-high-frequency GaAs tunnel diodes tends to decrease when the diodes are operated beyond the valley at a current level comparable to, or greater than, the peak current. The reasons for this effect are now well understood, and the conditions required for safe operation have been established. For example,

Fig. 10—Solar Cell Efficiency at High Temperatures

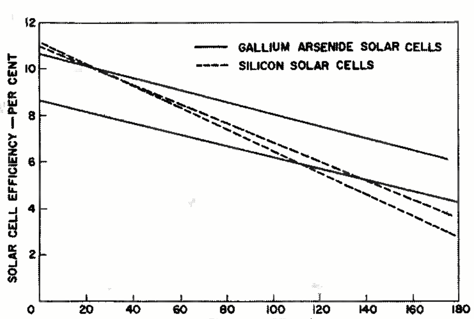


Fig. 11—Solar Cell Degradation as a Function of Time in the Van Allen Belt (Courtesy of P. Rappaport)

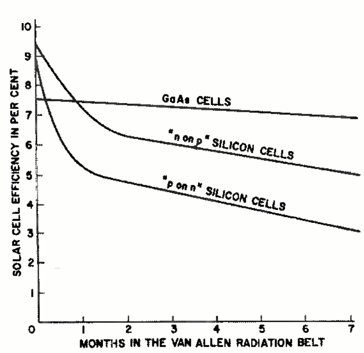
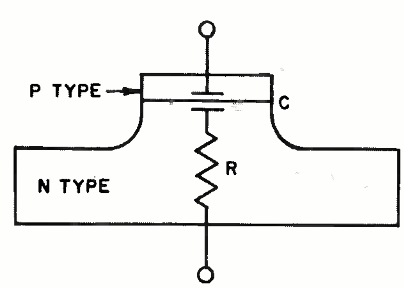


Fig. 12—Figure of Merit for Semiconductor Varactors



$$f_{co} = \frac{1}{2\pi RC} \alpha \frac{\mu_n}{\sqrt{\epsilon}}$$

high-frequency GaAs tunnel diodes can be used safely in oscillator circuits and switching circuits in which the diode is switched to a point only slightly beyond the valley. A particularly important example of the latter is the tunnel-diode-and-transistor hybrid switch, in which the diode voltage is limited to the transistor emitter-to-base voltage. Recently, a 200-Mc GaAs tunnel-diode oscillator was operated for 500 hours at 150°C with 10 mw of output, and showed *no change* in the current-voltage characteristics. Similar life tests indicate that the problem of changing peak current does not arise for a wide variety of applications.

Gallium arsenide offers the highest performance capabilities for high-frequency tunnel diodes. In the expression for the resistive cutoff frequency, it is seen that:

$$f_{max} = \frac{\sqrt{\frac{r_j}{r_s} - 1}}{\frac{r_j}{r_s}} \left(\frac{1}{2\pi r_s c_j} \right)$$

If the ratio r_j/r_s is optimized:

$$f_{max} = \frac{1}{4\pi r_s c_j{}^{-uc}} \left(\propto \frac{\mu}{\sqrt{\epsilon}} \right)$$

This result is similar to the varactor figure of merit (Fig. 12). Here, again, the low-dielectric constant and the high mobility (particularly at high doping levels) make GaAs superior. In fact, some of the highest-frequency and highest-power tunnel-diode oscillators have used GaAs diodes. Point-contact GaAs tunnel diodes[8] have been used for fundamental frequency oscillations at more than 100 Gc. Diodes achieving 4 mw at 6 Gc have been reported. The RCA Microwave Applied Research group at Princeton reports that they have attained 22 mw at 2 Gc from single diodes and even more from multiple-diode oscillators.

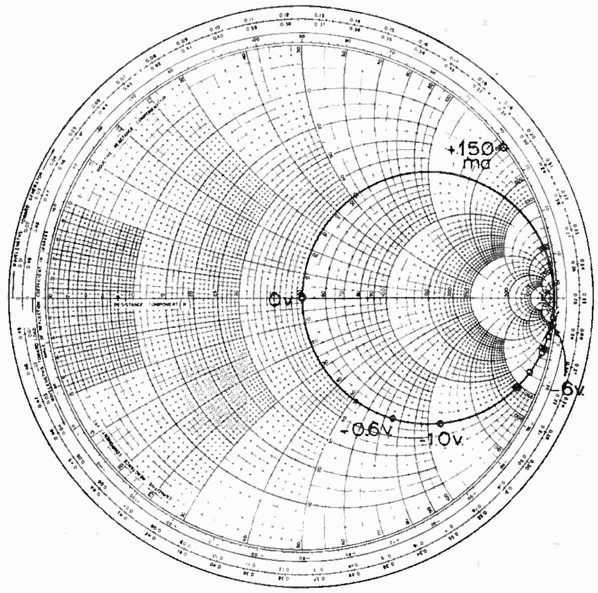


Fig. 14—Varactor Diode Cutoff Frequency Measurements

LIGHT-GENERATING DIODES

Recent research at the Princeton Laboratories has revealed that a forward-biased GaAs diode can be made to produce infrared light with efficiencies that may exceed 50 percent. This value is very respectable; incandescent bulbs achieve 4-percent and fluorescent lights 20-percent efficiency. Furthermore, the infrared output of the diode can be modulated at frequencies up to about 200 Mc, the limit of readily available photocell detectors. Such high-frequency modulation is far beyond the capability of any other known light source.

The light in the GaAs diodes is generated by injection luminescence, a process in which injected minority-charge carriers recombine across the forbidden-energy gap to emit photons. The theoretical efficiency of this process in GaAs approaches 100 percent because no phonons (lattice vibrations) are involved, as is the case with germanium and silicon. As shown in Fig. 2, there is no horizontal displacement of the valence and conduction band edges. The high-frequency modulation capability of light-generating diodes results from a very low junction resistance-capacitance time constant and the very low minority-carrier lifetime of the material.

The previously unrecognized ability of GaAs diodes as light sources will undoubtedly have many useful applications in the consumer, industrial, and military electronic fields.

TRANSISTORS

In transistor technology, GaAs offers great promise of improving the capabilities of the device at the higher temperature ranges. Fig. 16 illustrates some of the present limitations on transistor performance. Two time constants are important: the transit time across the base region and the $r_{bb'}c_c$ time constant for change of the collector capacitance. The relations in Fig. 16 indicate that both the electron and hole mobilities are important for transistors; as previously shown in Fig. 4, the hole mobility in

Fig. 13—Breakdown Voltage of Gallium Arsenide and Silicon p-n Junctions as a Function of Carrier Concentration

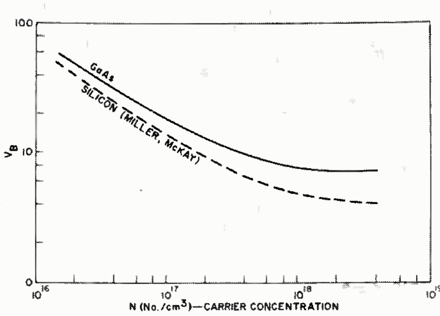


Fig. 15—Comparison of Current-voltage Characteristic of Germanium and Gallium Arsenide Tunnel Diodes

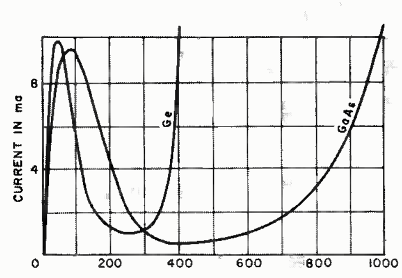
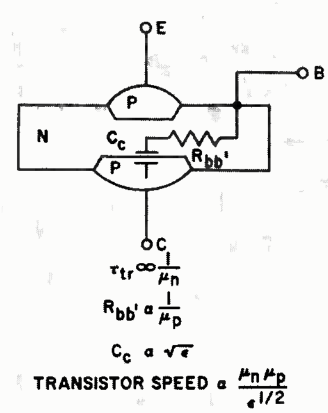


Fig. 16—Figure of Merit for Transistors



GaAs is the same as that in silicon. If numbers are substituted in the relation for transistor speed, it is evident that, for typical base dopings, GaAs is as good as germanium and is better than silicon by a factor of seven. As previously shown, GaAs is capable of operation at much higher temperatures than germanium or silicon. Although the relation (Fig. 16) for transistor speed indicates an advantage of a factor of seven for germanium over silicon, these two transistor types are separated at most by a factor of two as a result of the advance in silicon technology over the past five years. A similar gap exists between the performance capabilities of GaAs and the development of effective fabrication techniques. Only in the past three years have GaAs transistors been developed which have appreciable current gain. This advance was based on the use of very narrow base widths, which permit carriers having short lifetimes (less than those shown in Fig. 6) to diffuse through the base region. The structure used is an alloyed-emitter, diffused-base mesa transistor with base widths of about 1 micron (see Table III). Although attempts are presently being made to use today's advanced technologies in the fabrication of even better units, GaAs transistors are not yet available on the commercial market.

TABLE III — Gallium Arsenide Switching Transistor

h_{fe} (1 kc, $I_E = 10$ ma)	29.5
B^V_{CBO} ($I_c = 100\ \mu$a)	18 volts
I_{CBO} (5 volts)	0.77 nsec
C_{ob} (5 volts)	8.8 pf
R_{CS} ($I_c = 10$ ma)	45 ohms
t_d	16 nsec
t_r	24 nsec
t_s	8 nsec
t_s	16 nsec

RECTIFIERS

The previous discussion of varactor and tunnel diodes highlighted the high-electron mobility of GaAs. However, it is not possible to take advantage of this property in GaAs rectifiers. A silicon rectifier structure has three distinct semiconductor regions: the heavily p- and n- doped end regions and a thick, almost intrinsic, intermediate region. In reverse bias, this latter region sustains the high voltage across the device; in forward bias, injected carriers from both ends modulate its resistance. In typical silicon rectifiers, this conductivity modulation process requires lifetimes of 1 to 10 μsec. Because of problems of purification and lifetime, today's GaAs technology does not permit the use of the above structure. Instead, p-ν structure is used which results in higher forward resistances (there is only slight conductivity modulation) and lower breakdown voltages. In this structure, the low leakage currents and good high-temperature performance inherent to GaAs are still preserved. The Semiconductor and Materials Division is now developing a rectifier which can operate at temperatures up to 350°C. Fig. 17 compares the normalized rectified current of this unit to that of a silicon rectifier. This extended performance is a result of the low-leakage which GaAs rectifiers and diodes generally have at high temperatures. As shown in Fig. 18, the leakage current reaches 1 ma only at 350°C.

A low leakage current at room temperature is desirable for diode vibrating reed-type amplifiers. These amplifiers are essentially low-noise low-frequency varactor amplifiers. Fig. 19 shows the characteristic of a diode designed for such applications.

CONCLUSIONS

The two basic crystal properties which give GaAs an advantage over silicon and germanium are its high electron mobility and wide forbidden-energy gap. These features have led to three specific device areas where GaAs can be expected to dominate: high-frequency varactor diodes, solar cells, and tunnel diodes for microwave-oscillator applications. Future developments in material and device technology can also be expected to lead to the use of GaAs in high-temperature transistors and rectifiers.

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Fig. 17—Average Rectified Current as a Function of the Operating Temperature

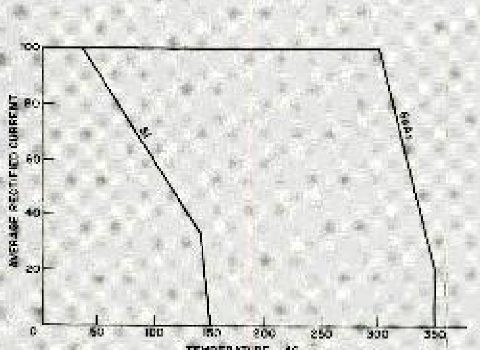


Fig. 18—Reverse Characteristic of Gallium Arsenide

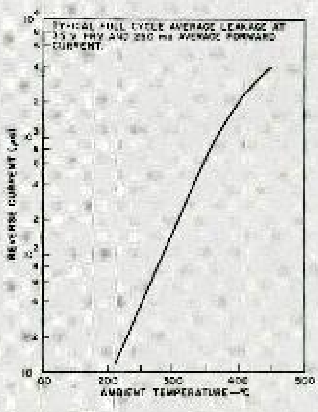
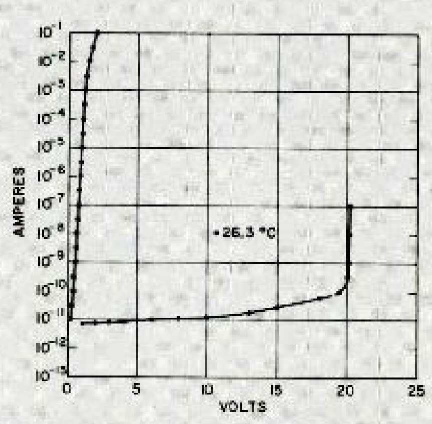


Fig. 19—Current-voltage Characteristic (in vacuum) of Gallium Arsenide Diode (courtesy of J. Halpern and R. H. Rediker)



A PROPOSED VARIABLE-DATA-RATE COMMUNICATION TECHNIQUE

More-efficient communications may be achieved by varying the data rate so that it is always a maximum relative to ambient channel conditions, compared to the conventional fixed-rate approach which requires design for worst-case conditions. The method proposed here is practical for troposcatter and satellite systems, and promises improvements in interference and primary power consumption.

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Many radio communication systems have channel characteristics that change with time. The received signal and available bandwidth vary as functions of interference, multipath reception, local weather conditions, and other factors. Common practice today is to design a communications system around the worst case expected. While *worst-case* design assures transmission with a certain long-term confidence, transmitter power and/or channel capacity are wasted whenever the channel is much better than the worst case used in system design.

A communications system in which data rate is varied so that it is always maintained at the maximum level compatible with the channel conditions utilizes the channel more efficiently. Such a system, proposed herein, results in an improved over-all system performance. Prominent examples of communications systems which would benefit from variable-data-rate techniques are those employing tropospheric scatter, where received signal levels and available bandwidths vary widely with time.

GENERAL BANDWIDTH RELATIONSHIPS

Data rate can be varied in a straightforward manner by changing the bandwidth of the transmitted signal. When a suitable detection scheme is employed at the receiver, this shift in bandwidth also changes the effective signal-to-noise ratio at the receiver. For example, when a signal with an energy per digit E_o is keyed at a rate K_o and transmitted in a channel where the noise power per unit cycle of bandwidth is N_o, then in a bandwidth B_o at the receiver is: $R_o = E_o/N_o$.

If, however, the signal is keyed at a rate K with energy per digit equal to $E_o(K_o/K)$, i.e. transmitter power held constant and transmitted in a bandwidth B, where $B = B_o(K/K_o)$, the signal-to-noise ratio is given by: $R = (E_oK_o/N_oK) = (R_oB_o/B)$.

For a variable-bandwidth detector (or equivalently, a variable integration time detector) matched to the received signal's bandwidth, signal-to-noise ratios into the detector are given by the above two equations for R_o and R. The change in signal-to-noise ratio (expressed in db) when the bandwidth varies from B_o to B is then: $R = 10 \log (B_o/B)$.

Thus, halving the bandwidth causes an effective *increase* in signal-to-noise ratio of 3 db, while doubling the bandwidth causes an effective *decrease* in signal-to-noise ratio of 3 db.

In systems where received signal power varies over a wide range, a variable-bandwidth system could be used to maintain a constant signal-to-noise ratio at the input to the detector. Important system parameters are:

1) The mean value of the receiver signal-to-noise ratio R_o in a bandwidth B_o.
2) An average value of the data rate corresponding to B_o. This value would be chosen on the basis of the average long-term error rate desired.
3) The instantaneous bandwidth of the transmitted information B. This bandwidth would be determined from the instantaneous received signal-to-noise ratio R according to the relationship $B/B_o = R/R_o$.

Thus, compensation for decreases in the received signal level are made by transmitting with a reduced bandwidth (lower data rate); increases in signal strength allow transmissions with expanded bandwidth. In theory, the signal-to-noise ratio at the receiver would be maintained always at a constant value R_o.

In practice however, the data rate (bandwidth) cannot be allowed to vary from zero to infinity; a better approach is to allow data rate to vary only over realistic limits. For the case of bandwidth variation over a limited range of data rates, the analysis is somewhat more complex. As shown in Fig. 1, detector operation is divided into three separate regions, determined by the upper limit of the received signal-to-noise ratio. No further bandwidth expansion takes place above this limit (upper truncation level); below the lower limit of received signal-to-noise ratio (lower truncation level), no further reduction in bandwidth takes place.

In Region II, between the two truncation levels, bandwidth of transmitted signal is varied according to $B/B_o = R/R_o$; a constant signal-to-noise ratio of R_o is maintained at the detector input.

The Region I, the channel signal-to-noise ratio is below the lower truncation level R_L; at all times, signal bandwidth is the smallest permissible value corresponding to the ratio at the lower truncation level. The signal-to-noise ratio into the detector falls below R_o as the signal-to-noise ratio in the nominal channel (bandwidth B_o) falls below R_L.

In Region III, the signal-to-noise in the nominal channel is above the upper truncation level R_u. The signal bandwidth is the largest value permissible at all times. The signal-to-noise ratio into the detector rises above R_o as the signal-to-noise ratio in the nominal channel rises above R_u.

SYSTEM PERFORMANCE IN RAYLEIGH FADING WITH NO DIVERSITY

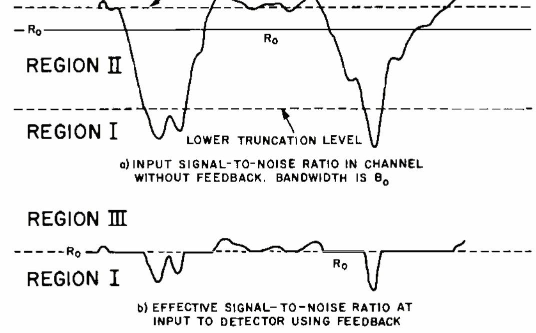
The behavior of an FSK (frequency shift keying) data channel was investigated under three conditions:

1) Fixed bandwidth.
2) Variable bandwidth, no limits on the range of variation.
3) Variable bandwidth, limits on range of variation.

When transmission is at a fixed rate, the probability of a digit error is given by (see Curve F, Fig. 2):

$$P_e = \tfrac{1}{2}\int_o^\infty \exp(-R/2)\exp(-R/R_o)\ \frac{dR}{R_o} = \frac{1}{2(1 + R_o)}$$

Fig. 1—Waveforms showing detector operation (effect of bandwidth variation on received S:N ratio) for three different regions of bandwidth variation.



*This paper was written while Mr. Palmer was with DEP Applied Research; he is now associated with Gannon College, Erie, Pa.

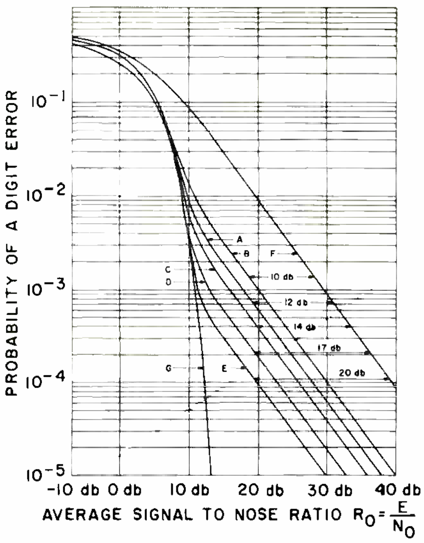


Fig. 2 — Performance of FSK channel in the presence of Rayleigh fading (no diversity).
A—variable bandwidth, 100:1 range, average data rate is D_0.
B—variable bandwidth, 50:1 range, average data rate is D_0.
C—variable bandwidth, 100:1 range, average data rate is 0.983 D_0.
D—variable bandwidth, 100:1 range, average data rate is 0.865 D_0.
E—variable bandwidth, 100:1 range, average data rate is 0.632 D_0.
F—fixed bandwidth, average data rate is D_0.
G—variable bandwidth, infinite range, average data rate is D_0.

When the fading is completely cancelled by unlimited data rate variation, the probability of a digit error is given by[1] (see Curve G, Fig. 2):

$$P_e = \tfrac{1}{2} \exp(-R_o/2)$$

When the fading is only partially cancelled by putting limits on the range of variation of data rate, the effect is to "slice out" the portions of the signal where the feedback is active (see Fig. 1), producing a signal for detection which has fades only when the signal is outside Region II.

In Region I, the effective signal-to-noise ratio is given by R_o/R_L; in Region III, the effective signal-to-noise ratio is given by R_o/R_u; where x and y are defined as R_L/R_o and R_u/R_o, respectively.

Actually, the operation takes place in each of the three regions at various times. To obtain the long-term average probability of system error, the average probability of error, P_e, must be determined in each of the regions as well as the percentage of operating time spent in each region. P_e can then be obtained from:

$$P_e = \int_o^\infty P_e(R)P(R)\,dR$$

$$= \int_o^{xR_o} P_{eI}(R)P(R)\,dR + \int_{xR_o}^{yR_o} P_{eIII}(R)P(R)\,dR. \quad \int_{yR_o}^{\infty} P_{eII}(R)P(R)\,dR +$$

Where: P_{ej} = probability of error in Region j.

The upper and lower truncation levels can be chosen so as to maintain the average data rate at that value corresponding to B_o. This requires that x and y must satisfy the relation:

$$x + \exp(-x)\exp(-y) = 1.$$

For purposes of illustration, values of x and y satisfying this equation with $y/x = 100$ were chosen as representative values for a truncated variable rate system; the long-term probability of error is plotted as Curve A of Fig. 2. To further illustrate the range of data-rate variations y/x and the effect of varying the average data rate, several specific cases are listed in Table I; these considerations are plotted in Fig. 2.

TABLE I—Specific Cases of Data-Rate Variations

Curve	*x*	*y*	*Range of Variation*	*Average Data Rate*
A	0.063	6.3	100:1	D_o
B	0.102	5.1	50:1	D_o
C	0.04	4.0	100:1	0.983 D_o
D	0.02	2.0	100:1	0.865 D_o
E	0.01	1.0	100:1	0.632 D_o

Note: D_o is the reference data rate used in computing Curves F and G.

The curves of Fig. 2 clearly show the use of truncated bandwidth variation produces a substantial increase in system performance compared with the fixed-bandwidth case. Also, the amount of improvement depends on both the range of data-rate variation and the average data rate transmitted; although the relationship has not been determined completely, it is indicated here qualitatively.

SYSTEM PERFORMANCE WITH DUAL DIVERSITY

The preceding analysis was repeated for the case of dual-diversity reception. The probability density of signal-to-noise ratio out of an "ideal" Nth-order diversity combiner is given by[2]:

$$P(R)\,dR = \left(\frac{R}{R_o}\right)^{N-1} \frac{\exp(-R/R_o)}{(N\text{-}1)\,!}\,\frac{dR}{R_o}$$

Specifically for dual diversity, this equation reduces to:

$$P(R)\,dR = \frac{R}{R_o}\exp(-R/R_o)\,\frac{dR}{R_o}$$

The mean value of this function is equal to $2R_o$. The control signal will be taken from the output of the combiner and the operation will be identical to operation in the nondiversity case. Since the bandwidth variations occur in both diversity channels, the combiner bandwidth will vary the same amount as that of the transmitter. However, since the new average value of signal-to-noise ratio out of the combiner is $2R_o$, it will be necessary to specify the upper and lower truncation limits in terms of $2R_o$ rather than R_o, as in the nondiversity case. For dual diversity, the x and y parameters are redefined as: $x = R_L/2R_o$, and $y = R_u/2R_o$.

In Region II, between the upper and lower truncation limits the bandwidth of the transmitted signal will be varied according to $B/B_o = R/2R_o$. In this case, the output signal-to-noise ratio will be maintained constant at $2R_o$.

In Region I, below the lower truncation level, the bandwidth will be fixed by $x = B_L/B_o = R_L/2R_o$. Here, the effective signal-to-noise ratio into the detector would be given by: $R_{eff} = R(2R_o/R_L) = R/X$.

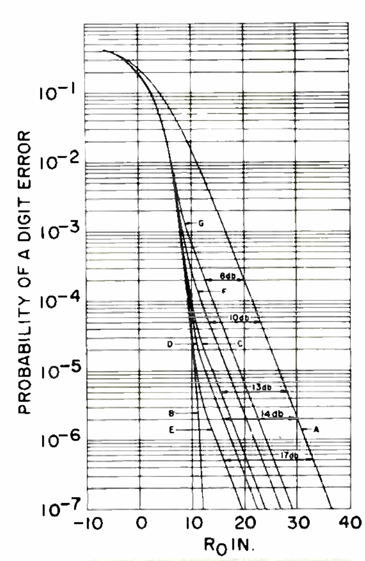
In Region III, above the upper truncation level the bandwidth at the transmitter would be fixed by: $y = B_u/B_o = R_u/2R_o$. And, the effective signal-to-noise ratio out of the combiner would be given by: $R_{eff} = R(2R_o/R_u) = R/y$.

To maintain an average data rate equal to the fixed-bandwidth case, x and y must satisfy:

$$x + (x+1)\exp(-2x) - (y+1)\exp(-2y) = 1$$

By the use of the distribution given earlier in the reduced equation for

Fig. 3—Performance of FSK channel in the presence of Rayleigh fading (dual diversity).
A—fixed bandwidth, fading.
B—fixed bandwidth, no fading.
C—variable bandwidth, 100:1 range, average data rate is D_0.
D—variable bandwidth, 100:1 range, average data rate is 0.998 D_0.
E—variable bandwidth, 100:1 range, average data rate is 0.729 D_0.
F—variable bandwidth, 50:1 range, average data rate is D_0.
G—variable bandwidth, 20:1 range, average data rate is D_0.



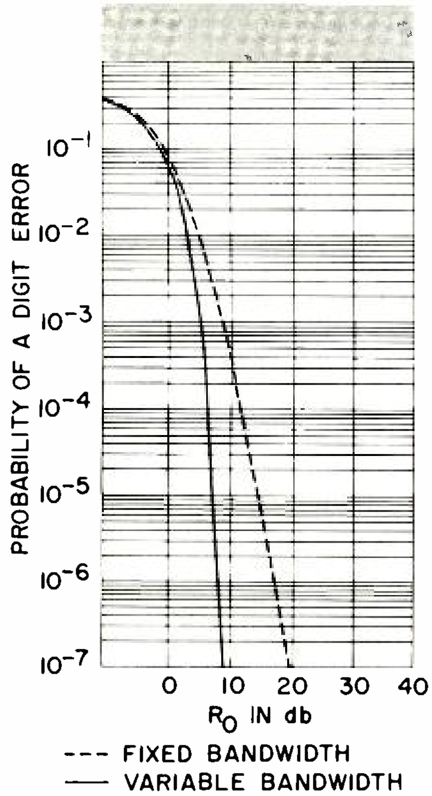


Fig. 4—Performance of FSK channel in the presence of Rayleigh fading (fourth-order diversity).

$P(R)\ dR$, and the redefined parameters x and y, the performance of an FSK channel was analyzed. The cases of interest, as before, are:

1) Rayleigh fading, fixed bandwidth.
2) Rayleigh fading, unlimited range of bandwidth variation.
3) Rayleigh fading, limited range of bandwidth variation.

The performance curves for several separate cases are shown in Fig. 3; the cases considered are listed in Table II.

PERFORMANCE WITH QUADRUPLE DIVERSITY

The performance of variable-bandwidth systems, when used in conjunction with fourth-order diversity reception was analyzed in a manner similar to the non-diversity and dual-diversity examples; Fig. 4 shows the results.

For fourth-order diversity the probability distribution out of the combiner is:

$$P(R)\,dR = \left(\frac{R}{R_o}\right)^3 \frac{\exp(-R/R_o)}{6} \frac{dR}{R_o}$$

The average (rms) value of the distribution given by this equation is $4R_o$, so that the data rate at any time is governed by:

$B/B_o = R/4R_o$, for the case of the unlimited feedback and for operation between the upper and lower truncation levels in the case of truncated operation.

$B/B_o = R_u/4R_o = y$, for operation above the upper truncation level (Region III) in truncated variable bandwidth operation.

$B/B_o = R_L/4R_o = x$, for operation when the received signal falls below the lower truncation level (Region I) in truncation variable rate operation.

In this case to achieve reference data rate, x and y must satisfy:

$$6x + \exp(-4x)(6 + 12x + 24x^2 + 16x^3) - \exp(-4y)(6 + 12y + 24y^2 + 16y^3) = 6$$

THE TRANSMIT TERMINAL

The proposed *transmit terminal* consists of a multiplexer, a tape loop, a modulator and a feedback receiver. Its operation (Fig. 5) is as follows: Multiplexed data with a fixed bandwidth appears at the output of the multiplexer, and is then recorded on the tape. The tape speed at the recording head is determined by the bandwidth of the multiplexer. The output data is recorded, and at the same time, a tone generated by a highly stable oscillator is also recorded on the tape for control purposes.

After a delay, the recorded information is played back from the tape; playback speed is variable, however, and determined by the feedback signal. The output of the feedback receiver is the signal-to-noise ratio at the main *receive* terminal relative to its long-term average (or R/R_o for nondiversity operation). Thus, the bandwidth of the signal at the playback head is directly proportional to the signal-to-noise ratio at the main *receive* terminal. The signal out of the playback head is sent to the modulator where it serves as the modulating signal. Note that the modulating signal now contains the control tone mentioned earlier. In case the RF transmission is AM-SSB, the RF transmission bandwidth is varied directly with the bandwidth of the playback signal.

When high- and low-speed governors are placed on the variable-speed capstan, the variation in playback-signal bandwidth is truncated, as discussed previously. If truncation limits are set to produce an average data rate equal to that of the system without variable bandwidth (according to previously discussed procedures), there is no net pile-up of data in the system. However, since data is put on the tape at a fixed rate and

TABLE II—Cases Considered in Analyzing FSK Performance (Fig. 3)

Curve	x	y	*Range of Variation*	*Average Data Rate*
A	1	1	0 Fixed Rate	D_o
B	0	∞	∞	D_o
C	0.055	5.5	100:1	D_o
D	0.04	4.0	100:1	0.998 D_o
E	0.02	2.0	100:1	0.729 D_o
F	0.105	4.8	47:1	D_o
G	0.18	3.6	20:1	D_o

taken off at a variable rate, there will be a variable delay in the time between recording and playback. When *playback* speed *is less* than the *record* speed, tape will "pile up"; piled-up tape is used up when the *playback* speed *exceeds* the *record* speed.

The function of the tape loop at the transmit terminal is to convert the fixed-bandwidth output of the multiplexer to a variable-bandwidth output which varies in direct proportion to the signal-to-noise ratio at the receive terminal.

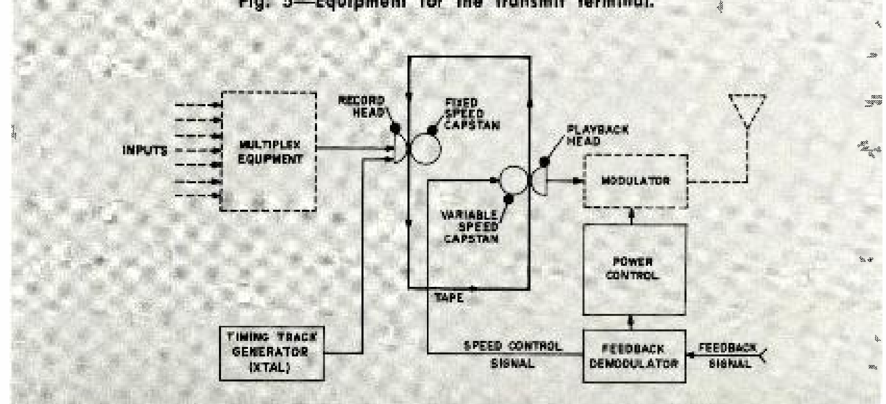
THE RECEIVE TERMINAL

The *receive* terminal (Fig. 6) consists of a demodulator, a tape loop, a feedback transmitter, and a demultiplexer. In operation, the signal is received and demodulated. The output of the demodulator is the modulating signal described in the preceding discussion. The bandwidth of the signal out of the demodulator will vary in direct proportion to the received signal-to-noise ratio (defined by $B/B_o = R/R_o$) when transit-time effects are neglected.

The output of the demodulator is then recorded on a tape; the speed of the tape passing over the record head varies in direct proportion to the signal-to-noise ratio. After a short delay, the recorded signal is played back at a fixed speed. Therefore, the bandwidth out of the tape loop at the receiver terminal is always equal to the original fixed bandwidth of the transmitter multiplexer and can be fed directly into the fixed-bandwidth receiver demultiplexer. No modification of the demultiplexer is required.

However, due to transit time delay, the value of R/R_o available at the

Fig. 5—Equipment for the transmit terminal.



receiver will differ slightly from the value of R/R_o used at the transmitter to set the transmitted bandwidth. Thus, the value of R/R_o available at the receiver can be used to set tape speed at the record head on an approximate basis; fine adjustment of tape speed is obtained by using the control tone generated at the transmitter. This control tone (after passing through the tape loop at the transmitter) is transmitted to the receive terminal; next it is inspected at the record head and compared with a highly stable local tone, equal in frequency to the oscillator control tone originally generated. The tape speed is adjusted until the received control tone agrees with the local tone at the receiver. Selection of the transmitted control tone is made easier by the approximate adjustment of tape speed according to the value of R/R_o existing at the receiver.

In view of the methods outlined for controlling the tape loop at the receive terminal, it seems desirable to measure both R and R_o at the receive terminal and to transmit the ratio R/R_o over the feedback loop. The most direct method of determining R/R_o is to measure the signal-to-noise ratio always in the widest bandwidth transmitted, regardless of the spectral occupancy of the signal.

If this widest bandwidth is B_{max}, then the signal-to-noise ratio measured in B_{max} must be multiplied (amplified) by a fixed amount $A = B_{max}/B_o$ before transmission over the feedback link.

In tropospheric scatter, the fading signal has two components, a fast fading Rayleigh distributed component and an unspecified slow fade. The action of the tape machine as outlined above will compensate for the fast fades. If desired, the slow fades could be compensated for by controlling the transmitter power in a manner similar to that suggested by Axelby and Osbourne[3].

In addition, it should be noted that while the signal presented to the demultiplexer input has a constant signal-to-noise ratio, the signal level is not constant but varies with the fading. It then appears highly desirable from a practical standpoint to place an AGC circuit ahead of the demultiplexing operation.

SYSTEM DELAY

Tape loops introduce a delay into the system; for example, consider the transmitter tape loop described previously. Speed of the tape at the record head is fixed at V_o while the speed of the tape at the playback is variable at V_R. In a time Δt, the net buildup of information between record and playback heads is given by $I\delta$, where: $I\delta = (V_o - V_R)\ \Delta t$. It will be positive if $V_o > V_R$ and some information is stored in the tape; it will be negative if $V_o < V_R$ and some information previously stored is withdrawn from the tape.

The total delay introduced into the system is equal to:

$$\delta = \frac{I\delta}{V_o} = (1 - \frac{V_R}{V_o})\ \Delta t$$

And, the cumulative delay for an interval $t_2 - t_1$ is given by:

$$\delta = \int_{t_1}^{t_2} (1 - \frac{V_R}{V_o})\ dt = \int_{u}^{t} (1 - \frac{R_{(u)}}{R_o})\ du$$

While the delay at any given time cannot be determined, the probability of the system delay exceeding a given amount of tape storage can be determined to a good approximation. Calculations show that the probability of exceeding a storage capacity is given by:

$$P(S > C) = 1 - H\left(\frac{C}{\sqrt{2}}\ \delta\right)$$

Where:

$$H(x) = \sqrt{\frac{2}{\pi}} \int_{o}^{x} \exp(-\beta^2)\, d\beta$$

and $\alpha = (f'/2)$ ln 2 and f' = half-power point of the fading mechanism's power spectrum.

Fig. 7 shows the probability of exceeding various storage capacities as functions of operating time for the case of no diversity.

CONCLUSIONS

The variable-data-rate method of operation described in this article is practical for tropospheric scatter communications and for other systems as well, including satellite communications. The technique promises to reduce interference and primary power consumption significantly by reducing the average transmitter power necessary for a reliability equal to that of a conventional fixed-rate system. Variable data-rate permits extension of the tropospheric scatter system to geographic areas now "out-of-range" and makes it possible to locate more stations within a given area, thereby conserving the radio spectrum.

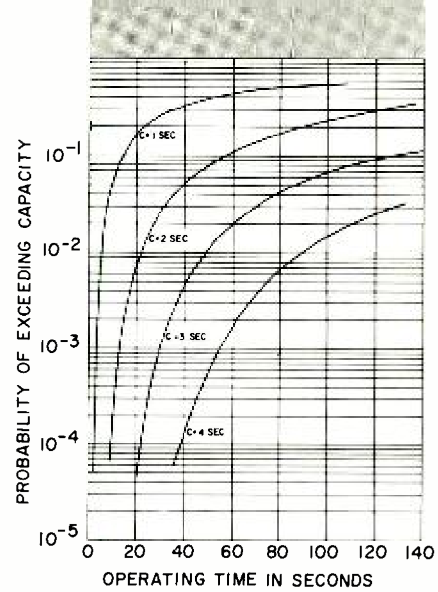


Fig. 7—Probability of exceeding a given storage capacity, C, as a function of operating time.

ACKNOWLEDGEMENT

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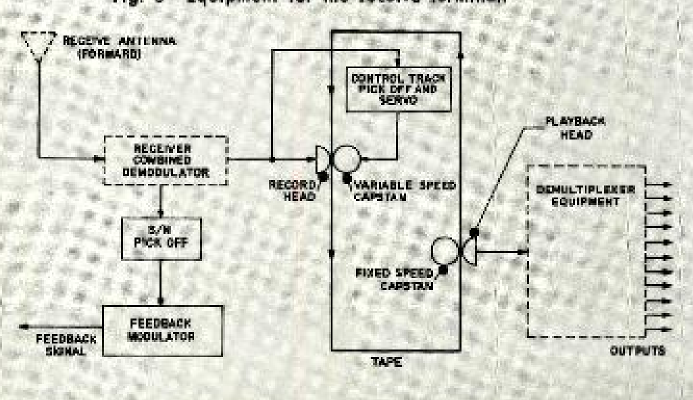
Fig. 6—Equipment for the receive terminal.



J. E. PALMER received a BSc (Math and Physics) from the University of Western Ontario in 1954 and joined RCA as a specialized trainee. He left RCA to serve two years in the U.S. Military Service and then rejoined to RCA in DEP Applied Research. He worked on the design and development of digital data processing and communication equipment and on the theoretical analysis of digital communication systems using wire lines and tropospheric scatter links. In 1960, Mr. Palmer received an MSEE degree from the University of Pennsylvania as a member of the RCA graduate study program. Mr. Palmer recently left RCA to join the Department of Pure and Applied Science of Gannon College, Erie, Pennsylvania.

Engineering and Research NOTES

BRIEF TECHNICAL PAPERS OF CURRENT INTEREST

Sun-Pumped Continuous Laser



by R. C. Duncan, Jr., Dr. Z. J. Kiss, and Dr. H. R. Lewis
RCA Laboratories, Princeton, New Jersey

Laser (optical-maser) action has been achieved in $CaF_2:Dy^{2+}$ at liquid neon temperature (27°K) using the sun as the pumping source. The minimum power required to obtain laser oscillations could be supplied with a 10-inch-diameter condensing mirror. Laser action at liquidnitrogen temperature is anticipated using a 20-inch-diameter condensing mirror.

Laser action in the $CaF_2:Dy^{2+}$ system was reported[1] at 2.36 microns. The laser oscillation takes place in the sharp $^5I_7 \rightarrow {}^5I_8$ $4f$-$4f$ transitions, and it is pumped in broad $4f$—$5d$ absorption bands starting at 10,000 cm^{-1} and extending throughout the visible region of the spectrum. The low pulsed laser threshold, the long lifetime (10 msec for a 0.05 molar % Dy^{2+} in CaF_2) and the convenient location of the broad pumping bands of this system make it especially suitable for sun-pumped operation.

Fig. 1 shows the experimental arrangement. A 1-inch long, ¼-inch-by-⅛-inch cross-section CaF_2: 0.05 M % Dy^{2+} laser crystal is placed in a dewar filled with liquid neon just outside the focal point of a 14-inch spherical mirror. The dewar was wrapped with aluminum foil except for the area of illumination to insure better optical coupling. The output laser beam was chopped, fed into a Perkin-Elmer monochrometer, and recorded with a PbSe detector having a time-constant of 15 msec. The laser output is shown on Fig. 2. The lower trace of Figure 2a is the scope trace when the monochrometer was set just off the laser line to give some idea of the chopped scattered light and the noise level of the trace. The upper trace is the laser output showing the relaxation oscillations. The trace was taken using the full 14-inch aperture of the mirror. The laser threshold was observed with a 10-inch-diameter portion of the mirror exposed.

To estimate the required power of the laser threshold, we assume[2] the power density from the sun to be 100 milliwatts/cm^2 (a "fairly" clear August afternoon at 3:30 p.m. in Princeton, N. J.). Assuming 80% reflectivity, the pump power collected by the 10-inch mirror is about 50 watts. Considering the relatively poor optical coupling, and matching the absorption spectra of the crystal to the emission curve of the sun, we estimate the power absorbed by the crystal at threshold to be about 3 watts. From the known values of the pulsed laser threshold at 27°K and at 78°K, we estimate that a 20-inch-diameter condensing mirror will be sufficient to operate the laser at liquid nitrogen temperature. Experiments using much larger mirrors are in progress to evaluate the high power output capabilities of the sun-pumped laser.

One possible use of a sun-pumped laser is in a satellite where radiation cooling might be possible.

Acknowledgements: The crystal used in this experiment was grown at RCA Laboratories by Harold Temple. We acknowledge with pleasure the assistance of R. B. Marotte with the experiments. The research reported in this work was partially supported by the Electron Technology Laboratory, Aeronautical Systems Division, Air Force Systems Command, Wright-Patterson Air Force Base, Ohio under Contract Number AF33(616)-8199, and the RCA Laboratories, Princeton, N. J.

1. Z. J. Kiss and R. C. Duncan, Jr., *Proc. IRE*, 50, 1531 (1962).
2. Z. J. Kiss, *to be published*.

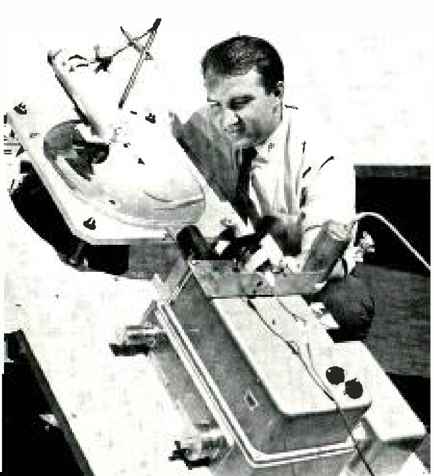


Fig. 1—Experimental arrangement showing the liquid neon dewar, 14" spherical mirror, 30-cps beam chopper, monochrometer, and PbSe detector; Dr. Kiss observes.

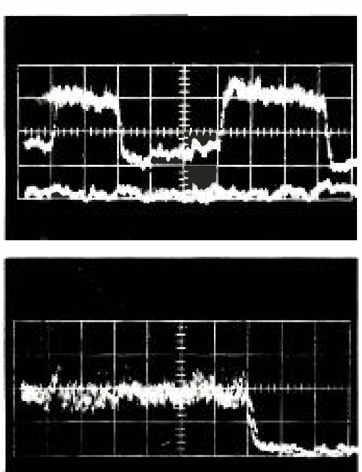


Fig. 2—Oscilloscope trace of sun pumped $CaF_2:Dy^{2+}$ CW laser output at 2.36μ. The laser beam is mechanically chopped at 30 cps. (a) (top) Time scale is 5 msec/cm. In the lower trace, monochrometer slightly off the laser frequency, showing noise level and low chopped scattered light intensity. In the upper trace, monochrometer is on the laser frequency, showing 30 cps chopped signal and laser relaxation oscillations. (b) (bottom) Laser output with time scale of 2 msec/cm showing the relaxation oscillations more clearly.

Wideband Microwave Phototubes for Laser Communications Systems

D. J. Blattner

H. C. Johnson

Dr. Sterzer

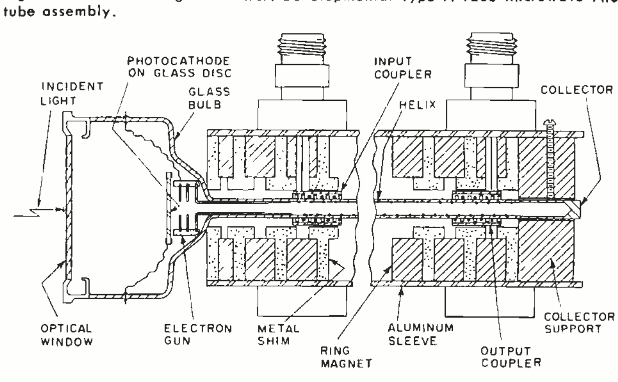
Dr. Reudy

by D. J. Blattner, H. C. Johnson, Dr. F. Sterzer (*Microwave Applied Research Lab.*) and Dr. J. E. Reudy (*Conversion Devices Lab.*), *Electron Tube Division, Princeton, N. J.*

The frequency response of conventional photomultiplier tubes is limited to a few hundred megacycles.[1] With the advent of lasers and the consequent effort to develop optical communication systems having thousands of megacycles of bandwidth, a need has arisen for wideband phototubes. The Electron Tube Division is developing such microwave phototubes under Signal Corps Contract DA36-039-SC-90846, and is prepared to furnish similar tubes to other RCA divisions.

Fig. 1 is a schematic diagram of the RCA developmental type A-1283 Microwave Phototube. The light to be demodulated passes through the optical window onto a transmission type photocathode. The photoelectrons emitted by the cathode are bunched at the modulation frequency of the light. As these electrons pass through a traveling-wave-tube type helix, they excite a traveling wave that is taken out at the output coupler. The tube, and its focusing periodic permanent magnets, is 18 inches long and weighs 5 pounds. The transmission-type photocathode can be made to provide any response available in conventional RCA phototubes.[2] Present tubes use an L-band helix (800 to 2200 Mc); however, tubes can also be built having S-, C-, *or* X-band helices. A photograph of the tube is shown in Fig. 2.

Fig. 1—Schematic diagram of RCA Developmental Type A-1283 Microwave Phototube assembly.



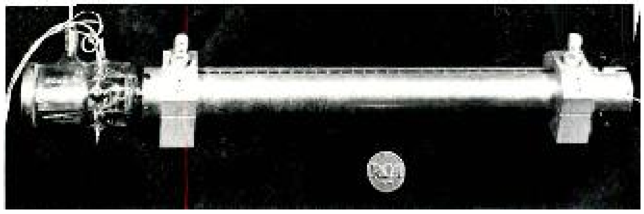


Fig. 2—Photograph of RCA Developmental Type A-1283 Microwave Phototube.

The RCA A-1283 has two important advantages over the microwave phototube previously described by Siegman, *et al.*[3] Siegman's tube used a thermionic oxide cathode as the photoemitter. Thus, the incoming light had to be incident on the front of the cathode, complicating the optical alignment of the tube. Moreover, the photosensitivity of a thermionic oxide cathode is orders of magnitude less than that of the transmission type photocathode.

Fig. 3 shows some of the ways in which the microwave phototube can be used, as follows:

1) When the light incident on the photocathode is modulated at a frequency f_s which is in the passband of the helix, the RF output from the tube is also at f_s.

2) When the light incident on the cathode consists of a laser beam having two or more modes at light frequencies $f_1, f_2 \cdots$, where $(f_1 - f_2) = (f_2 - f_3) = \cdots = f_s$, and f_s is in the passband of the helix, the RF output from the phototube is at frequency f_s. Employed in this way, the microwave phototube is a valuable tool for studying the mode spectra of lasers.

3) When two laser beams, having light frequencies f_L and f_{LO} and a difference $|f_{LO} - f_L|$ that is in the passband of the helix, are incident on the photocathode, the RF output is at frequency $|f_{LO} - f_L|$. In optical communication systems, the light beam at f_L is used to carry the information, and the light beam f_{LO} is a local oscillator supplied at the receiver.

4) When the incident light beam is modulated at a frequency f_s that is outside the frequency range of the helix, the tube can still be used to demodulate the signal by supplying an RF local oscillator frequency f_{LO} at the input of the helix, provided f_{LO} and $|f_s \pm f_{LO}|$ are in the operating range of the helix. The RF output of the helix will contain frequencies f_{LO} and $|f_s \pm f_{LO}|$, and f_s can be recovered by simply passing the output into a crystal mixer. *This novel technique greatly extends the useful frequency range of the microwave phototube.* For example, a tube having a helix that operates in the range from 1 to 2 Gc can be used to demodulate signals ranging from 0 to 4 Gc by this technique. It should be noted that f_s need

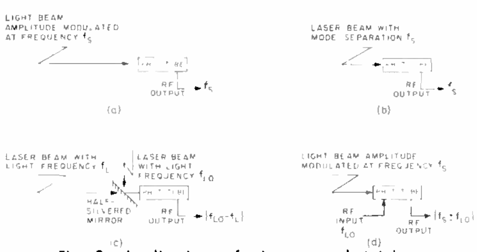


Fig. 3—Applications of microwave phototubes.

not be simply the modulation on a single light beam, but can be obtained by mixing together two laser beams, as in (2) or (3) above.

The helix voltage for the L-band tube ranges from 300 to 350 volts, and the collector can be operated at this same potential. Typical beam currents range from 0.01 to 20 μa; the light intensity required to produce these currents depends on both the wavelength of the incident light and the response of the photocathode. For example, an S-20 cathode can provide a photocurrent of approximately 60 μa/mw of light at 5000 angstroms.

The signal-to-noise ratio of the microwave phototube is evaluated elsewhere in this issue.[4] The analysis shows that for weak signal conditions the signal-to-noise ratio of a receiver using optical heterodyning (Fig. 3c) is significantly better than the signal-to-noise ratio of a video receiver (Fig. 3a).

Acknowledgment: The authors are grateful to H. E. McCandless and D. B. Wenner for building tubes, and to Dr. G. A. Morton and F. E. Vaccaro for many helpful suggestions and discussions.

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AM to PM Conversion in Tunnel Diode Amplifiers

by R. M. KURZROK, *Systems Laboratory, Surface Communications Division, DEP, New York City, N. Y.*

In RF and IF amplifiers, AM to PM conversion can be a source of distortion in frequency-modulated systems.[1] This *Note* discusses AM to PM conversion in tunnel-diode amplifiers.

A typical germanium tunnel diode *I-V* characteristic can be approximated by[2]:

$$i = G_o v - g v^2$$

Where: i and v represent the current and voltage variations, respectively, about the bias point (I_o, V_o). The conductances G_o (at the bias point) and g (at any point of interest α) are:

$$G_o = \frac{\partial i}{\partial v}\Bigg|_{I_o, V_o} \qquad g = \frac{G_o - \frac{\partial i}{\partial v}}{2V_\alpha}\Bigg|_\alpha$$

It can be shown[3] that the rectified voltage V_R due to an AC signal input voltage $v = V_p \cos \omega t$ will be:

$$\left| V_R \right| = \frac{1}{2} g V_p^2. \tag{1}$$

Although the third-order nonlinearity is usually more significant than the second-order nonlinearity when operating in the negative-resistance region of a tunnel diode, this third-order linearity can be neglected, since it will not contribute to the rectified voltage V_R.

For a hybrid-coupled or circulator-coupled tunnel-diode amplifier, it can be shown[4,5] that amplifier gain, G (in db), is:

$$G = 10 \log_{10} \left\{ \frac{(1+n)^2 + X^2}{(1-n)^2 + X^2} \right\} \tag{2}$$

And, amplifier phase shift, θ, is:

$$\theta = \tan^{-1}\left(\frac{-X}{1+n}\right) - \tan^{-1}\left(\frac{X}{1-n}\right) \tag{3}$$

Where: $n = Z_o \div |-R|$; $X = \omega C Z_o$; Z_o = characteristic impedance of source/load; $-R$ and C are circuit elements of the tunnel diode; and ω = angular frequency. (It is assumed that *ideal* hybrids or circulators have been employed; lead inductance and spreading resistance of the tunnel diodes have been neglected.)

For small values of θ, Equation 3 reduces to:

$$|\theta| = \frac{2X}{1-n^2} = \frac{2\omega C Z_o}{1 - \frac{Z_o^2}{R^2}}$$

Differentiating $|\theta|$ with respect to C and replacing $d\theta$ and dC by $\Delta\theta$ and ΔC:

$$\Delta\theta = \frac{2\omega Z_o}{1 - \frac{Z_o^2}{R^2}} \Delta C \tag{4}$$

For typical values of $n = 1/2$, $|R| = 100$ ohms, and $Z_o = 50$ ohms:

$$\Delta\theta \cong \omega R \Delta C \tag{5}$$

Equations 4 and 5 relate small changes in the phase shift through tunnel-diode amplifiers to small changes in the tunnel-diode junction capacitance.

The junction capacitance of the tunnel-diode amplifier is a function of the applied voltage. Theoretically, this dependence of capacitance upon voltage may be approximated by[6] (K and V are constants):

$$C \cong K(\phi - v)^{-1/2} \tag{6}$$

Differentiating C with respect to v, replacing dC and dv by ΔC and Δv, and letting $v = V_o$:

$$\Delta C \cong \frac{K}{2}(\phi - V_o)^{-3/2} \Delta V \tag{7}$$

Now, ΔV represents the change in voltage from the establishing operating potential V_o. Consequently, ΔV will be equal to the rectified voltage V_R because of the detection of a high-level RF signal in the tunnel-diode amplifier.

Combining Equations 1, 5, and 7:

$$\Delta\theta \cong \frac{\omega R K g}{4} (\phi - V_o)^{-3/2} V_p^2 \qquad (8)$$

For a 1-milliampere-peak-current germanium tunnel diode, biased in its negative-resistance region, typical values of the constants are: $K = 1.4 \times 10^{-12}$ (volt)$^{1/2}$ farad, $\phi = 0.58$ volts, $V_o = 0.13$ volts, and $g = 5 \times 10^{-3}$ μmhos/volt. Letting $\omega = 2\pi f = 2\pi$ (10^9) cps, R = 50 ohms, and $V_p = 0.1$ volt:

$$\Delta\theta = 3.64 \times 10^{-5} \text{ radians}$$

For the assumed operating conditions, the AM to PM conversion in the tunnel-diode amplifier is negligible for most applications. Although several approximations have been used in deriving Equation 6, this equation is convenient for obtaining a "ballpark" computation of the phase jitter in a tunnel-diode amplifier due to AM detection of the applied RF signal. This is a measure of AM to PM conversion within the tunnel-diode amplifier.

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New Electronic Direct-Reading Hygrometer Has Calibrated Output

by Loy E. Barton, *RCA Laboratories, Princeton, N. J.*

Accurate and fast acting electronic thermometers, using transistors or diodes as temperature-sensing elements, were developed some time ago at RCA Laboratories. In a new development, two of these transistor thermometers have been combined to provide an electronic hygrometer for the direct reading of relative humidity.

Unlike other hygrometers, this hygrometer has an electrical output of about 1/2 millivolt across 1000 ohms per 1-percent relative humidity which may be applied to a recorder, or to a device to maintain a given relative humidity.

This hygrometer is easy to calibrate in that no relative humidity reference standard is needed. In the process of assembly of the instrument, the meter is left out but remains electrically connected to the circuit in order that the device may be turned upside down to permit submerging the sensing elements in water of known temperatures. The sensing elements are at the top of the two pedestals shown in Fig. 1. In operation, the sensor at the left is exposed to ambient temperatures and the one on the right is wick wetted to measure the depressed temperature of the "wet bulb."

The sensor at the left also has thermally in contact with it a diode that maintains equal currents to the two separated transistor sensors when the transistors are at equal temperatures. Therefore, the meter reads zero when the two sensors are at the same temperature for a range of temperatures, such as 0° to 40°C.

Fig. 1—Electronic hygrometer.

Now, if the left sensor (which may be considered the ambient or dry bulb) is put into water at a given temperature, and the right sensor placed in water at a lower or "wet bulb" temperature, the meter will read the difference temperature and may be calibrated in degrees difference. This difference represents a given percentage of relative humidity at a given ambient temperature. Therefore, if we wish to use a single linear scale on the meter for indicating relative humidity over a desired temperature range, some compensation is required. Since the difference temperature between the wet and the dry bulb sensors for a given ambient is greater at higher temperatures, a second diode must be thermally associated with the sensor at the left to render the instrument less sensitive as the ambient temperature rises. The compensation thus afforded by the second diode permits the use of a single linear meter scale for indicating percent of relative humidity over a selected temperature range.

The circuits and adjustments in the hygrometer provide for an ambient range of 15°C to 40°C or about 60°F to 100°F. Full-scale deflection of the meter without the linearity diode is about 11°C. The accuracy is about 2 percent but depends upon the accuracy of the various water temperatures used in calibration and the accuracy of the compensating adjustments.

An Improved Method of Measuring the Complex Dielectric Constant of Ferrite-Like Toroids



by R. L. Harvey, Dr. I. Gordon, and R. A. Braden

RCA Laboratories, Princeton, N. J.

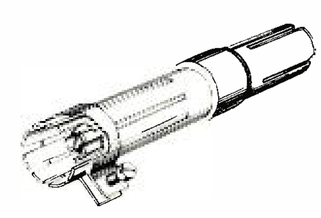
The measurement of complex dielectric constant at frequencies of 200 Mc to 3000 Mc is usually performed by placing the specimen in the maximum voltage position of a concentric line that is attached to a slotted line device. The real component (ϵ') is determined by the shift in the voltage minimum and the imaginary component (ϵ'') is determined by the change in the voltage-standing-ratio. The literature on this type of measurement states that most workers used samples of solids or liquids that were made to fit exactly the concentric line. In this case, a size correction is not necessary.

To determine the dielectric properties of a large number of ceramic-like toroids, such as ferrite cores, it is impractical and expensive to grind each specimen to a definite size. If the toroids are made to fit loosely into the concentric line, a size correction can be made[1] that involves the measured dielectric constant, the dimensions of the concentric line, and the dimensions of the specimen. We find that it is almost impossible to measure the specimen with sufficient accuracy (especially if it is slightly elliptical) to obtain satisfactory accuracy in the final result.

Because of these problems, a simple holder (Fig. 1) for the sample was devised that circumvents all these difficulties. It consists of an open-ended line that can be attached to the slotted line measuring equipment. The inner and outer conductors are made with a number of longitudinal slots at the end that is remote from the measuring equipment. This is a position of maximum voltage and the position where the sample is inserted.

The sample to be measured is given a conductive coating on its inner- and outer-diameter surfaces (e.g., one of the commercially available silver pastes). The coated core is then inserted in the holder. The inner conductor is expanded by the center screw to fit the inside of the sample; the outer conductor is tightened by the clamp around the outside of the sample. Under these conditions, no size correction is needed, and we find that the true dielectric properties can be measured within an accuracy of at least 5%.

Fig. 1—An adjustable sample holder for toroidal samples which have small variations in dimensions. The holder obviates size correction when measuring the complex dielectric constant.



The dimensions of the sample and the dimensions of the concentric line should be such that there is only an insignificant change in the impedance of the line when it is adjusted to fit the sample.

Acknowledgement: The research reported in this work was sponsored by the U.S. Army Signal Research and Development Laboratories, Fort Monmouth, New Jersey, under contract number DA36-039-sc-87433.

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Superconducting Ribbon for Solenoid Applications

by Dr. J. J. Hanak, *RCA Laboratories, Princeton, N. J.*

Very-high-field solenoids capable of generating fields of 100,000 gauss now made with copper winding require about 100 tons of equipment and dissipate more than one megawatt of power as heat. Superconductors have been long considered for the construction of light-weight solenoids, because at cryogenic temperatures they can carry large electric currents with zero power dissipation. However, most superconductors are unsuitable for this application, since they revert to the normal state when placed in low or moderate magnetic fields. Recently, it has been reported that certain alloys and compounds retain superconductivity at high magnetic fields.[1] Important in this connection is the compound niobium stannide, Nb_3Sn, which was found to remain superconducting at fields over 200,000 gauss.

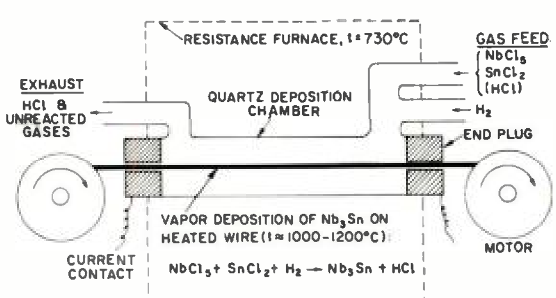
In the past, Nb_3Sn was prepared by metallurgical sintering techniques, which resulted in a porous and extremely brittle material not suitable for widespread use. In 1960, scientists in our Materials Research Laboratory developed a vapor-phase transport process for preparing this compound for the first time in a dense crystalline state—and *in forms suitable for widespread use in both research and application.* It consists of a simultaneous reduction of gaseous mixed chlorides of niobium and tin by hydrogen at 900 to 1200°C.[2]

Based on this process, an apparatus (Fig. 1) was developed for continuous coating of refractory metal wire and ribbon with Nb_3Sn. Noble metals, including rhodium, osmium, iridium, platinum, and gold, are especially suitable as substrate materials because they promote nucleation of Nb_3Sn. The Nb_3Sn coated ribbon has both electrical and mechanical properties desirable for solenoid construction. It is very thin (typical cross section is 2 × 30 mil, thickness of deposit about 0.3 mil) and hence sufficiently ductile to wrap around diameters as small as 1 inch and it can support enormous currents densities: 1×10^6 amp/cm^2 at zero field, 5×10^5 amp/cm^2 in a transverse DC field of 93,500 gauss, and 1.5×10^5 amp/cm^2 in a pulsed longitudinal field of 170,000 gauss. By comparison, copper can carry only 1×10^3 amp/cm^2 safely. Hence, superconducting solenoids approaching a field of 200,000 gauss appear feasible. As a first step toward high-field magnets, a superconducting solenoid, 2 inches in diameter and generating 10,300 gauss, has been built in the Materials Research Laboratory with 450 meters of this ribbon. It is noteworthy that the field of this magnet corresponded closely to what was expected from the measured current capacity of the ribbon.

Vapor-deposited Nb_3Sn will be useful not only for high field superconducting magnets, but also for other electronic applications. Examples are the fabrication of lossless microwave cavities, at high frequencies superconducting-rotor gyroscopes, and the construction of AC magnets for use in AC magneto-hydrodynamic energy conversion. These applications will be facilitated by coating of Nb_3Sn on ceramics, also successfully demonstrated at the RCA Laboratories.

Because of its device potential, the ribbon-coating process attracted the attention of the Electron Tube Division, Harrison, which has now assumed the responsibility for commercial development of the process. As a part of this activity, ETD recently established at David Sarnoff Research Center a new group known as the Superconductor Materials and Devices Laboratory. This project is being funded by ETD's New Business Development activity. In addition to further research and development of the process, this Laboratory is also building a sizeable 50,000-gauss solenoid.

Fig. 1—Continuous vapor deposition of niobium stannide wire or ribbon.



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Spilled Milk Belongs Down The Drain

by L. R. Sheppard, *Aerospace Communications and Controls Division, DEP, Camden, N. J.*

It is frustrating when a carefully thought out change proposal is rejected with a comment like, "We've spent $23,471.50 for tools for the present configuration and though it may be cheaper to produce the design you propose—we've got this large investment—etc., etc.

To forestall such frustrations, we have included the following in the *RCA Value Engineering Manual:*

> "Not so obvious, however, is the apparently contradictory statement that once these expenditures" (i.e. investment in fixed costs to produce the original design) "are made, they should be considered as *spilled milk* when evaluating the adviseability of a further change—This is money spent—it cannot be recovered, and if the cost comparisons show that there is a net saving in going to the new method—all we can do is fall into line in the march of progress, perhaps shedding a tear or two over the spilled milk enroute."

Far from having the desired effect, however, this gave rise to protests from previously unheard-from quarters where the manual had been read. What we were saying in a nutshell was: *The money's been spent! Forget it!*

Intuitively, the objectors felt that when an investment of sizeable magnitude must be amortized at almost any cost. Equally intuitively we feel that the only useful function of past errors is to serve as a reminder to prevent repeating them in the future. After proving this point by calculating out specific examples each time they arose, we formulated a general proof of the proposition, as follows.

Let: Q_A = Quantity of parts completed before change can be implemented. Q_B = Remaining quantity of parts to be made by proposed method, if accepted. T_A = Tooling and other nonrecurring costs of producing by present method. T_B = Tooling and nonrecurring costs for instituting the proposed change. (This must include such factors as ECN, retrofit, stock updating, manual and spare parts list changes and similar costs as appropriate.) L_A = Recurring cost per piece, original method. L_B = Recurring cost per piece, proposed method. If no change is made, the lot will cost:

$$(Q_A + Q_B)\left(\frac{T_A}{Q_A + Q_B} + L_A\right) \qquad (1)$$

If Q_A is made by the old method and Q_B by the new method, the lot will cost:

$$Q_A\left(\frac{T_A}{Q_A} + L_A\right) + Q_B\left(\frac{T_B}{Q_B} + L_B\right) \qquad (2)$$

To yield a savings (1) must be greater than (2), setting up this inequality and simplifying, we find:

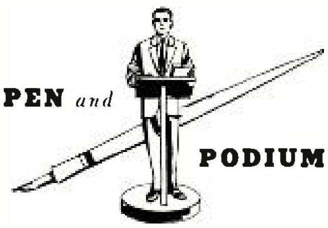
$$L_A Q_B > T_B + L_B Q_B \qquad (3)$$

Note that the spilled milk, T_A, has disappeared down the drain, proving our point.

To simplify use of this inequality in evaluating the advisability of a proposed change, we can divide through by Q_B to go to a per-piece basis:

$$L_A > L_B + \frac{T_B}{Q_B};\ T_B < Q_B(L_A - L_B)$$

If we can satisfy this inequality—*if the nonrecurring costs of going to the new method are less than the product of the quantity to be made times the per-piece saving*—let's seat the stopper firmly in the drain and go to the new method.



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DATES and DEADLINES

PROFESSIONAL MEETINGS AND CALLS FOR PAPERS

In January 1963, the IRE and the AIEE merged to form the Institute of Electrical and Electronics Engineers, Inc.—the IEEE. This has been reflected below in noting the sponsorship of conferences. Also, note that former IRE Professional Groups (PG's) and AIEE Technical Groups (TG's) are now called "Professional Technical Groups," or PTG's.

Meetings

March 15-16, 1963: Pacific Computer Conf., IEEE; Calif. Inst. of Technology, Pasadena, Calif. *Prog. Info.:* E. J. Schubert, 2400 Harbor Blvd., Fullerton, Calif.

March 18-20, 1963: American Rocket Soc. and Inst. of the Aerospace Sciences Space Flight Testing Conf.; Cocoa Beach, Fla. *Prog. Info.:* Dr. Charles L. Carroll, Jr., Mgr., Technical Staff, Guided Missiles Range Div., Pan American World Airways, Inc., Patrick Air Force Base, Fla.

March 25-28, 1963: IEEE Intl. Convention, all PTG's; Coliseum and New York Hilton, New York, N. Y. *Prog. Info.:* D. B. Sinclair, IEEE Hdqs., 1 East 79th St., New York, N. Y.

March 26-28, 1963: American Power Conference, Ill. Inst. of Tech., Co-Operating IEEE Tech. Grps.; Sherman Hotel, Chicago, Ill. *Prog. Info.:* W. A. Lewis, Ill. Inst. of Technology, Chicago, Ill.

April 9-10, 1963: 6th Biennial Electric Heating Conf., IEEE; Penn-Sheraton Hotel, Pittsburgh, Pa. *Prog. Info.:* R. A. Summer, Ohio Crankshaft Co., 3800 Harvard Ave., Cleveland, Ohio.

April 10-11, 1963: 4th Symp. on Engineering Aspects of Magnetohydrodynamics, IEEE-Univ. of Calif., LAS; Univ. of Calif., Berkeley, Calif. *Prog. Info.:* G. S. Jones, Avco Everett Research Lab., 2385 Revere Beach Parkway, Everett 49, Mass.

April 14-18, 1963: The Electrochemical Soc., Inc., Penn-Sheraton Hotel, Pittsburgh, Pa. *Prog. Info.:* Society Hdqs., 30 E. 42nd St., New York, N. Y.

April 16-18, 1963: Cleveland Electronics Conf., Cleveland Section, IEEE; Hotel Sheraton-Cleveland, Cleveland, Ohio. *Prog. Info.:* Roland Gogolick, S. Sterling Co., 5827 Mayfield Rd., Cleveland 24, Ohio.

April 16-18, 1963: Optical Masers Symp., PIB-IEEE; United Eng. Center, New York, N. Y. *Prog. Info.:* IEEE Hdqs., 1 East 79th St., New York, N. Y.

April 17-19, 1963: Intl. Conf. on Nonlinear Magnetics, IEEE; Shoreham Hotel, Washington, D. C. *Prog. Info.:* J. J. Suozzi, BTL Labs., Allentown, Pa.

April 17-19, 1963: Southwestern IEEE Conf. and Elec. Show. Dallas Memorial And., Dallas, Texas. *Prog. Info.:* Al Mitchell Graduate Res. Center of the S.W., P.O. Box 8478, Dallas 5. Texas.

April 22-23, 1963: Rubber and Plastics Conf., Rubber and Plastics Indus. Committee; Sheraton Hotel, Akron, Ohio. *Prog. Info.:* R. Schofield, Cutler-Hammer, Inc., 1521 W. 117th St., Cleveland, Ohio.

April 22-24, 1963: 3rd Ann. San Diego Symp. for Biomedical Engineering, IEEE, et al; Del Webb's Ocean House, San Diego, Calif. *Prog. Info.:* John H. McLeod, 8484 La Jolla Shores Dr., La Jolla, Calif.

April 24-26, 1963: Power Industry Computer Conf., IEEE; Westward-Ho Hotel. Phoenix, Ariz. *Prog. Info.:* G. W. Stagg, American Elec. Pr. Serv. Co., New York, N. Y.

April 29-30, 1963: Electro-Nuclear Conf., IEEE; Richland, Wash. *Prog. Info.:* E. J. Mollerus, 1218 Geo. Washington Way, Richland. Wash.

April 29-May 2, 1963: 1963 Spring URSI-IEEE Mtg., URSI, PTGAP, CT, GE, IT, I, MT, and T; Natl. Academy of Sciences, Washington, D. C. *Prog. Info.:* Dr. Millett G. Morgan, Thayer School of Eng., Dartmouth College, Hanover, N. H.

May 2-3, 1963: 4th Natl. Symp. on Human Factors in Electronics, PTGHFE; Marriott Twin Bridges, Washington, D. C. *Prog. Info.:* Rube Chernikoff, U. S. Naval Res. Lab. Code 5124, Wash. 25, D. C.

May 7-9, 1963: Electronic Components Conf., IEEE-EIA; Intl. Inn, Wash., D. C. *Prog. Info.:* E. J. Kaputa, Bu. of Ships, Code 681A2, Dept. of Navy, Wash. 5, D. C.

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May 17-18, 1963: Symp. on Artificial Control of Biology Systems, PTGBME-Buffalo Niagara Section; Univ. of Buffalo, School of Medicine, Buffalo. N. Y. *Prog. Info.:* D. P. Sante, 4530 Greenbriar Rd., Williamsville 21, N. Y.

May 20-22, 1963: Natl. Symp. on Microwave Theory and Techniques, PTGMTT; Miramar Hotel, Santa Monica, Calif. *Prog. Info.:* Irving Kaufman, Space Tech. Labs., Inc., 1 Space Park, Redono Beach, Calif.

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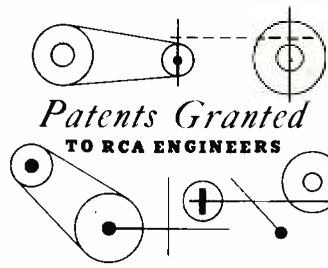

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3,060,422—Indicator and/or Control System, Oct. 23, 1962; R. D. Kell

Des. 193,692—Photographic Copying Apparatus, Oct. 23, 1962; C. J. Young

3,061,657—Thermoelectric Compositions and Devices Utilizing Them, Oct. 30, 1962; E. F. Hockings

3,061,683—Gain-Control Circuit for Stereophonic Radio Receivers, Oct. 30, 1962; L. A. Freedman and J. O. Preisig

3,062,970 — Converter Circuits Employing Negative Resistance Elements, Nov. 6, 1962; K. Li

3,064,133—Layer Type Storage Light Amplifier, Nov. 13, 1962; H. O. Hook and J. Murr, Jr.

3,064,134—Display Device, Nov. 13, 1962; R. D. Kell

3,064,154—Cathode Ray Tube, Nov. 13, 1962; H. B. Law

3,064,210—Harmonic Generator, Nov. 13, 1962; M. C. Steele and A. L. Eichenbaum

3,065,392—Semiconductor Devices, Nov. 20, 1962; J. I. Pankove

3,067,348—Pickup Tube Target Structure, Dec. 4, 1962; S. A. Ochs

3,068,360—Radar Light Amplifier Device, Dec. 11, 1962; F. H. Nicoll

3,068,360—Radar Light Amplifier Device, Dec. 11, 1962; F. H. Nicoll

3,068,414—Radar with Traveling-Wave Duplexer, Dec. 11, 1962; R. W. Peter

3,069,632—Parametric Oscillator Random Number Generator, Dec. 18, 1962; F. Sterzer

3,070,441—Art of Manufacturing Cathode-Ray Tubes of the Focus-Mask Variety, Dec. 25, 1962; J. W. Schwartz

3,070,520—Semiconductor Devices and Methods of Fabricating Them, Dec. 25, 1962; L. Pensak

3,070,654—Chrominance Channel Control Apparatus, Dec. 25, 1962; A. Macovski

3,070,711—Shift Register, Dec. 25, 1962; S. M. Marcus and J. T. Wallmark

3,071,644—Control Systems, Jan. 1, 1963; G. A. Olive

3,071,754—Magnetic Memory Systems Using Transfluxors, Jan. 1, 1963; J. A. Rajchman

3,072,741—Television Brightness and Contrast Control Circuit, Jan. 8, 1963; R. W. Ahrons and L. L. Burns

3,072,751—Magnetic Recording and Reproducing Means, Jan. 8, 1963; V. K. Zworykin and P. K. Weimer

3,073,411—Acoustical Apparatus, Jan. 15, 1963; J. C. Bleazey and J. Preston

3,068,475—Stereophonic Sound Signalling System, Dec. 11, 1962; J. Avins

SEMICONDUCTOR AND MATERIALS DIVISION

3,061,546—Magnetic Cores, Oct. 30, 1962; J. J. Sacco, Jr.

RCA VICTOR HOME INSTRUMENTS DIVISION

3,061,786—Signal Translating Circuit, Oct. 30, 1962; G. E. Theriault

3,061,790—Signal Detectors, Oct. 30, 1962; G. E. Theriault

3,071,639—Portable Radio Receiver Casing and Battery Holder Means, Jan. 1, 1963; C. I. Tighe, Jr. and H. Mendelson

BROADCAST AND COMMUNICATIONS DIVISION

3,061,815—Signal Translating System, Oct. 30, 1962; S. R. Tomes

3,064,075—Method of and Apparatus for Transmission of Television Signals by Frequency Modulation, Nov. 13, 1962; W. L. Hurford

3,064,206—Oscillation Generator and Mixer, Nov. 13, 1962; G. F. Rogers

3,068,338—Thermostatically Controlled Circuits, Dec. 11, 1962; R. R. Bigler

3,073,975—Crystal Unit, Jan. 15, 1963; R. R. Bigler and E. M. Washburn

RCA VICTOR RECORD DIVISION

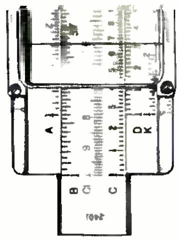
3,068,327—Transistor Amplifier Circuit, Dec. 11, 1962; J. J. Davidson

ELECTRONIC DATA PROCESSING

3,073,971—Pulse Timing Circuit, Jan. 15, 1963; E. J. Daigle, Jr.

3,073,972—Pulse Timing Circuit, Jan. 15, 1963; R. H. Jenkins

Engineering NEWS *and* HIGHLIGHTS



Color, Computers, Space, Broadcasting, and Operating Efficiency Were Important Factors in Record 1962 RCA Sales and Profits

RCA achieved record sales and profits in 1962—the best twelve-month period in its 43-year history. RCA's 1962 sales will approach $1.7 billion and operating profit after taxes will exceed $50 million, gains of at least 10 percent and 40 percent, respectively, over the 1961 levels. The previous RCA earnings record was $47.5 million, in 1955. In addition to operating profit, a capital gain of $7 million was realized in 1962 from the sale of common stock of the Whirlpool Corporation. This nonrecurring income will add $.41 per share of common stock to the operating earnings in 1962.

The strong profit base developed in 1962 has placed RCA in a position to advance to higher sales and earnings in 1963. Significant elements in the company's 1962 success were:

1) A strong upward thrust in consumer products and services, paced by sales of more than twice as many color television sets as in 1961, and by an estimated five-fold increase during 1962 in profits on color apparatus and related services.

2) Growing strength in electronic data processing, reflected in the more than doubling during 1962 of revenue from domestic and international sale and rental of commercial systems, and the continued substantial reduction of related costs.

3) Continued advances in areas of space and defense electronics.

4) Record sales and profits of the National Broadcasting Company.

5) Successful implementation of an intensive companywide program to increase operating efficiencies and reduce costs.

A 30 percent increase over 1961 in total sales of all RCA Victor home instruments exceeded the previous record high (1956). Unit sales of television sets were well in excess of $2 million achieving a dollar volume beyond the previous all-time peak established in 1950.

The electronic data processing program is proceeding toward the development of a profitable growth business. RCA has shipped more than 280 electronic data processing systems to government and commercial users in this country and overseas, and the foreign orders for RCA systems rose to 158, a 125-percent increase over the 1961 year-end total. Also, the first RCA 601 was placed in operation in December 1962 at the New Jersey Bell Telephone Company—and will be featured on the cover of the next issue (April-May 1963) of the RCA Engineer.

NJSPE HONORS RCA

The New Jersey Society of Professional Engineers has selected RCA for its 1963 *Industrial Professional Development Award*. The honor is presented annually to an industrial organization located in New Jersey for outstanding effort in advancing the engineering profession and for recognizing and promoting the interests of its employees in becoming professional engineers. **Arthur L. Malcarney,** Group Executive Vice President, will accept the award at the NJSPE Convention in Newark, N.J., on May 3.

RCA HONORED FOR METEOROLOGICAL SERVICES OF TIROS

RCA was honored recently for *outstanding services to meteorology by a corporation* with the receipt of an award from the American Meteorological Society for the Tiros meteorological satellites, six satellites into operational orbit in six tries.

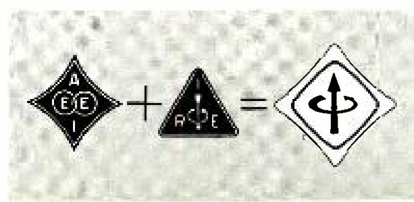
Receiving the award on behalf of RCA was **Barton Kreuzer,** Division Vice President and General Manager of the DEP Astro-Electronics Division—where Tiros was developed and built for NASA.

DR. HAMMOND AWARDED IRE "MEDAL OF HONOR"

Dr. John Hays Hammond, Jr., a Director of RCA since 1923, was recently named recipient of the highest IRE annual technical award in electronics, the *Medal of Honor* —one of two such awards given for 1962. Mr. Hammond, who is President of the Hammond Research Corp., Gloucester, Mass., was honored for "*pioneering contributions to circuit theory and practice, to the radio control of missiles, and to basic communication methods.*" Dr. Hammond is the originator of more than 800 inventions, is a Fellow of the IRE, and has served the IRE in many official capacities.

IRE + AIEE = IEEE

Illustrated below is the genesis of the new emblem recently approved by the Institute of Electrical and Electronics Engineers (IEEE). It will replace emblems of the former AIEE and the IRE, which merged Jan. 1 to form the 150,000-member IEEE. The emblem retains the four-cornered shape of the old AIEE emblem and the symbol itself is a modification of the IRE symbol representing electric current and magnetic force.



S. H. WATSON AWARDED STANDARDS MEDAL

Samuel H. Watson, Manager of Standardizing, Product Engineering, Camden, has been selected by the American Standards Association for its annual *Standards Medal.*

In honoring Mr. Watson, ASA praised him for merging "a broad technical background with a maturing philosophy of standardization to benefit colleagues in his chosen field of electronics as well as related industries." The ASA also cited him for "an exemplary record of accomplishments" in technical society work conducted by the Institute of Radio Engineers and the Standards Engineers Society.

RABINOVICI AND KLAPPER HONORED FOR PAPER

B. Rabinovici, Senior Staff Scientist, and **J. Klapper,** Senior Member Technical Staff, New York Systems Laboratories, DEP Surface Communications Division, were presented on December 6, 1962, with an award at the IRE National Convention on Vehicular Communications in Los Angeles, California, for the paper, "A Synchronous Mode of Radas Communication," published in the *IRE-PGVC Transactions* in August 1962.—*M. P. Rosenthal*

KISOR HONORED FOR PACKAGING

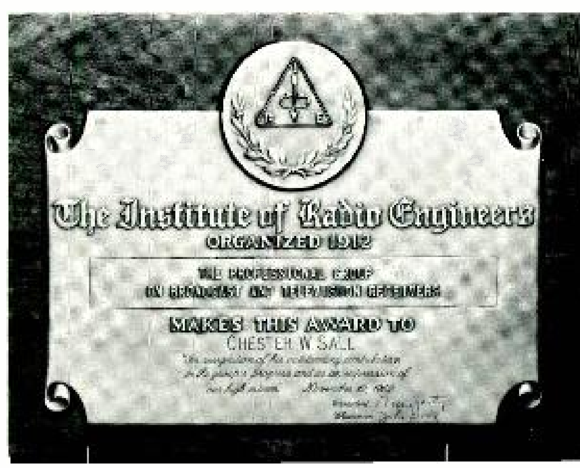
Ted W. Kisor, of the Semiconductor and Materials Division, Somerville, N.J., received the *Harold Jackson Trophy* and a $500 bond for designing the "Best in Show" package at the 1962 National Packaging and Handling Show. The competition was sponsored by the Society of Packaging and Handling Engineers. Mr. Kisor's package is an extruded tube that orients transistors, protects them in a shoulder carton, and provides a tray for convenient handling.

AVENESSIANS CITED FOR OUTSTANDING GRAD WORK

A. Avenessians, of RCA Communications, New York, has been cited for "*outstanding achievement*" as a graduate student in the Columbia University School of Engineering. He was one of three students so honored in a class of 46. He received his MSEE there in June 1962 and is continuing work toward a PhD.—*W. Jackson*

C. W. SALL HONORED BY IRE-PGBTR

Chester W. Sall, Technical Publications Administrator of the RCA Laboratories, was recently honored by the IRE Professional Group on Broadcast and Television Receivers for his "*many contributions to the progress of the organization.*" The handsome plaque shown below was presented to Chet at the Radio Fall Meeting in Toronto. A Senior Member of IRE (now IEEE), Chet is Editor of the *Transactions of the PGBTR.* (He is also the National Chairman of the IEEE Professional-Technical Group on Engineering Writing and Speech.) A graduate of DePauw University and the New York State College of Teachers, Chet has been with RCA since 1943, when he joined the Industry Service Laboratory in New York City. He subsequently transferred to Camden, and came to RCA Laboratories in 1960. In 1961 he was appointed to his present position. He has been especially active as an RCA Engineer Engineering Editor and Editorial Representative for a number of years.



...PROMOTIONS...

to Engineering Leader and Manager

As reported by your Personnel Activity during the past two months. Location and new supervisor appear in parenthesis.

RCA Service Company

C. S. Cummings: from Mgr., Systems Evaluation to *Mgr., Systems Analysis* (N. D. Mallory, Mgr., of Missile Test Project, Quality Analysis)

L. L. Lasman: from Data Reduction Analyst to *Mgr., Mathematical Analysis* (J. R. Garrett, Mgr., Missile Test Project Mathematical Svc.)

E. MacDonald: from Engr., BMEWS, to *Ldr., Engineers* (K. R. Lewis, Mgr., Operations—Home Office)

W. K. McCaskill: from Engr., BMEWS to *Ldr., Engineers* (E. A. Amundson, Mgr., Site II—Clear, Alaska)

H. R. Phillips: from Mgr., Displays, Switching & Data Handling to *Ldr., Engineers* (S. N. Levy, Mgr., Operations—Home Office)

M. Wauters: from Engr., BMEWS, to *Ldr., Engineers* (K. R. Lewis, Mgr., Operations —Home Office)

M. S. Gaspierik: from Engr. to *Ldr., Engineers* (T. G. Rutherford, Mgr., Missile Test Project Telemetry Timing/Firing Engineering)

A. H. Thomas: from Test Operations Analyst to *Ldr., Test Operations Analysts* (M. W. Boggs, Mgr., Missile Test Proj. Tech. Operations)

Electronic Data Processing

J. R. Hammond: from Ldr., D&D Eng. Staff to *Mgr., Peripheral and Sustaining Engineering* (H. Kleinberg, Mgr., Eng., Palm Beach Operations)

C. M. deVillars: from Sr. Member, Eng. Staff to *Ldr., D&D Eng. Staff* (J. R. Hammond, Mgr., Peripheral & Sustaining Eng., West Palm Beach)

T. A. Franks: from Engr. to *Ldr., D&D Engineers* (R. E. Montijo, Mgr., Systems Eng., Camden)

B. Walker: from Engr. to *Ldr., D&D Engineers* (J. A. Burstman, Mgr., Computer Advanced Product Dev., Camden)

Electron Tube Division

B. H. Vine: from Sr. Engr., Prod. Dev. to *Eng. Ldr., Prod. Dev.* (Mgr., Advanced Dev. Eng.—Conversion Tube, Lanc.)

Semiconductor and Materials Division

H. Becke: from Engr. to *Eng. Ldr.* (W. Bosenberg, Mgr., Prod. Dev., Somerville)

E. H. VanWagner: from Eng. Ldr. to *Eng. Ldr.* in Mountaintop.

E. A. Czeck: from Engr., Mfg. to *Eng. Ldr.*, Mfg., Mountaintop.

Astro-Electronics Division, DEP

L. W. Gage: from Engr. to *Ldr., Publication Engineers* (L. Thomas, Mgr., Publication Svcs.)

W. H. Thomas: from Engr. to *Ldr., Publications Engineers* (L. Thomas, Mgr., Publication Services)

D. Y. Moy: from Engr. to *Ldr., Publication Engineers* (L. Thomas, Mgr., Publication Services)

L. Berko: from Sr. Engr. to *Ldr., Engineers* (R. W. Northup Mgr., Mech. Design Development and Integration)

B. L. Matonick: from Engr. to *Ldr., Engineers* (R. W. Northup, Mgr., Mech. Design Development and Integration)

Aerospace Communications and Controls Division, DEP

N. Aron: from Member, Tech. Staff to *Ldr., Technical Staff* (J. H. Woodward, Burlington)

S. N. Richter: from Member, Tech. Staff to *Ldr., Technical Staff* (M. E. Siegal, Burlington)

V. P. Frolich: from Member, Tech. Staff to *Ldr., Technical Staff* (C. W. Bringham, Burlington)

W. R. Hunsicker: from Member, Tech. Staff to *Ldr., Technical Staff* (P. M. Toscano, Burlington)

P. L. Neyhart: from Engr. to *Ldr., Design and Development Engineers* (D. I. Troxel, Mgr., Reliability and Maintainability, Camden)

Applied Research, DEP

F. D. Kell: from Engr. to *Ldr., Design and Development Engineers* (W. R. Isom, Mgr., Electro-Mechanics)

Surface Communications Division, DEP

M. H. Plofker: from Mgr., Value Engineering and Test Coordination to *Mgr., Product Assurance and Reliability* (I. K. Munson, Mgr., Cambridge)

J. Acunis: from Ldr. Tech. Staff to *Mgr., Systems Engineering* (M. L. Ribe, Mgr., Systems Projects, N.Y.)

G. Ashendorf: from Engr. to *Ldr., Engineering Systems Projects* (M. Raphelson, Mgr., Camden)

Data Systems Division, DEP

E. E. Griffith: from Member, Engineering Staff to *Ldr., Systems Engineering Staff* (G. G. Murray, Mgr., Information Processing Systems, Van Nuys)

B. E. Kempf: from Sr. Member, D&D Engineering Staff to *Ldr., D&D Engineering Staff* (E. L. Byrne, Mgr., Digital Engineering, Van Nuys)

W. J. Davis: from Principal Member, D&D Engineering Staff to *Staff Engineering Sientist* (L. M. Seeberger, Mgr., Display Engineering, Van Nuys)

L. M. Perkins: to *Ldr., D&D Engineering Staff* (J. C. Wight, Act. Mgr., ECM Engineering, Van Nuys)

Milton Goldin: from Member, Systems Engineering Technical Staff, to *Ldr., Systems Engineering Staff* (Bethesda)

Missile and Surface Radar Division, DEP

D. L. Lundgren: from Engr. to *Ldr., Engineering Systems Projects* (J. M. Seligman, Mgr., System Projects, Moorestown)

R. K. Kaye: from Engr. to *Ldr., Engineering Systems Projects* (R. W. Lapidos, Mgr., System Products, Moorestown)

W. E. Scull: from Engr. to *Ldr., Engineering Systems Projects* (R. A. Newell, Mgr., TRADEX-PRESS Program, Moorestown)

R. D. Kemp: from Engr. to *Ldr., Design and Development* (L. G. Peterson, Mgr., Engineering Manuals, Moorestown)

E. J. Schmitt: from Engr. to *Ldr., Engineering Systems Projects* (F. L. Bernstein, Mgr., Systems Projects, Moorestown)

W. T. Patton: from Engr. to *Ldr., Design and Development* (R. C. Spencer, Mgr., Equipment Design and Development, Moorestown)

O. M. Woodward: from Engr. to *Ldr., Design and Development* (R. C. Spencer, Mgr., Equipment Design and Development, Moorestown)

D. Keys: from Engr. to *Ldr., Engineering Systems Projects* (L. T. Carapellotti, Mgr., System Projects, Moorestown)

F. V. Cavoto: from Engr. to *Ldr., Design and Development* (E. Hatcher, Data Handling and Simulation, Moorestown)

G. A. Brusca: from Engr. to *Ldr., Design and Development* (E. Hatcher, Data Handling and Simulation, Moorestown)

E. E. Roberts, Jr.: from Ldr., Engineering Systems Projects to *Admin., Engineering Ground Support* (F. J. Gardiner, Supervisor, Moorestown)

RCA Victor Record Div.

L. Jones: from Engineer, Factory Engineering, to *General Foreman, Quality Control Section* (Indianapolis)

DEGREES GRANTED

J. M. S. Neilson, SC&M, Mountaintop	MS, Physics, University of Pa.
F. M. Brock, BCP, Camden	MSEE, University of Pa.
E. W. MacLaren, DSD, Bethesda	MEA, George Washington Univ.
A. Avanessians, RCAC, N.Y.	MSEE, Columbia University
A. B. El-Kareh, RCA Labs	Ph.D., University of Delft, Netherlands
D. Donze, SCD, Camden	MSEE, Drexel Inst. of Technology
B. L. Dickens, SCD, Camden	MSEE, University of Pa.
J. Monsay, SCD, Camden	MSEE, University of Pa.
L. Abbott, SurfCom, N.Y.	BEE, The City University of N.Y.
A. Ashkinazy, SurfCom, N.Y.	BEE, The City University of N.Y.
C. Atzenbeck, SurfCom, N.Y.	MEE, Polytechnic Institute of Brooklyn

ARE YOU REGISTERED?

In the June-July 1962 RCA ENGINEER the editorial by **J. C. Walter,** "Registered Professional Engineers in Industry" included a list of some 230 RCA engineers who were known to possess state licenses. The Editors realized the list was incomplete, and asked that readers who were licensed (and not included in that list) send that information in. Since then, additions to the roster (including those below received in the past few weeks) bring the total published to date to nearly 400. *If you are licensed and have not yet informed the* RCA ENGINEER, *send that information to:* RCA ENGINEER, 2-8, Camden, N.J.

L. L. Caviness, DEP, PE-A275, *North Carolina*
W. H. Enders, RCA Labs., PE-8134, *Mass.*
G. G. Gerlach, DEP, PE-6956E, *Pa.*
F. A. Hartshorne, DEP, PE-2022-E, *Pa.*
M. Landis, DEP, PE-28325, *N.Y.*
E. Leshner, DEP, PE-5804-E, *Pa.*
A. Lochanko, BCD, *Providence of Ontario, Canada*
L. N. Merson, DEP, PE-12497, *N.J.*
R. J. Pfeifenroth, DEP, PE-40621, *Calif.* and PE-13418, *Texas*
S. A. Raciti, DEP, PE-39124, *N.Y.*
J. L. Seibert, NBC, EE-4732, *Calif.*
A. W. Sinkinson, DEP, PE-15548, *Mass.* and PE-1207, *Vt.*
A. Z. Szczepkowski, DEP, PE-7359, *Mass.*
C. E. Thomas, Jr., EDP, PE-3572, *Ga.*
G. J. Lawrence, DEP, PE-12629, *N.J.*
A. Drake, ACCD-DEP, PE-11060, *N.J.*
J. A. Brooks, Jr., SCD-DEP, PE-011154, *Pa.*

STAFF ANNOUNCEMENTS

Electronic Data Processing, Camden: **A. D. Beard,** Chief Engineer, Engineering, names his Engineering organization as follows: **R. A. Alexander,** Mgr., Common Components and Standards Engineering; **C. M. Breder,** Mgr., Administration and Control; **L. Iby,** Mgr., Mechanical Coordination; **R. E. Montijo,** Mgr., Systems Engineering; **G. D. Smoliar,** Mgr., Advanced Computer Product Design Engineering; **A. J. Torre,** Mgr., Peripheral Product Engineering; and **J. J. Worthington,** Mgr., 601 Project.

DEP Data Systems Division, Van Nuys: **H. R. Wege,** Vice President and General Manager, DEP-DSD, has announced that **R. W. Landee** is appointed Chief Engineer, Engineering Department, and that **W. M. McCord** is appointed Mgr., Projects Management. Both report to Mr. Wege.

DEP Defense Engineering, Camden: **Dr. H. J. Watters,** Chief Defense Engineer, has named **D. Shore** as Chief Systems Engineer, Systems Engineering Evaluation and Research (SEER). He reports to Dr. Watters.

Electron Tube Division, Industrial Tube Products Department: **D. W. Epstein,** Mgr., Conversion Tube Operations, announces his organization as follows: **R. W. Engstrom,** Mgr., Advanced Development Engineering—Conversion Tube; **W. G. Fahnestock,** Mgr., Operations Planning and Controls—Conversion Tube; **L. W. Grove,** Mgr., Display Tube Operation; **J. K. Johnson,** Mgr., Photo Cell Operation—Mountaintop; **J. A. Molzahn,** Mgr., Quality Control—Conversion Tube Operations; **G. A. Morton,** Director, Conversion Devices Laboratory—Princeton; **M. Petrisek,** Mgr., Camera Tube Operation; **R. C. Pontz,** Mgr., Photo and Image Tube Operation; and **F. S. Veith,** Staff Engineer.

H. K. Jenny, Manager, Microwave Engineering, announced his staff as: **M. Nowogrodzki,** Mgr., Microwave Product Engineering; **R. G. Talpey,** Mgr., Microwave Support Engineering and Special Products Manufacturing; **F. E. Vaccaro,** Mgr., Microwave Applied Research; **P. R. Wakefield,** Mgr., Microwave Engineering Projects and Research and Development Liaison; and **B. Walley,** Mgr., West Coast Microwave Engineering Operation.

W. P. Bennett, Manager, Regular Power Tube Engineering, announces his organization as: **J. W. Gaylord,** Mgr., Regular Power Tube Design Engineering; and **A. P. Sweet,** Mgr., Regular Power Tube Applications Engineering.

Semiconductor and Materials Division, Somerville: **A. M. Glover,** Vice President and General Manager, SC&M, announces his staff as: **F. R. Buchanan,** Controller; **N. H. Green,** Mgr., Commercial Semiconductor Products Department; **T. R. Hays,** Mgr., Marketing Department; **E. O. Johnson,** Chief Engineer, Engineering; **R. L. Kelly,** Administrator, Product Assurance; **R. E. Koehler,** Mgr., Microelectronics Department; **C. H. Lane,** Mgr., Industrial Semiconductor Products Department; **K. M. McLaughlin,** Mgr., Memory Products Department; **W. H. Painter,** Division Vice President, Operations Planning, and **C. E. Sharp,** Mgr., Personnel.

Mr. Johnson announces his staff as follows: **R. B. Janes,** Mgr., Advanced Development; **E. O. Johnson,** Acting Mgr., Engineering Services; **H. V. Knauf,** Mgr., Equipment Development; **R. D. Lohman,** Mgr., Integrated Circuits Project; and **A. H. Noll,** Mgr., Engineering Administration.

Mr. Green announces the organization of the Commercial Semiconductor Products Department as follows: **R. M. Cohen,** Mgr., Commercial Products Engineering; **G. J. Feder,** Mgr., Somerville Plant, **R. J. Hall,** Mgr., Findlay Plant; **J. W. Karoly,** Mgr., Finance and Consumer Operations Planning; **R. E. Rist,** Mgr., Computer Operations Planning; **J. W. Ritcey,** Administrator, Product Evaluation; and **L. R. Shardlow,** Mgr., Commercial Products Development Shop.

Mr. Lane announces the organization of the Industrial Semiconductor Products Department as follows: **D. J. Donahue,** Mgr., Engineering, **P. T. Valentine,** Mgr., Aerospace Reliability Products, **D. H. Wamsley,** Staff Engineer; **D. Watson,** Mgr., Operations Planning and Financial Controls; and **W. H. Wright,** Plant Mgr., Mountaintop Plant.

Mr. Koehler announces the organization of the Microelectronics Department as follows: **B. V. Dale,** Acting Mgr., Battery Project; **L. H. Good,** Mgr., Engineering; **P. Greenberg,** Mgr., Operations Planning and Controls; **R. E. Koehler,** Acting Mgr., Financial Plans; and **L. J. McGrath,** Mgr., Manufacturing.

ETD ESTABLISHES DIRECT ENERGY CONVERSION DEPARTMENT

A Direct Energy Conversion Department has been established in the Electron Tube Division at Harrison, N.J. The function of this department is to: 1) Direct and integrate the direct energy conversion programs of the Electron Tube Division, such as thermionic energy converters; thermoelectric materials and devices for power generation; thermoelectric materials and devices for cooling; solar cells; and super-conducting materials and devices; and 2) provide effective liaison between this effort and other affected activities inside and outside RCA.

The Direct Energy Conversion Department will continue to utilize the services of other departments, as feasible. **L. R. Day** was appointed Manager, Direct Energy Conversion Department, and will report to **D. Y. Smith,** Vice President and General Manager, ETD.

In addition to his duties as outlined above, Mr. Day has been Acting Manager, New Business Development, and will continue to report in that capacity to Mr. Smith.

ETD ESTABLISHES FIFTH RESIDENT LAB AT RCA LABORATORIES

The Electron Tube Division has established its fifth affiliate laboratory at the David Sarnoff Research Center. The new group—the Materials Display Devices Laboratory of Kinescope Engineering—is directed by **Dr. Harold B. Law.** It is responsible for research and development in the areas of advanced display, systems materials, and devices as well as for providing advanced technical assistance to ETD product activities on conventional display problems.

INCREASED FACILITIES FOR MONTREAL LABS

The RCA Victor Co., Ltd., Research Laboratories, Montreal, Canada, announce the addition of 7400 sq. ft. to their facilities. The infrared, semiconductor device applications and systems analysis groups are now located in offices and laboratories in this new area. The Company library, a larger Research conference room and facilities for report preparation and assembly are also included in the additional space—*H. J. Russell*

DR. D. A. ROSS NAMED TO HEAD RCA GRADUATE RECRUITING

Appointment of **Dr. D. A. Ross** as Manager, Graduate Recruiting, was announced recently by **Dr. George H. Brown,** Vice President, RCA Research and Engineering. Dr. Ross, who has been associated with RCA since 1958, will supervise and coordinate all technical recruiting on the graduate level for both RCA Laboratories and the R&D activities of the RCA product divisions. He will place increased emphasis on RCA's Ph.D. recruiting activities, enabling a greater assessment of the overall corporate needs for highly trained technical personnel. Dr. Ross will conduct his activities from the Princeton Laboratories and also will have an office at the RCA College Relations headquarters at Cherry Hill, N.J.

A native of Montreal, Canada, Dr. Ross was graduated in 1947 from McGill University. He later attended Yale University, receiving an M.S. degree in physics in 1955 and his Ph.D. in 1957.

During his four-year association with RCA, Dr. Ross has headed the RCA Laboratories' Radiation Group at Industrial Reactor Laboratories, Inc., the nuclear reactor research facility at Plainsboro, N.J., operated jointly by RCA and nine other companies.

MONTREAL LABS CONTRIBUTE TO TOPSI SATELLITE

The RCA Victor Co., Ltd., Research Laboratories, Montreal, are supplying telemetry equipment for the S.48 TOPSI satellite which is to be fired from Vandenberg AF Base early in 1963. This will be similar to the 136-Mc 2-watt FM transmitter which is still operating with great success aboard the ALOUETTE satellite, S.27, which was fired from Vandenberg in September 1962. (See *News & Highlights,* Dec. '62-Jan. '63 issue.)
—*H. J. Russell*

SC&M CHARTERS SPECIAL "INTEGRATED CIRCUITS PROJECT"

The Semiconductor and Materials Division has established a special project for work on low power integrated semiconductor circuits and digital microcircuits, to be identified as the *Integrated Circuits Project.* **R. D. Lohman** was named Manager of the Integrated Circuits Project. Market Development assistance to this program is being rendered by the SC&M Market Development activity, **W. W. Martenis,** Manager. Mr. Lohman will report to **E. O. Johnson,** Chief Engineer, SC&M.

DSD-BETHESDA REPRESENTED IN NEW ACM GROUP ON INFORMATION RETRIEVAL

The Council of the Association for Computing Machinery has recently approved the formation of a "Special Interest Group on Information Retrieval." **Doug Climenson,** Leader, Systems Engineering Staff was elected vice president and **Sidney Kaplan,** Manager Advanced Information Storage and Retrieval Systems was selected as chairman of the Liaison Subcommittee thereof. Both are with the DEP-DSD Data Systems Center, Bethesda, Md.

The SIGIR group has been asked to assist the computation center of the University of Maryland in the preparation of a set of seminar lectures on Information Retrieval.
—*H. J. Carter*

PROFESSIONAL ACTIVITIES

DEP-SurfCom, Camden: **Roger Knuth** is serving as a Member of the Administrative Committee of the Philadelphia Chapter, IEEE-PTGEWS. **T. H. Story,** Chairman Philadelphia Sec., AIEE, was elected a member of the national AIEE nominating committee from District 2. He was also invited to address the Section Delegates Conference at the National IEEE Convention in New York.—*C. W. Fields*

DEP-SurfCom, Cambridge: Product engineering supervisors attended a local seminar on "Management of Scientific and Technical Personnel"—one of two seminars planned during an extensive training program being conducted covering the interpretation of company policies and procedures. The seminar was conducted by Dr. L. Danielson, Assoc. Prof. of Industrial Relations, U. of Michigan.—*P. J. Riley*

DEP-DSD, Van Nuys, Calif.: Engineers' Week, February 17-23, 1963, was celebrated in the San Fernando Valley by a group of technical societies consisting of the San Fernando Valley Chapter of the Institute of Plant Engineers, the San Fernando Valley Chapter of the California Society of Professional Engineers, the San Fernando Valley Chapter of the American Society of Tool and Manufacturing Engineers, the American Society for Metals, and the San Fernando Valley Sub-section of the IEEE. An "Engineer of the Year" award was made in a joint meeting on 22 February. **William H. Miller,** DEP-DSD, First Vice President, San Fernando Valley Chapter of the CSPE, coordinated the Engineers' Week celebration for the Valley. **Walter Swarthout,** DEP-DSD, Treasurer of the San Fernando Valley Chapter of the CSPE, was also active in this program.—*D. J. Oda*

ETD, Lancaster: **Jules M. Forman,** Engineering Manager of Environmental, Special Equipment & Specification Engineering, of the Electrical Measurements & Environmental Engineernig Laboratory, has been appointed by the Pennsylvania Society of Professional Engineers State Executive Board of Directors as a State Temporary Officer of the Professional Engineer in Industry Functional Section Committee. Mr. Forman is the immediate Past President of Lincoln Chapter, PSPE, which encompasses Lancaster, York, and Adams Counties. Mr. Forman also was recently appointed as RCA consultant to the JT-15 Committee. Mr. Forman was past chairman and founder of the original JT-15 space committee.

John B. Grosh, Acting Engineering Leader of Environmental Engineering at RCA, Lancaster, Pa., has been officially approved as the RCA Company Member of JT-15 "Committee on Environments for Electron Tubes." The JT-15 Committee is an industry-wide committee of the Electronic Industries Association under the Joint Electron Device Engineering Council (JEDEC).—*G. G. Thomas*

DEP-ACCD, Burlington: **O. T. Carver** acted as Chairman of the Session on "Estimation & Prediction by Checkout Equipment" at the Seminar on Automatic Checkout Techniques co-sponsored by BMI and AF, September 5, 6, 7 in Columbus, Ohio.

Zachary Kachemov attended a seminar at Battle Memorial Institute, Ohio, on Automatic Checkout Equipment and obtained information on Checkout Status Control Weapon Systems, etc.

Vincent Mancino attended an IRE meeting and discussed Standards for methods of measurement of transmitter spurious output. —*D. B. Dobson*

RCA Service Co., EDP Service Dept., Cherry Hill: **A. L. Christen** graduated from the Harvard University Program for Management Development on December 15, 1962. He was one of 56 young business executives from 47 companies and 10 foreign countries attending this 16 week course given by the Graduate School of Business Administration. Average age of the participants was 33 years, with 10 years of business experience. Mr. Christen was elected President of his class. Mr. Christen, who is 35 years old received his BSEE (with honors) from Princeton University in 1951. He has been working in the EDP Service activity since its inception in 1960 and was Camden-Philadelphia District Manager prior to attending the Harvard Program. —*J. J. Lawler*

Record Division, Indianapolis: **Stephen W. Liddle** and **Arthur G. Evans** completed a Univ. of Chicago course in "Communications" given by Ind. Central College for RCA engineers.

Construction of a new facility for the Warehouse and Distribution Operation in Indpls. will take place during 1963 with occupancy in 1964. The one-story building covers 280,000 sq. ft. and will adjoin the present warehouse on E. 30th St.—*M. L. Whitehurst*

DEP-ACCD, Camden: **E. W. Keller** is serving on two IEEE Technical Committees: Information Theory and Modulation Systems; and Modulation Systems.

DEP, Central Engineering, Camden: **W. W. Thomas,** DEP Documentation Administrator, in Central Engineering, has been named Chairman of a Government-Industry Task Group for Effective Standardization. Task Group objective is to review the DOD Program related to uniformity in design, engineering, procurement, developing a plan for a supply management with a view toward developing a plan for a realistic and suitable standardization and cataloging effort. Ultimate recommendations of the Task Group will include improvements in policies and procedures, communications, timing of efforts and management organization structure.

Jerome W. Kaufman participated in a special Technological Activities program of the 44th National Recreation Congress held at the Sheraton Hotel, Phila., Pa., on Oct. 4, 1962. Mr. Kaufman was a principal speaker at the Congress. The subject of his paper concerned engineering familiarization at the secondary school level through the medium of JETS clubs. Mr. Kaufman's JETS activities stem from RCA's sustaining membership in the American Society for Metals. Mr. Kaufman is Delegate to the Engineering and Technical Societies Council of Delaware Valley of which the American Society for Metals is a part. The Engineering and Technical Societies Council is the Coordination Office for JETS and Mr. Kaufman is the Chairman of the JETS Committee.—*J. J. Lamb*

DEP-DSD, Data Systems Center, Bethesda, Md.: **Hrand L. Kurkjian** of the Advanced Data Systems Activity attended the International Symposium on Arms Control and Disarmament at the University of Michigan, December 17 through 20, 1962.

Betty Holz of the Advanced Data System Activity, was Chairman of a session on "Experimental Design and Simulation" at the American Association for the Advancement of Science meeting held in Philadelphia's Bellevue Stratford Hotel on December 27, 1962.

The Data Systems Center Human Factors Group in Advanced Data Systems, consisting of **Dr. Richard Krumm, Albert Farina,** and **Roger Graves** participated in a Symposium on the Human Factors Aspects of Photo Interpretation at the Rome Air Development Center on the 7th and 8th of November 1962. Other companies represented at the Symposium included the Boeing Company, Minneapolis-Honeywell Regulator Company, Cornell Aeronautical Laboratories, and Applied Psychology Services. —*H. J. Carter*

SC&M, Mountaintop, Pa.: All SC&M division production engineers at Mountaintop are taking a 12-weeks Work Simplification Course conducted by **E. Klein** of Mfg. Stds. —*M. N. Slater*

RCA Victor Co. Ltd., Montreal, Labs.: **K. A. Graf,** of the RCA Victor Research Laboratories, has been granted leave of absence to work towards his Ph.D. at the Aerophysics Institute, University of Toronto. His work will be closely related to that of the plasma physics laboratory which he is temporarily leaving.

Dr. J. R. Whitehead, Director of Research and **Dr. F. G. R. Warren,** Laboratory Director, Electronics, of the RCA Victor Research Laboratories were session chairmen at the IRE Communications Symposium held in Montreal on Nov. 16, 17, 1962. (The RCA Victor papers presented there are listed in *Pen & Podium,* this issue.)

Dr. M. P. Bachynski, Laboratory Director, Microwave and Plasma Physics in the RCA Victor Research Laboratories in Montreal has been elected Chairman of Commission VI of the Canadian National Committee of URSI (International Scientific Radio Union). Dr. Bachynski is not only the first industrial representative to serve in this capacity, but he is probably, at 32 years old, the youngest.—*H. J. Russell*

RCA Victor Home Instruments, Indianapolis: H. I. Division and Purdue Univ. is sponsoring a graduate level course on transistor theory at the Indpls. Purdue Campus.

E. Montoya, RV engineering, is in charge of a plant training school for hourly employees on basic electricity, drafting, machine shop, math and electronics. They will last 28 weeks.

R. C. Graham, RV engineer, is Editor of the IRE *Indiana Reporter for Engineers.* **M. C. Mehta,** RV engineer, is Assistant Editor. Local Editors from RCA are: **Dave Ballard,** Marion plant; **Lorin Watters,** Indpls. plant; and **W. Barr,** Bloomington plant. The October 1963 Indpls. Section IRE meeting will be sponsored by the RCA Tube Division.—*R. C. Graham*

RCA Laboratories, Princeton: **Louis Pensak,** Electronic Research Laboratory, is on leave of absence from September 1, 1962 to August 31, 1963 as Visiting Associate Professor of Electrical Engineering, at Duke University, Durham, North Carolina. He will be concerned with medical electronics. —*C. W. Sall*

RCA Staff: **Madhu S. Gokhale,** RCA engineer on leave of absence in India and U.N.O. adviser to the Indian Standards Institution on Company Standardization, recently visited the Standards Institution of Israel and lectured before the senior staff on the importance of company standardization and Quality Control.

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