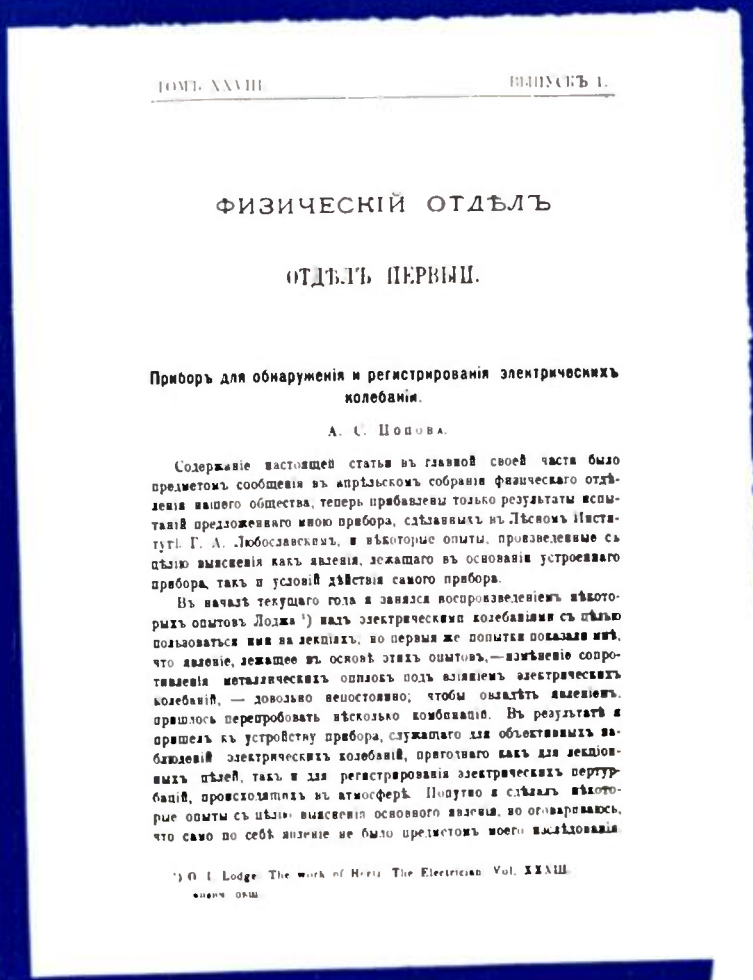
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Proceedings of the IRE

in this issue

POPOV AND RADIOTELEGRAPHY
THE NEURISTOR
SIMULATING NERVE AXON
MODULATED WAVEFORMS
SATELLITE IONIZATION EFFECTS
DIELECTRIC MICROWAVE RESONATOR
RESOLUTION IN RADAR SYSTEMS
TRANSACTIONS ABSTRACTS
ABSTRACTS AND REFERENCES



ТОМЪ XXVIII. ВЫПУСКЪ 1.

ФИЗИЧЕСКІЙ ОТДѢЛЪ

ОТДѢЛЪ ПЕРВЫЙ.

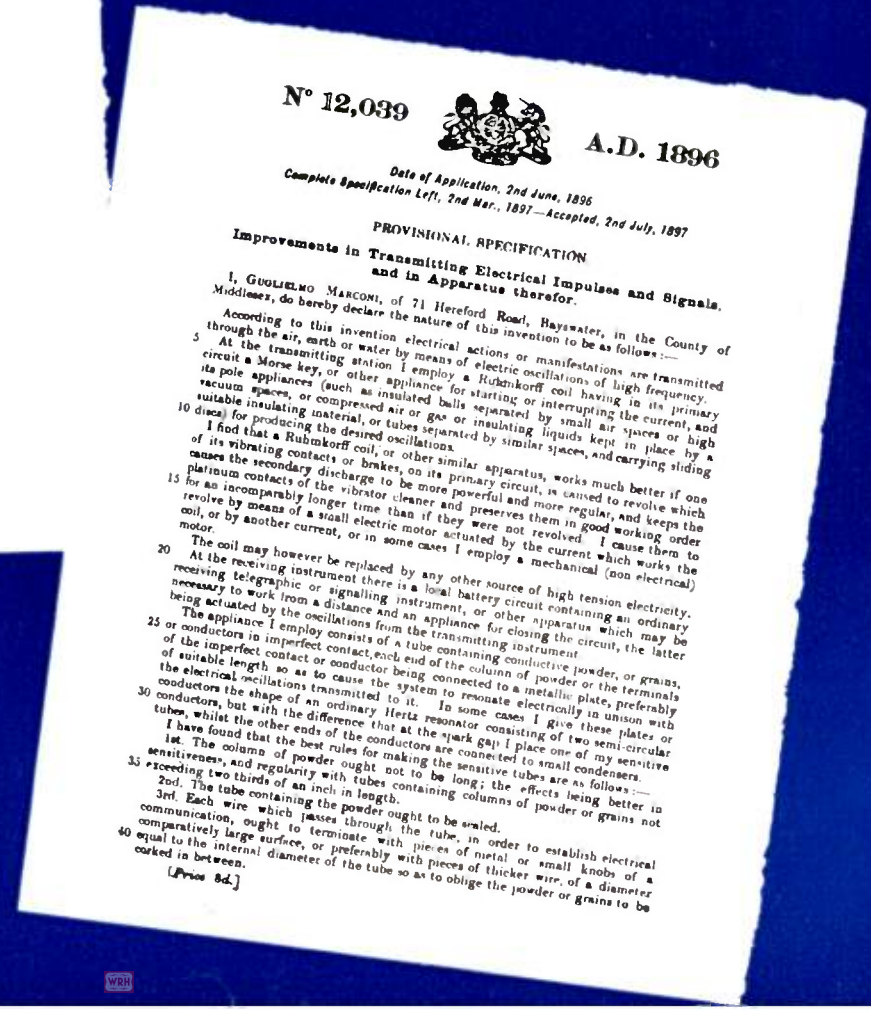
Приборъ для обнаруженія и регистрированія электрическихъ колебаній.

А. С. Поповъ.

Содержаніе настоящей статьи въ главной своей части было предметомъ сообщенія въ апрѣльскомъ собраніи физическаго отдѣленія нашего общества, теперь прибавлены только результаты испытаній предложеннаго мною прибора, сдѣланныхъ въ Лѣсномъ Институтѣ Г. А. Любославскимъ, и нѣкоторые опыты, произведенные съ цѣлью выясненія какъ явленія, лежащаго въ основаніи устроеннаго прибора, такъ и условій дѣйствія самого прибора.

Въ началѣ текущаго года я занялся воспроизведеніемъ нѣкоторыхъ опытовъ Лоджа [1]) надъ электрическими колебаніями съ цѣлью пользоваться ими на лекціяхъ, но первыя же попытки показали мнѣ, что явленіе, лежащее въ основѣ этихъ опытовъ,—измѣненіе сопротивленія металлическихъ опилокъ подъ вліяніемъ электрическихъ колебаній, — довольно непостоянно; чтобы овладѣть явленіемъ, пришлось перепробовать нѣсколько комбинацій. Въ результатѣ я пришелъ къ устройству прибора, служащаго для объективныхъ наблюденій электрическихъ колебаній, пригоднаго какъ для лекціонныхъ цѣлей, такъ и для регистрированія электрическихъ пертурбацій, происходящихъ въ атмосферѣ. Попутно я сдѣлалъ нѣкоторые опыты съ цѣлью выясненія основного явленія, но оговариваюсь, что само по себѣ явленіе не было предметомъ моего изслѣдованія.

[1]) O. J. Lodge. The work of Hertz. The Electrician. Vol. XXXIII. стран. 486.



N° 12,039 A.D. 1896

Date of Application, 2nd June, 1896
Complete Specification Left, 2nd Mar., 1897.—Accepted, 2nd July, 1897

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I, Guglielmo Marconi, of 71 Hereford Road, Bayswater, in the County of Middlesex, do hereby declare the nature of this invention to be as follows:—

According to this invention electrical actions or manifestations are transmitted through the air, earth or water by means of electric oscillations of high frequency.

At the transmitting station I employ a Rubmkorff coil having in its primary circuit a Morse key, or other appliance for starting or interrupting the current, and its pole appliances (such as insulated balls separated by small air spaces or high vacuum spaces, or compressed air or gas or insulating liquids kept in place by a suitable insulating material, or tubes separated by similar spaces, and carrying sliding discs) for producing the desired oscillations.

I find that a Rubmkorff coil, or other similar apparatus, works much better if one of its vibrating contacts or brakes, on its primary circuit, is caused to revolve which causes the secondary discharge to be more powerful and more regular, and keeps the platinum contacts of the vibrator cleaner and preserves them in good working order for an incomparably longer time than if they were not revolved. I cause them to revolve by means of a small electric motor actuated by the current which works the coil, or by another current, or in some cases I employ a mechanical (non electrical) motor.

The coil may however be replaced by any other source of high tension electricity.

At the receiving instrument there is a local battery circuit containing an ordinary receiving telegraphic or signalling instrument, or other apparatus which may be necessary to work from a distance and an appliance for closing the circuit, the latter being actuated by the oscillations from the transmitting instrument.

The appliance I employ consists of a tube containing conductive powder, or grains, or conductors in imperfect contact, each end of the column of powder or the terminals of the imperfect contact or conductor being connected to a metallic plate, preferably of suitable length so as to cause the system to resonate electrically in unison with the electrical oscillations transmitted to it. In some cases I give these plates or conductors the shape of an ordinary Hertz resonator consisting of two semi-circular conductors, but with the difference that at the spark gap I place one of my sensitive tubes, whilst the other ends of the conductors are connected to small condensers.

I have found that the best rules for making the sensitive tubes are as follows:—

1st. The column of powder ought not to be long; the effects being better in sensitiveness, and regularity with tubes containing columns of powder or grains not exceeding two thirds of an inch in length.

2nd. The tube containing the powder ought to be sealed.

3rd. Each wire which passes through the tube, in order to establish electrical communication, ought to terminate with pieces of metal or small knobs of a comparatively large surface, or preferably with pieces of thicker wire, of a diameter equal to the internal diameter of the tube so as to oblige the powder or grains to be corked in between.

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POPOV OR MARCONI? Page 2036

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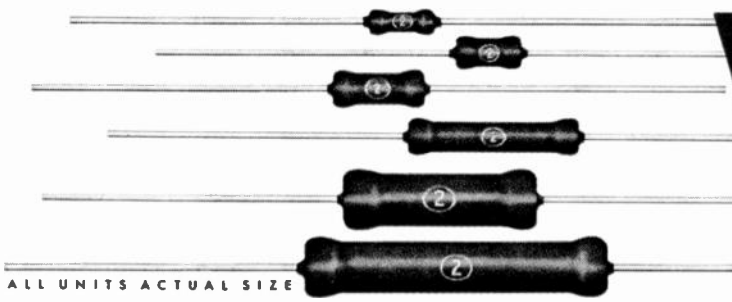
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Effectivity—the Reliability of Complex Systems

It has long been recognized that, when considering the dependability or reliability of any but the simplest equipment, the probability of survival is an inadequate measure. This is most apparent in the case of equipment that may perform a number of functions, since the loss of one whole function might be tolerated. A useful measure for the dependability of complex systems must make allowance for the fact that many failures cause only a relatively minor degradation in system performance. To be really meaningful, such a measure must appropriately weight the importance of all possible failures. This is illustrated in Figure 1, where different failures or combinations of failures define different states of the equipment and provide different degrees of performance. The important elements of this measure of dependability are:

1. The probability that the system is in a particular state, and
2. The relative performance when in that state.

A useful approach to this problem has been to define a *loss coefficient*, $a_i(t)$, as the fractional loss in performance associated with the i^{th} state at time t. In the continuous case, this loss coefficient will be a continuous function of the system variables and will not generally constitute an ordered set. To cope with this situation, we may substitute arbitrarily defined discrete states of the system so as to form a partially ordered set. The loss within each state will then be a continuous random variable having some probability density function, and the *loss coefficient*, $a_i(t)$, is taken as the mean value of the loss in the i^{th} state. We further define the *state probability*, $P_i(t)$, as the probability that the system is in the i^{th} state at time t. From these two factors, we derive the expected fractional loss in performance at time t as

$$\mathcal{L}(t) = \sum_i a_i(t) \cdot P_i(t)$$

where the summation is taken over all possible states. The expected fraction of full system capability realized at a given time is

$$\epsilon(t) = 1 - \mathcal{L}(t)$$

and this we call the *point effectivity* (a point function).

For most well-behaved probability functions we may also form the time

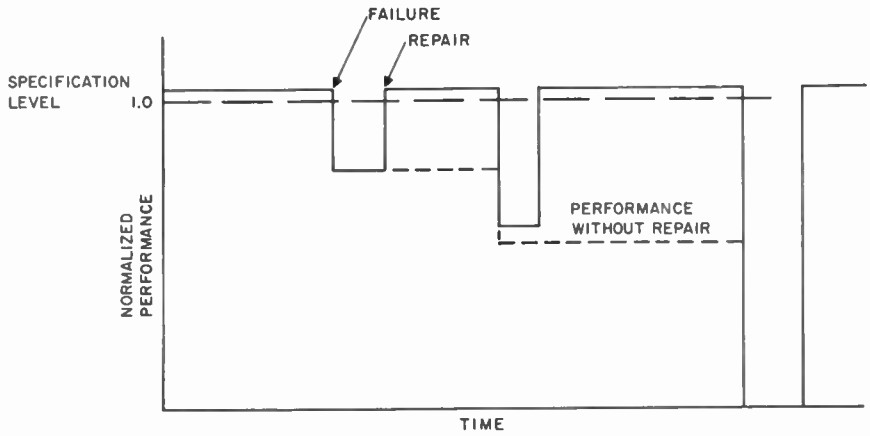


FIGURE 1. PERFORMANCE DEGRADATION OF A GENERALIZED SYSTEM AS A FUNCTION OF TIME.

averages, or interval functions,

$$L(T) = \frac{1}{T}\int_0^T \mathcal{L}(t)dt$$

and

$$E(T) = 1 - L(T).$$

We call E(T), which is the expected fraction of full system capability realized during a given time interval, the *effectivity* of the system.

These concepts provide us with a measure of the generalized performance to be expected, this performance being defined to suit the system under consideration. This permits dependability to be specified as the mean-value of a measure—an approach more analogous to the usual performance factors, whereas the concept of reliability is inherently different. The analysis may be in terms of a point function if instantaneous capability is the important consideration or, for systems having cumulative value, an interval function may be used. Systems having a time-dependent value may also be analyzed. Either discrete or continuous variations may be handled so that slow degradation may be included in the analysis. To relate these measures to the more conventional ones, note that if the loss coefficient, $a_i(t)$, is restricted to zero or unity, that is $a_j = 1$ for $j \varepsilon U$, where U is the set of all unsatisfactory (failed) states, otherwise $a_i = 0$, then

$$E_1(T) = \frac{1}{T}\int_0^T \left[1 - \sum_j P_j(t)\right] dt.$$

Thus, $E_1(T)$ is the expected fraction of the interval that the system is in a satisfactory state and, therefore, it corresponds to a common definition of *availability*. Furthermore, if U is an absorbing set (that is, no repairs are possible), then

$$\epsilon_2(t) = 1 - \sum_j P_j(t) = R(t),$$

or the *point effectivity* is also the *reliability*, and

$$E_2(T) = \frac{1}{T}\int_0^T R(t)dt$$

is the *average reliability*. Thus, as should be the case for any generalization, the concept of effectivity when used in special cases reduces to common terms. When applied to preliminary system design, this concept aids in the comparison of alternative systems, and by combining performance and dependability also provides a firm basis for value analysis. It is hoped that it will help others who must evaluate complex systems for which concepts such as mean time-between failures may not be meaningful.

A more complete discussion of this important concept and pertinent references will be found in the technical paper by K. S. Packard and M. H. Goldstein, "Effectivity: Its Application to a Long-Lived Space System," Sixth Symposium on Ballistic Missile and Aerospace Technology, Los Angeles, August 1961.

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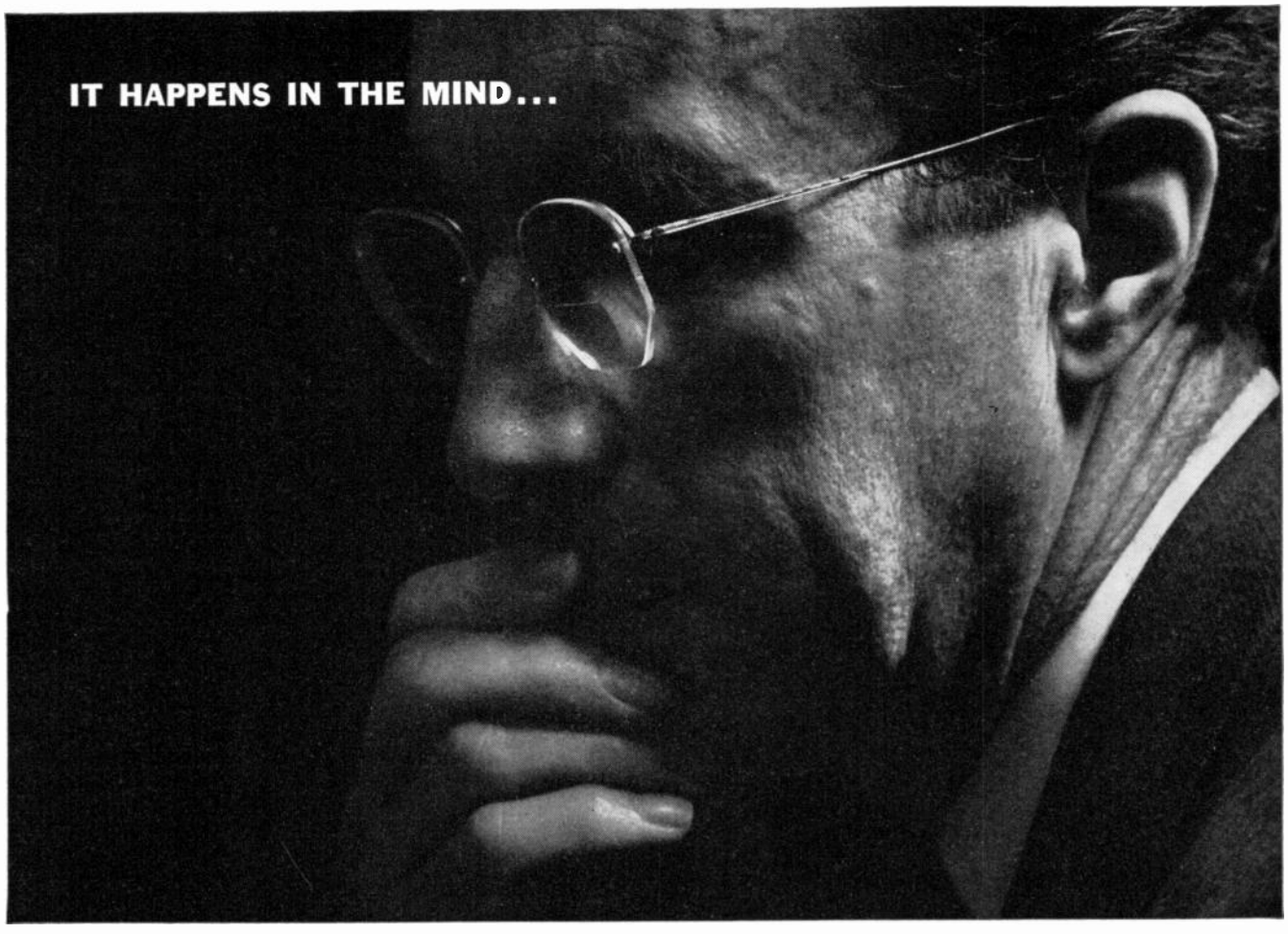
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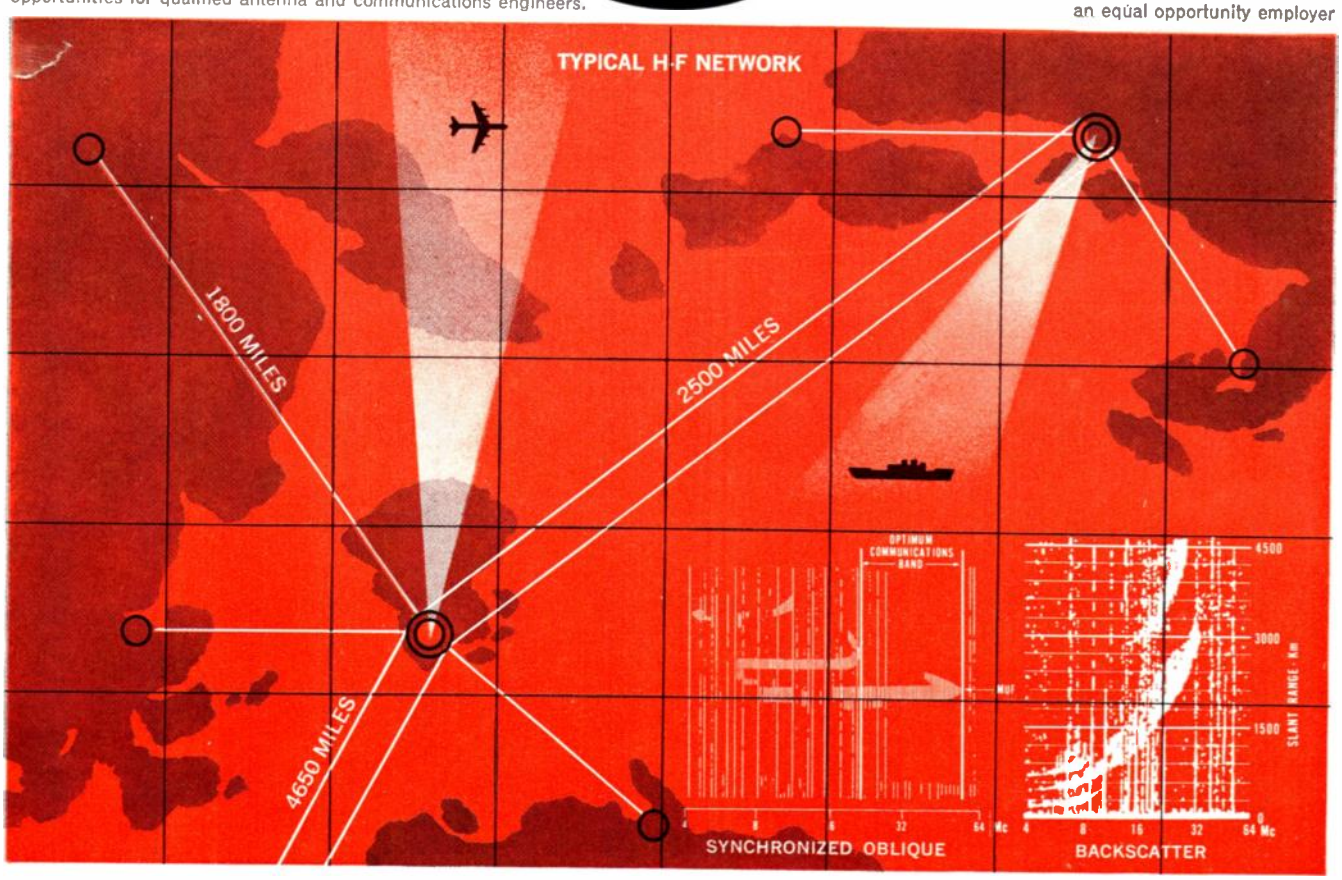
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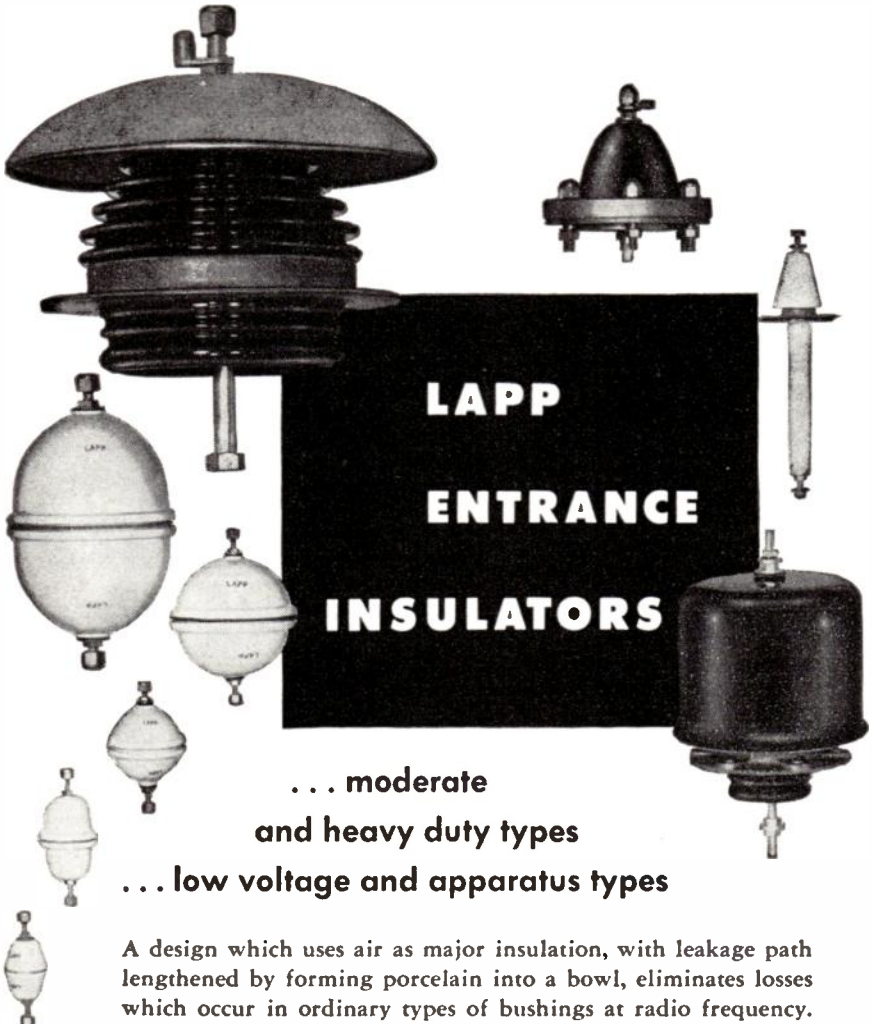
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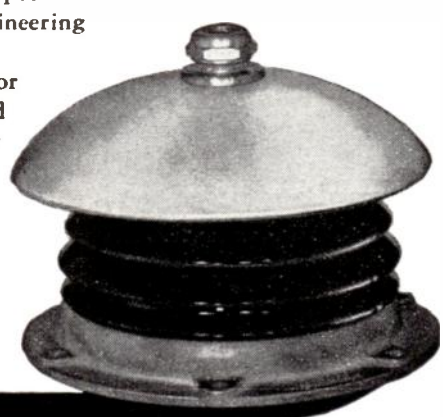
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Meetings with Exhibits

● As a service both to Members and the industry, we will endeavor to record in this column each month those meetings of IRE, its sections and professional groups, which include exhibits.

Δ

October 8-10, 1962

National Electronics Conference, McCormick Place, Chicago, Ill.

Exhibits: Mr. Rudy Napolitan, National Electronics Conference, 228 N. LaSalle St., Chicago, Ill.

October 15-18, 1962

Symposium on Space Phenomena & Measurement, Statler-Hilton Hotel, Detroit, Mich.

Exhibits: Mr. J. B. Bullock, University of Michigan, Ann Arbor, Mich.

November 1-2, 1962

Sixth National Conference on Product Engineering and Production, Jack Tar Hotel, San Francisco, Calif.

Exhibits: Mr. W. Dale Fuller, Lockheed Missiles & Space Div., P.O. Box 504, Sunnyvale, Calif.

November 4-7, 1962

Fifteenth Annual Conference on Electronic Techniques in Medicine & Biology, Conrad Hilton Hotel, Chicago, Ill.

Exhibits: Professional Associates, Inc., 6520 Clayton Rd., St. Louis 17, Mo.

November 5-7, 1962

NEREM (Northeast Research & Engineering Meeting), Commonwealth Armory & Somerset Hotel, Boston, Mass.

Exhibits: Mr. S. K. Gibson, Instruments of New England, 108 Greenwood Lane, Waltham 5, Mass.

November 12-15, 1962

Eighth Annual Conference on Magnetism & Magnetic Materials, Penn-Sheraton Hotel, Pittsburgh, Pa.

Exhibits: Mr. J. L. Whitlock, John Leslie Whitlock Associates, 253 Waples Mill Rd., Oakton, Va.

November 16-17, 1962

Communications Symposium, Queen Elizabeth Hotel, Montreal, P.Q., Canada

Exhibits: Mr. Arthur H. Gregory, Northern Electric Co., Ltd., 1600 Dorchester Blvd. W., Montreal, P.Q., Canada

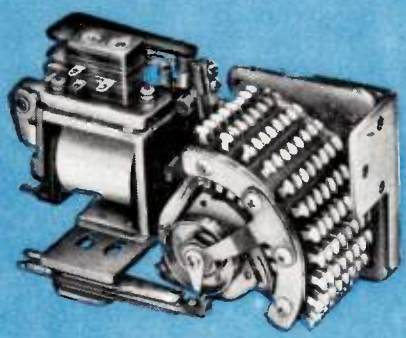
November 19-20, 1962

MAECON (Mid-America Electronics Conference), Continental Hotel, Kansas City, Mo.

Exhibits: Dr. Arthur Goldsmith, Wilcox Electric Co., 1400 Chestnut, Kansas City 27, Mo.

(Continued on page 10A)

SPRING-DRIVEN

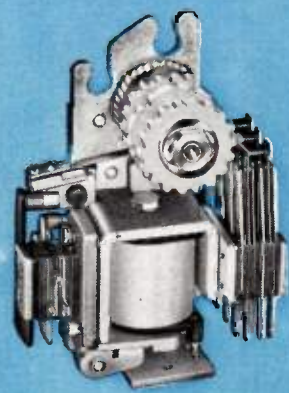


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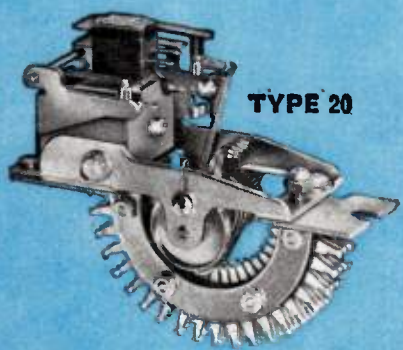


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or four 33-point levels

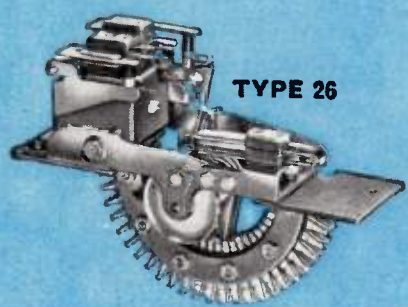
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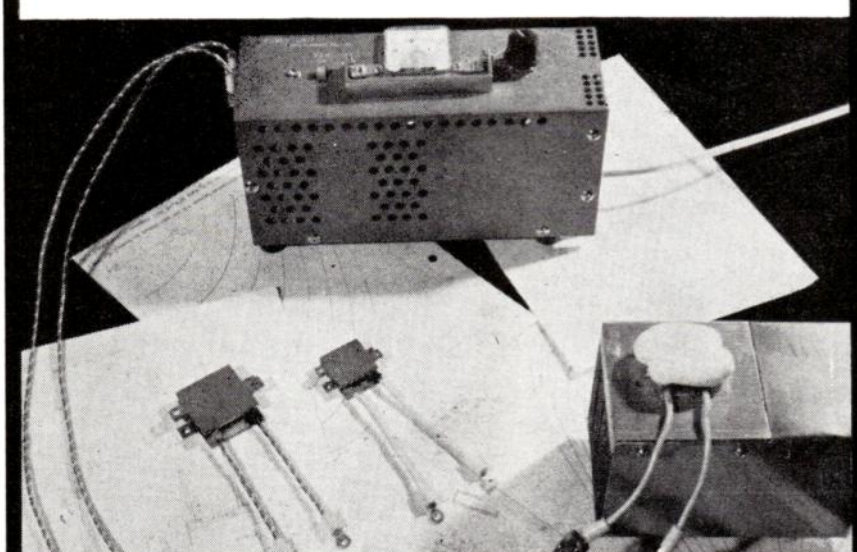


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STM-1021	10	21	0.9	55	50	57×43×10.5 (2.24×1.69×0.413)
STM-1016	10	16	0.9	55	50	57×43×10.0 (2.24×1.69×0.393)
STM-1012	10	12	0.9	55	50	41×32.5×9.5 (1.62×1.28×0.374)
STM-1006	10	6	0.9	55	50	41×29×9.5 (1.62×1.28×0.374)

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Note: (1) No heat load: T_h is 27°C.
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Meetings with Exhibits

(Continued from page 8A)

December 4-6, 1962

Fall Joint Computer Conference, Sheraton Hotel, Philadelphia, Pa.

Exhibits: Mr. R. A. C. Lane, RCA Building 204-1, Camden 8, N.J.

December 6-7, 1962

13th National Conference on Vehicular Communications, Mayfair Hotel, Los Angeles, Calif.

Exhibits: Mr. Leslie M. Walker, Los Angeles County Dept. of Communications, 500 West Temple St., Los Angeles 12, Calif.

January 30-February 1, 1963

Fourth Winter Convention on Military Electronics, Ambassador Hotel, Los Angeles, Calif.

Exhibits: IRE Los Angeles Office, 1435 La Cienega Blvd., Los Angeles, Calif.

February 11-15, 1963

Third International Symposium on Quantum Electronics, UNESCO Bldg., Paris, France

Exhibits: Madame Cauchy, 7 rue de Madrid, Paris 8, France

March 25-28, 1963

International Radio & Electronics Show and IRE International Convention, New York Coliseum and Waldorf-Astoria Hotel, New York, N.Y.

Exhibits: Mr. William C. Copp, IRE Advertising Dept., 72 West 45th St., New York 36, N.Y.

April 17-19, 1963

SWIRECO (Southwestern IRE Conference & Electronic Show), Dallas Memorial Auditorium, Dallas, Texas

Exhibits: Prof. A. E. Salis, E.E. Dept., Arlington State College, Arlington, Texas

May 13-15, 1963

NAECON (National Aerospace Electronics Conference), Biltmore Hotel, Dayton, Ohio

Exhibits: IRE Dayton Office, 1414 E. Third St., Dayton, Ohio

May 21-23, 1963

Spring Joint Computer Conference, Cobo Hall, Detroit, Mich.

Exhibits: Mr. E. Calvin Johnson, Bendix Corp., Research Laboratories Div., Southfield, Mich.

Δ

Note on Professional Group Meetings: Some of the Professional Groups conduct meetings at which there are exhibits. Working committeemen on these groups are asked to send advance data to this column for publicity information. You may address these notices to the Advertising Department and of course listings are free to IRE Professional Groups.



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8063

Motorola Integrated Circuit Electronics...

Types of circuits available; Advantages and disadvantages of hybrid types and functional electronic blocks; The approach to custom-designed circuits

In recent months a dramatic change has occurred in the electronics field.

The widely discussed era of integrated circuits electronics has graduated from an engineering vision to practical reality. And the next few years will find integrated circuits progressing from designers' drawing boards to production lines in electronic equipment ranging from space gear to consumer products.

The total impact of this new technology on the industry, and the magnitude of the envisioned progress is still a matter of widespread conjecture. Though by this time many facts have been substantiated, there is still considerable confusion and misconception as to what can and what cannot be done with integrated circuits *at this time.*

You may have heard, for example, that integrated circuits will greatly reduce equipment size and weight.

This is certainly true. In fact, size and weight reduction is the most dramatic advantage. What the transistor has done to vacuum-tube equipment of a decade ago, integrated circuits will ultimately do to present equipment. Thus, the erstwhile rack of complex equipment which the transistor reduced to single-drawer size, can be further cut to "match-box" proportions.

You've heard, too, that integrated circuits will improve equipment reliability.

And this should be true.

Though reliability is difficult to prove in integrated circuits, since interwoven components cannot be individually tested to their maximum ratings, none of the manufacturing processes by which integrated circuits are made (diffusion, epitaxial growth, alloying, etc.) is new. Through the *exclusive* use of these transistor-proven techniques, complete integrated circuits are expected to have the same order of reliability as other time and field tested semiconductor devices.

You may also have heard that integrated circuits will slash the cost of electronic equipment.

This, unquestionably, is one of the most compelling reasons for circuit integration — but it bears qualification.

Integrated circuits by the thousands are fabricated simultaneously. A single paper-thin wafer of semiconductor material contains hundreds of identical circuits, and dozens of wafers are processed simultaneously. The basic *material* cost of an integrated circuit is little greater than that of a transistor and, with reasonable yields, the cost of the circuit approaches that of the device. In addition, the assembly time of integrated circuit equipment will be but a fraction of that required for circuit wiring.

But the design and production of a functional electronic block circuit (those in which all component parts are fabricated on or within a *single* tiny block of semiconductor material) are as yet costly operations unless your requirements permit large scale use of identical circuits. The great cost-reducing features of such circuits become apparent only when quantity requirements are so large that distributed design costs do not add appreciably to the price of each unit. On the other hand, hybrid (multiple chip) integrated circuits can be built now at a fraction of the cost of functional electronic blocks in small quantities.

If you are in the equipment manufacturing field you may have the impression that integrated circuits will permit a reduction of your engineering staff.

This is probably not the case unless your equipment can be built entirely from a limited number of "stock" circuits. You will need a staff of specialists capable of correlating conventional circuit design with the unique requirements of integrated circuit technology. *Motorola* can provide design assistance, but retention of specific features of each manufacturer's equipment demands an in-house engineering effort.

Finally, you may have heard that integrated circuits limit equipment design flexibility due to the relatively small number of circuits currently available.

Nothing is further from the truth.

Today, the integrated circuit capability of *Motorola's Semiconductor Products Division* has been developed to the point where working samples of a broad range of custom-designed circuits are being delivered in less than six weeks after receipt of an order. Right now, much the same design flexibility obtainable with transistors can be achieved with integrated circuits.

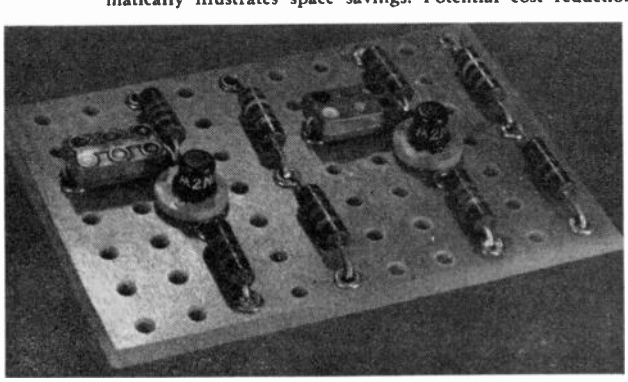
Types of Integrated Circuits

Today's integrated circuit technology involves basically two fundamental processes — a thin-film process, by means of which passive electronic components (resistors, capacitors, etc.) are deposited as material layers on a passive substrate, and a semiconductor device technology where both active and passive elements are formed on or within a tiny block of semiconductor material (silicon).

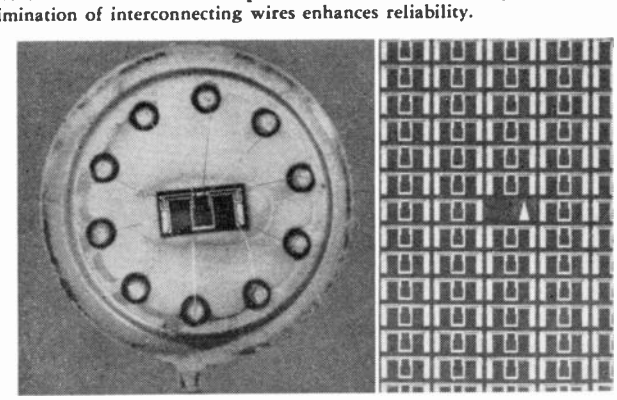
The semiconductor device technology utilizes multiple masked diffusion, surface layer passivation and patterned metal film alloying to make active and passive elements, all properly interconnected and isolated, in a single block. The ranges of component values are somewhat more restricted than in standard circuits and useful values of some functions (i.e. inductance) are difficult to obtain. Parasitics differ from those of conventionally wired circuits.

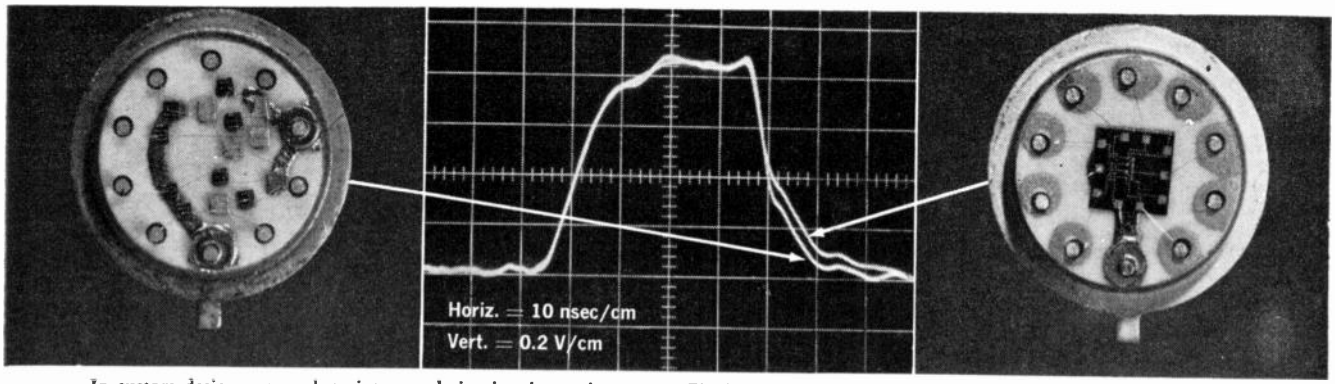
The thin-film technology utilizes multiple evaporation, sputtering, or vapor

Comparison of 50 mc amplifier breadboard using conventional components with equivalent integrated circuit (in center) dramatically illustrates space savings. Potential cost reduction of integrated circuits is indicated by simultaneous fabrication of dozens of identical circuits on portion of wafer shown at right. Elimination of interconnecting wires enhances reliability.



Actual Size





In custom design approach to integrated circuits, the equipment manufacturer's conventional wired prototype is converted into a multiple chip circuit (left) for evaluation. Necessary design changes can be made quickly and inexpensively at this stage.

Final single block circuit (right) offers mass-production cost savings. Comparison of output waveforms of multiple-chip and single block integrated circuits used in an operating arithmetic unit shows that performance of both types is quite similar.

decomposition to deposit (at the present time) the passive circuit components on an insulating (or dielectric) substrate. Active elements must be attached separately to their film circuits and, while the range of values of some components is greater than obtainable through semiconductor technology, such circuits require more material types and processes, and are often larger in size.

Motorola, through extensive research in both integrated circuit fields, has successfully developed a capability for combining the best features of both techniques in yet another, or third, technology—*compatible integrated circuits*. Playing a vital part in this process is the epitaxial growth technique that has recently made an explosive impact in the semiconductor field. Through this combination *Motorola* has produced integrated circuits with increased versatility, greater functional scope, tighter component tolerances and higher speed.

Custom Design Approach

As a first step, the equipment engineer must evaluate the complete system and determine whether to "integrate" all or a portion of the equipment. His next step is to produce conventionally wired prototype circuits for performance evaluation. Often parts values and other parameters must be changed to make the conventional design suitable for circuit integration. Usually this involves consultation with *Motorola* engineers.

Then comes the actual development of the integrated prototype. This can be started by the equipment manufacturer with special integrated components obtainable from the *Motorola Semiconductor Products Division*. These parts, each one individually packaged, have all the electrical characteristics of the elements in the final circuit, but they can be conventionally wired for integrated circuit evaluation and final modifications.

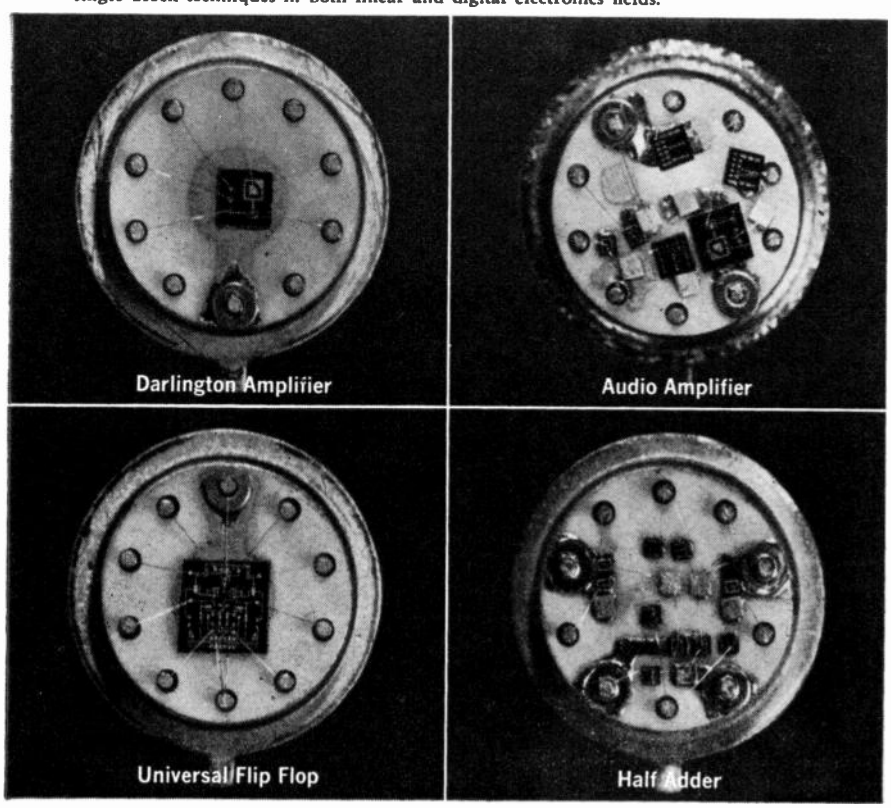
The next step, at *Motorola*, is to convert the equipment manufacturer's prototype into one or more hybrid integrated circuits in which parts identical to those of the prototype are bonded together on multi-lead headers. This becomes an interim circuit form which, in some instances, represents the final product.

Hybrid circuits have many of the same characteristics as fully integrated, single-block devices (see photos above). They have the additional advantages that production changes can be easily made and that design modifications are relatively simple and inexpensive. Moreover, they can be produced quickly at *Motorola*, since a great many integrated "components" are stocked for immediate use. The hybrid circuit, therefore, has much to recommend it as an interim, and for small runs, as a final product.

From the hybrid to the final single block circuit is a matter of developing the optimum pattern and the masks necessary for the various process steps. From a manufacturing standpoint, this is the most costly part of the entire operation. Still, it is entirely justified for equipments involving a large number of the same circuits. The reason — single-block circuits, after development, are the least expensive to make, and potentially the most reliable. With many individual bonding operations eliminated, it is the single-block circuit that offers the greatest cost-saving potential. It does this, however, at the expense of flexibility since such units cannot be modified during production without starting back at the beginning of the manufacturing operation.

The era of integrated circuits has arrived. They are *practical* and they are *available*, even though the future, unquestionably, will bring many improvements to this infant technology.

Integrated circuit capability of *Motorola's Semiconductor Products Division* is illustrated by practical circuits ranging from simple to complex, using multiple chip and single block techniques in both linear and digital electronics fields.



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IRE International News

First Meeting
IRE Region 9

On June 6, 1962 representatives from the IRE Sections included in the new Region 9 met in Geneva, Switzerland, in the office of Gerald C. Gross, Secretary-General of the International Telecommunications Union. Region 9 comprises "Greater Europe," as defined by the ITU.

Dr. R. L. McFarlan, IRE Past President and Chairman of the IRE Committee for International Activities, convened the meeting. He opened the proceedings by outlining the duties of a Regional Director and the functions of a Regional Committee as follows: The Regional Director keeps the Section Chairmen informed of Board decisions, and for this reason is expected to attend each Board meeting. However, a Section Chairman may approach IRE Headquaters directly if he so wishes. A Regional Committee meets at least once and possibly twice a year. It consists of the Director and two representatives from each Section. In addition, the Director can nominate up to twenty-five per cent more members.

Dr. McFarlan then pointed out that, although the position of Director was normally voted for by the entire Region, in the case of the establishment of a new Region, the Committee itself could make the selection. Ir. H. Rinia, who had been Chairman of the Benelux Section for three years, was proposed and seconded. The Committee gave this proposal its unanimous approval, and Mr. Rinia became the first Director of Region 9. The new Director then appointed Professor E. Gatti Vice Chairman, and J. H. Gayer Secretary-Treasurer.

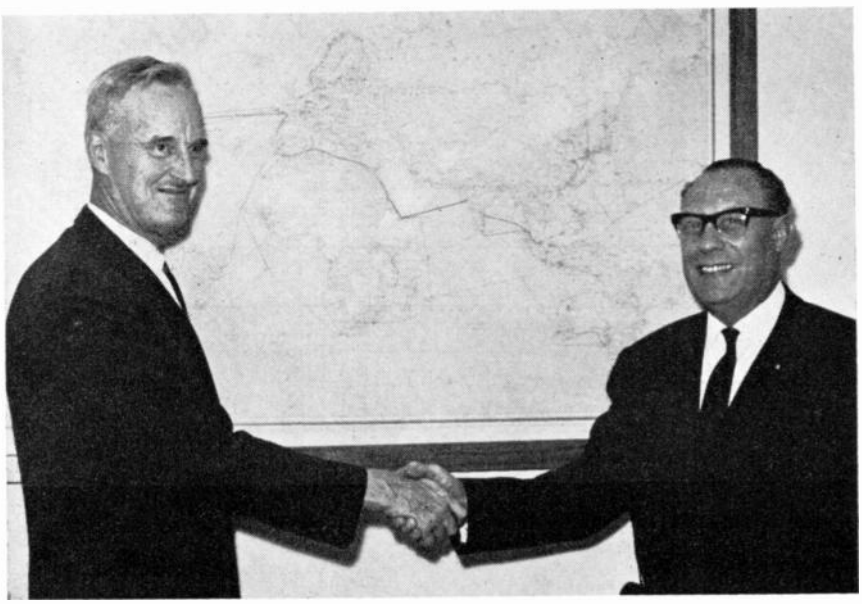
In the discussions which followed, a number of points were raised with regard to the precise position of the IRE and its successor, the IEEE. Referring to representation on organizations such as URSI and ITU, Dr. McFarlan called attention to the fact that such bodies have government representation, and the IRE is primarily concerned with the "dissemination of technical information," and is not a licensing body.

On the question of standardization efforts, Dr. Barrow pointed out that international standards exist which are at variance with IRE standards. He suggested that the appropriate IRE committee should be notified of this fact.

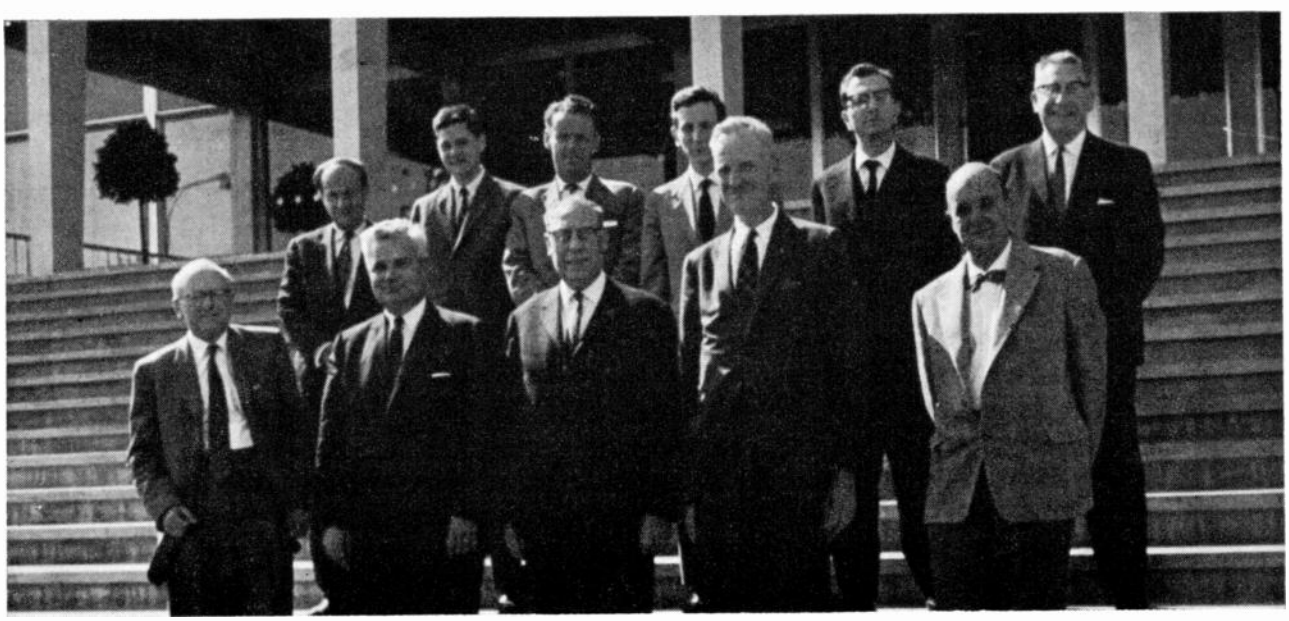
The possibilities of holding a European IRE Convention were also explored. It was speculated that such a meeting might be held in 1964. The representatives decided that the ITU Symposium on Space Communications, to be held in October, 1963, should receive the full support of the IRE.

The use of charter planes, the formation of an Educational Committee, and grading of members were also topics of discussion.

The following delegates attended this inaugural meeting: Benelux Section: Ir. H. Rinia, G. J. Siezen, B. B. Barrow, H. P. Williams; Geneva Section: J. H. Gauer, W. Baumgarten, W. Gerber; France Section: J. Lebel; Italy Section: V. Svelto, G. P. Tarchini; Israel Section: E. H. Frei; IRE Headquarters: R. L. McFarlan.



Dr. R. L. McFarlan, (*left*) Chairman of the IRE Committee on International Activities, and Ir. H. Rinia, (*right*) newly appointed Director of IRE Region 9.



Delegates at the inaugural meeting of IRE Region 9: First Row (*left to right*): H. P. Williams, J. H. Gayer, Ir. H. Rinia, R. L. McFarlan, and W. Gerber. Second Row (*left to right*): E. H. Frei, B. Barrow, G. J. Siezen, Ing. V. Svelto, Ing. G. P. Tarchini, and W. Baumgarten.

Current IRE Statistics

(As of April 30, 1962)

Membership—100,075
Sections*—111
Subsections*—33
Professional Groups*—29
Professional Group Chapters—299
Student Branches†—229

* See September, 1962 issue for a list.
† See this issue for a list.

Calendar of Coming Events and Authors' Deadlines*

1962

Oct. 1–3: 8th Nat'l. Communications Symp., Hotel Utica, Utica, N. Y.
Oct. 2–4: Nat'l. Symp. on Space Elec. and Telemetry, Fontainebleau Hotel, Miami Beach, Fla.
Oct. 7–12: AIEE 1962 Fall General Meeting—3rd Ann. Symp. on Switching, Circuit Theory and Logical Design, Chicago, Ill.
Oct. 8–10: Nat'l. Elec. Conf., McCormick Pl., Chicago, Ill.
Oct. 12–13: 7th Annual North Carolina Section Symp., Greensboro Coliseum, Greensboro, S. C.
Oct. 15–17: URSI-IRE Fall Meeting, Ottawa, Canada.
Oct. 15–18: Symp. on Space Phenomena and Measurement, Statler-Hilton, Detroit, Mich.
Oct. 22–24: ECCANE (East Coast Conf. on Aerospace and Navigational Elec.), Emerson Hotel, Baltimore, Md.
Oct. 25–27: Electron Devices Meeting, Sheraton Park Hotel, Wash., D. C.
Oct. 30–31: Conf. on Spaceborne Computer Engineering, Disneyland Motel, Anaheim, Calif.
Nov. 1–2: 6th Nat'l. Conf. on Product Engr. and Production, Jack Tar Hotel, San Francisco, Calif.
Nov. 4–7: 15th Annual Conf. on Engineering in Medicine and Biology, Conrad Hilton Hotel, Chicago, Ill.
Nov. 5–7: NEREM (Northeast Res. and Engr. Meeting), Commonwealth Armory, Somerset Hotel, Boston, Mass.
Nov. 12–14: Radio Fall Meeting, King Edward Hotel, Toronto, Ont., Can.
Nov. 12–15: 8th Annual Conf., on Magnetism and Magnetic Mat., Penn-Sheraton, Pittsburgh, Pa.
Nov. 16–17: 2nd Canadian IRE Communications Symposium, Queen Eliz. Hotel, Montreal, P.Q., Canada.
Nov. 19–20: MAECON (Mid-America Electronic Conf.), Hotel Continental, Kansas City, Mo.
Nov. 28–30: 1962 Ultrasonics Symp., Columbia Univ., New York, N. Y.
Nov. 28–30: Human Factors Society Annual Meeting, Barbizon-Plaza Hotel, New York, N. Y.
Dec. 4–6: FJCC (Fall Joint Computer Conf.), Sheraton Hotel, Philadelphia, Pa.
Dec. 6–7: 13th Nat'l. Conf. on Vehicular Communications, Disneyland Motel, Anaheim, Calif.

1963

Jan. 8–10: Millimeter and Submillimeter Conf., Cherry Plaza Hotel, Orlando, Fla.

* DL = Deadline for submitting abstracts.

(*Continued on page 16A*)

Call for Papers

1963 IRE International Convention

March 25–28, 1963

Waldorf-Astoria Hotel and the New York Coliseum, New York, N. Y.

Prospective authors are requested to submit all of the following information by the

Deadline Date of October 19, 1962

1. 100-word abstract *in triplicate*, title of paper, name and address
2. 500-word summary *in triplicate*, title of paper, name and address
3. Indicate the technical field in which your paper should be classified:

Aerospace & Navigational Electronics
Antennas & Propagation
Audio
Automatic Control
Bio-Medical Electronics
Broadcast & Television Receivers
Broadcasting
Circuit Theory
Communications Systems
Component Parts
Education
Electron Devices
Electronic Computers
Engineering Management
Engineering Writing & Speech
Geoscience Electronics
Human Factors in Electronics
Industrial Electronics
Information Theory
Instrumentation
Microwave Theory & Techniques
Military Electronics
Nuclear Science
Product Engineering & Production
Radio Frequency Interference
Reliability & Quality Control
Space Electronics & Telemetry
Ultrasonics Engineering
Vehicular Communications

Note: Only original papers, not published or presented prior to the 1963 IRE International Convention, will be considered. *Any necessary military or company clearance of papers must be granted prior to submission.*

Address all material to: Dr. Donald B. Sinclair, Chairman
1963 Technical Program Committee
The Institute of Radio Engineers, Inc.
1 East 79 Street, New York 21, N. Y.

PGRQC Conference Proceedings Available

The *Proceedings of the Second Annual New York Conference on Electronic Reliability, 20 October 1961* are now available. The papers are presented in a readily useable form and cover a wide range of subjects from mathematical analysis and design to the usage and maintenance of hardware in the field. Dr. Landis S. Gephart, Director of Reliability, National Aeronautics and Space Administration, had this to say about the Symposium:

> "I note the subjects selected for the program are both timely and appropriate to the theme. The session on "Reliability Prediction" presents three papers: the first on the general model for analyzing complex systems, by Benjamin Ellison and Victor Selman, describes a system employing continuous operation, on-line and standby redundancy, and maintenance during mission performance; the second, by David Ehrenpreis, provides a quantitative evaluation of mechanical reliability; and the third, by Anthony Coppola, discusses system effectiveness as a function of the joint relationships of reliability, maintainability, and performance. A session is devoted in detail to "Maintainability Factors," whereby Richard Barlow, John Campbell, and A. W. Green discuss, in turn, optimum maintenance intervals, establishment of requirements, and evaluation techniques. In the evening session, Clifford Ryerson and Louis Gamach discuss "Economic Factors," specifically, funding of the reliability engineering program and justification for reliability information on consumer products. The New Tork Section of the IRE is to be commended for the excellent technical program presented at this conference."

While the remaining supply lasts, copies may be obtained at $3.00 each from the Conference Treasurer, Victor Selman, International Electric Corporation, P.O. Box 285, Paramus, N. J. Checks should be made payable to the New York Conference on Electronic Reliability.

International Congress On Quantum Electronics

The Third International Congress on Quantum Electronics will be held at the UNESCO House, Paris, France, on February 10–15, 1963. The French Society of Electronic Technicians and Radio Electricians and the French Section of the IRE have organized this meeting for the occasion of the International Exhibition of Electronic Components. Supported by the U. S. Office of Naval Research and the French National Federation of Electronic Industries, it will follow the precedents of earlier Congresses on Quantum Electronics held in New York and California. Topics discussed will be masers, lasers, and stimulated radiation. There will be an exhibition of equipment in a special section at the International Exhibition of Electronic Components.

IFAC Bulletin Available

Beginning with the 1963 series, the Information Bulletin of the International Federation of Automatic Control (IFAC) will be available on a subscription basis through Headquarters of the Instrument Society of America. This publication has previously been accessible in this country only to officials and committee chairmen of the American Automatic Control Council and its constituent societies.

Published three times yearly, the Bulletin contains news of current events in the field of automatic control in the countries making up the Federation, and in other countries as well. In addition to notices of important meetings and symposia, lists of automatic control papers presented at these meetings, as well as recently published books on a world-wide basis, are given.

Subscriptions to this publication may be ordered from: IFAC Bulletin Department, ISA, Penn Sheraton Hotel, 530 William Penn Place, Pittsburgh 19, Pa. Subscription rate is $3.00 a yer.

Call for Papers
Aerospace Conference

The International Conference on Aerospace Support Systems will be held in Washington, D. C. on August 4–9, 1963. It is designed to be another of the free world's demonstrations of its efforts and achievements in the continuing endeavor toward superior technical competency in the field of Aerospace. The assistance and participation of all members of the aerospace professional community, in preparing for the Conference and Exhibit, is earnestly requested. It is hoped that the electrical electronic and mechanical engineer will team with the logistics expert, the medical and the biological man to improve coordination, and to exchange ideas for advancing the state of international development in Aerospace.

The program of Technical Sessions will include panel discussions and presentation of approximately 200 professional papers. Technical papers are being solicited, with the objective of encompassing as broad a subject area as possible within the context of "Aerospace Support." Subjects to be covered include, but are not limited to: electrical and electronic systems and techniques; mechanical structures, systems, and techniques; guidance control; telemetry; communications; instrumentation and check-out; training and education; management and administration. The scope of direct interest of the sponsoring and participating societies will be highlighted. However, papers on other related aspects of Aerospace Support will be appropriately included in the Technical Program.

Prospective authors may submit abstracts through the Committee and Subcommittee structure of the participating societies, or to the Technical Sessions Committee, Technical Papers, P.O. Box 6635, Washington 9, D. C. Abstracts of not more than 250 words are requested for review and approval by December 3, 1962. After acceptance, the author will receive an Author's Guide Kit for preparing the paper in final form. Deadline for final papers will be February 18, 1963.

Papers Solicited
For INTERMAG

The First International Conference on Nonlinear Magnetics (INTERMAG) will be held on April 17–19, 1963 at the Shoreham Hotel, Washington, D. C. The Conference is jointly sponsored by the IRE and the AIEE. It will serve as a forum for presenting new developments in Theory and Application on Nonlinear Magnetics. The term "Nonlinear Magnetics" refers to a technology where a nonlinear property, such as magnetic saturation, is essential to the function of the device or system.

Both American and foreign authorities in the field will present papers on the state-of-the-art. In addition, papers are invited in the following fields:

Computer Logic and Memories, including thin magnetic films, ferrites, multi-aperture cores, semipermanent and fixed-memory elements, nonstatic magnetic recording techniques involving tapes, discs, drums and heads.

Silicon Controlled Rectifiers, and other semiconductors in combination with saturable magnetic devices, used in inverters, frequency and phase converters, dc choppers, regulated power supplies and motor controls.

Instrumentation and Control Applications, such as magnetic amplifiers, single or multi-stage, frequency multipliers, pulse generators, dc instrument transformers, magnetometers, pressure and torque transducers.

Abstracts of 200 words should be submitted no later than November 5, 1962, to: J. J. Suozzi, Technical Program Chairman, Bell Telephone Labs., Allentown, Pa. The completed manuscript is due January 4, 1963. In addition, authors may request publication of their paper in the AIEE *Transactions*, in which case the manuscript must conform to AIEE standards, available from the Technical Program Chairman upon request.

Members of the INTERMAG Committee include: H. F. Storm, General Chairman; R. R. Booth, Vice Chairman; G. F. Pittman, Secretary; and W. D. Hirst, Publicity.

List of IRE Conference Publications Available

IRE Headquarters has compiled a list of all available publications which have resulted from conferences sponsored or co-sponsored by the IRE. The list, called "Miscellaneous IRE Publications Available," indicates the name and dates of each conference, the price, and how to order. Copies of the list are available upon request from IRE Headquarters, 1 East 79th Street, New York 21, N. Y.

Calendar of Coming Events and Authors' Deadlines*

(Continued from page 15A)

Jan. 21–24: 9th Nat'l. Symp. on Reliability and Quality Control, Sheraton Palace Hotel, San Francisco, Calif.

Jan. 30–Feb. 1: 4th Winter Convention on Military Electronics, Ambassador Hotel, Los Angeles, Calif.

Feb. 11–15: 3rd Internat'l. Symp. on Quantum Electronics, UNESCO Bldg., Paris, France.

Feb. 20–22: Internat'l. Solid State Circuits Conf., Sheraton Hotel and Univ. of Pa., Phila., Pa. (DL*: Nov. 1, 1962, S. K. Ghandi, Philco Scientific Lab., Blue Bell, Pa.)

Mar. 25–28: IRE International Convention, Coliseum and Waldorf-Astoria Hotel, New York, N. Y. (DL*: Oct. 19, 1962, Dr. Donald E. Sinclair, IRE, 1 E. 79 St., New York 21, N. Y.)

Apr. 10–11: 4th Symp. on Engrg. Aspects of Magnetohydrodynamics, Univ. of California.

Apr. 17–19: (Southwestern IRE Conf. and Elec. Show), Dallas Memorial Auditorium, Dallas, Tex.

Apr. 17–19: Internat'l. Non-Linear Magnetics (INTERMAG) Conf., Shoreham Hotel, Washington, D. C. (DL*: Nov. 5, 1962, J. J. Suozzi, BTL Labs., Whippany, N. J.)

Apr, 24–26: 7th Region Tech. Conf., San Diego, Calif.

May 2–3: 4th Nat'l. Symp. on Human Factors in Elec., Marriott Twin Bridges Motel, Washington, D. C.

May 7–9: Electronic Components Conf., Marriott Twin Bridges Hotel, Washington, D. C.

May 13–15: NAECON (Nat'l. Aerospace Electronics Conf.), Dayton, Ohio.

May 17–18: Symp. on Artificial Control of Biology Systems, Univ. of Buffalo, School of Med., Buffalo, N. Y.

May 20–22: Nat'l. Symp. on Microwave Theory and Techniques, Miramar Hotel, Santa Monica, Calif. (DL*: Jan. 15, 1963, Dr. I. Kaufman, Space Tech. Labs. Inc., 1 Space Pk., Redondo Beach, Calif.

May 21–23: Spring Joint Computer Conf., Cobo Hall, Detroit, Mich.

May 27–28: 7th Nat'l. Conf. on Product Engrg. and Production, Boston, Mass.

June 4–5: 5th Nat'l. Radio Frequency Interference Symp., Bellevue-Stratford Hotel, Philadelphia, Pa.

June 11–13: Nat'l. Symp. on Space Electronics and Telemetry, Los Angeles, Calif.

June 19–21: Joint Automatic Control Conf., Univ. of Minnesota, Minneapolis, Minn.

Aug. 20–23: WESCON (Western Elec. Show and Conf.), Cow Palace, San Francisco, Calif.

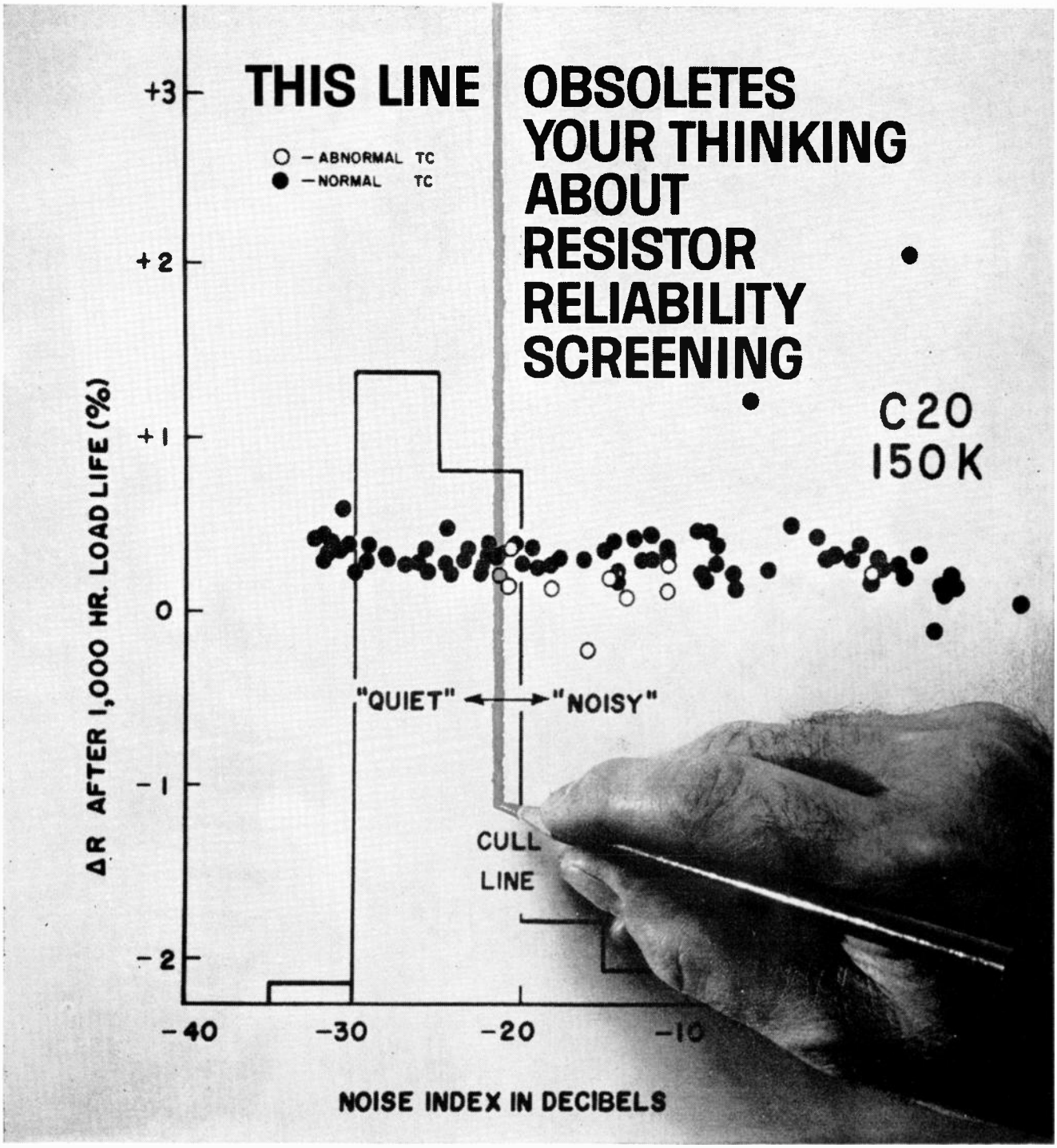
Aug. 27–Sept. 4: 2nd Congress, Internat'l. Fed. of Automatic Control, Basle, Switzerland.

Sept. 9–11: 7th Nat'l. Conv. on Military Electronics, Shoreham Hotel, Washington, D. C.

Sept. 18–19: 12th Annual Industrial Electronics Symp., Michigan State Univ., East Lansing, Mich.

Sept. 30–Oct. 2: Canadian Electronics Conf., Toronto, Ontario, Canada.

* DL = Deadline for submitting abstracts.



It's the reliability CULL* line we can draw on the basis of *current noise level* for any lot of Corning metal-oxide film resistors.

The good guys are on the left. Test them at 2½ times rated power and 25°C. We've been doing it for 23,000,000 part hours without one failure.

If one had failed, you'd have a failure rate of 0.01001%/1000 hours at 90% confidence, 0.00398 at 60%, and 0.00022 at 5%. But none have!

The CULL line falls at different db levels for different styles and sizes of Corning resistors. But we can draw one for any of them that is practical in cost and unsmirched to date for reliability.

Resistors to the right of the line may be OK, but our findings indicate that all mavericks in TC and load life will be on the noisy side of the line.

Read all about this new, *non-destructive* reliability screening tool, and how we'll put it to work for you at modest cost. Write for our new folder, "Current Noise Level: New Reliability Screening Technique for Corning Metal-Oxide Resistors," to Corning Glass Works, 3902 Electronics Drive, Raleigh, N. C.

*Corning Uniformity Limit Level

CORNING
Electronic Components

International Exhibition Of Electronic Components

The Sixth International Exhibition of Electronic Components will be held at the Porte de Versailles, Paris, France, on Februare 8–12, 1963. The Exhibition is supported by the French National Federation of Electronic Industries, and organized by the Society for the Diffusion of Arts and Sciences, with the collaboration of other professional associations.

The First French Components Exhibition was held in 1934, and in 1957 it became an international show. It is considered to be the world's largest display in the field of spare parts and electronic accessories. This year, the Exhibition will have more than 700 participants. Additional information may be obtained from: Salon International des Composants Électroniques, 23 Rue de Lübeck, Paris 16ème, France.

In order to maintain the specific character of the Exhibition, which is entirely devoted to the field of Electronic Components, exhibits are limited strictly to the following items:

Variable and Adjustable Capacitors;

Fixed Capacitors;

Radio-Frequency Coils—coil packs, intermediate frequency transformers, television coils;

Industrial Coils and Transformers for Industrial and Audio Frequencies—power supply transformers, power supply coils, low frequency, voltage regulators, by-pass filters, delay circuits;

Resistors—wire wound resistors, fixed and variable resistors;

Terminal and Wiring Accessories—plugs and sockets, tube holders, printed circuits;

Switching—switches, rotary and linear switches;

Safety Devices—fuses, thermal circuit breakers;

Power Supplies—accumulators, batteries, metal dry rectifiers, dynamos and convertors, vibrators;

Racks and Chassis;

Aerials—Aerials, Frame Aerials;

Electro-Acoustical Material—microphones; record players, amplifiers, loudspeakers, recording and reproduction equipment for magnetic tape, intercoms, hearing aids;

Tubes;

Semiconductors;

Meters;

Relays—thermal, magnetic, magnetic armature relays;

Equipment and Materials for Electronic Purposes—ferrite;

Miscellaneous—shock absorbers, piezoelectrical equipment.

International Radio and Television Exhibition

The First International Radio and Television Exhibition will be held in Paris, France in September, 1963. Manufacturers who are interested in participating in the Exhibition are requested to send notice of agreement before November 15, 1962. The show will be reserved exclusively for firms manufacturing receiving radio and television sets, tape recorders, and reproduction equipment sets for consumer use. Registration and additional information may be obtained from: R. Marty, Fédération Nationale des Industries Électroniques, 23 Rue de Lübeck, Paris 16éme, France.

North Carolina Section Seventh Annual Symposium

Greensboro Coliseum, Greensboro, N. C., October 12–13, 1962

October 12

Keynote Speaker: *T. E. Shea, Vice President, Engineering, Western Electric Company, New York, N. Y.*

"Generators with Gaseous Current Conductors," *Stewart Way, Research and Development Center, Westinghouse Electric Corporation, Pittsburgh, Pa.*

"Micro System Electronics," *Dr. R. M. Burger, Research Triangle Institute, Durham, N. C.*

"Modernizing Our Broadcast Engineering Concepts," *G. Bartlett, National Association of Broadcasters, Washington, D. C.*

"Telstar," *G. F. Critchlow, Satellite Communications Laboratory, Bell Telephone Laboratories, Murray Hill, N. J.*

"Electronics of Gemini," *Earl Paddon, McDonnell Aircraft Corporation, St. Louis, Mo.*

October 13

"Telemetry Systems for Spacecraft," *Robert Rochelle and Roland Van Allen, Spacecraft Technology Division, Goddard Space Flight Center, NASA, Greenbelt, N. C.*

"Optical Maser," *Dr. C. G. B. Garrett, Optical Electronics Research Department, Bell Telephone Laboratories, Murray Hill, N. J.*

Bio-Medical Panel

Moderator: *Dr. J. H. Meredith, M.D., Bowman Gray School of Medicine, Wake Forest College, Winston-Salem, N. C.*

Banquet Speaker: *Dr. Leonard Jaffee, Director of Communications Systems, NASA, Washington, D. C.*

The 1962 Electron Devices Meeting

Sheraton-Park Hotel, Washington, D. C., October 25–27, 1962

The 1962 Electron Devices Meeting will be held at the Sheraton-Park Hotel, Washington, D. C., on October 25–27, 1962. The program is as follows:

Thursday, October 25

Session 1A—Storage and Display Tubes

Chairman: *Russell R. Law, Hughes Research Laboratories, Malibu, Calif.*

Organizer: *Munsey Crost, U. S. Army Signal Corps, Research and Development Laboratory, Fort Monmouth, N. J.*

"A 100 Megabit 9-Digit Electron Tube Encoder," *M. H. Crowell, Bell Telephone Laboratories, Murray Hill, N. J.*

"Dielectric Storage Tape for a Storage Camera Tube," *Arthur S. Jensen and Walter G. Reininger, Westinghouse Electronic Tube Division, Applied Research Dept., Baltimore, Md.*

"Target Scan Parameters as They Affect the Performance of Camera Tubes," *J. Lempert and G. A. Klotzbaugh, Westinghouse Research Laboratories, Pittsburgh, Pa.*

"An Electrostatic Vidicon with High Resolution," *K. Schlesinger, Cathode Ray Tube Dept., General Electric Company, Syracuse, N. Y.*

"An Image Tube with Built-In Electrostatic Zoom Lens," *P. Schagen, Vacuum Physics Div., Mullard Research Laboratories, Redhill, Surrey, England.*

"Direct-View Three Dimensional Display Tube," *R. D. Ketchpel, Research Laboratories, Hughes Aircraft Company, Malibu, Calif.*

Session 1B—Variable Capacitors

Chairman: *Eduard A. Gerber, U. S. Army Signal Corps, Research and Development Laboratory, Fort Monmouth, N. J.*

Organizer: *James D. Meindl, U. S. Army Signal Corps, Research and Development Laboratory, Fort Monmouth, N. J.*

"High Voltage, High Q Epitaxial Gallium Arsenide Diodes," *H. Kressel and N. Goldsmith, Semiconductor and Materials Div., Radio Corporation of America, Somerville, N. J.*

Now...a Panoramic low-cost high performance spectrum analyzer 10mc to 43kmc with all these features:



PANORAMIC
MODEL
SPA-10
SPECTRUM ANALYZER

- **Single tuning head includes coaxial and waveguide input mixers, plus noise-free non-contacting klystron cavity shorts.**
- **High sensitivity** (SEE TABLE BELOW).
- **Dispersion adjustable to 80 mc.**
- **Selectivity adjustable 1-80 kc.**
- **Calibrated dispersion marker with modulation provision to measure narrow band dispersions accurately.**
- **Bright, easily read 5" CRT display with calibrated linear, 40 db log and power amplitude scales.**
- **Single knob tuning control with illuminated slide rule scale, accurate within 1%.**
- **Crystal controlled ± 0.01% markers** (OPTIONAL) **check signal frequency calibrations over entire SPA-10 range.**
- **Adjustable smoothing filter simplifies noise analysis.**

dependable **CERTIFIED SPECIFICATIONS** *for accurate data*

The low cost Model SPA-10 provides sensitive broadband spectrum analysis up through Ka band in a single compact unit. Wide dispersion range to 80 mc, internal calibration and self-checking facilities, and adjustable I-F filter bandwidth for optimized selectivity, are just a few of the characteristics which recommend this carefully designed and packaged analyzer. Non-professional lab and production personnel readily master the easy-to-use SPA-10. Tuning is accomplished with a single low-backlash control which is calibrated within 1% on an illuminated slide rule scale. Excellent shielding permits accurate analysis close to high-power signal sources.

The multi-purpose broadband Model SPA-10 complements Panoramic's unmatched array of widely accepted, exceptionally reliable RF and microwave spectrum analyzers. Many SPA-10 modular design elements are derived from the highly regarded, ultra-sensitive Panoramic Analyzer Model SPA-4a (also available).

Model SPA-10 provides quantitative visual analysis of pulsed signals measuring power points, band occupancy, pulse width, etc. Also for AM, FM, and random signals; distortion, stability and attenuation measurements; and RFI and ECM monitoring.

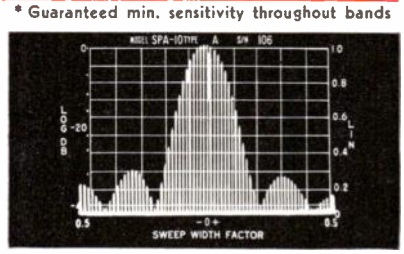
Write, wire or phone today for comprehensive technical bulletin on Model SPA-10, as well as • Model SPA-4a, 10 mc-44 kmc with one head • low cost Model SPA-1, 10 mc-4 kmc with 3 interchangeable heads.

CHECK THESE SPECIFICATION HIGHLIGHTS

Frequency Range—10mc to 43kmc.
Tuning Control Accuracy—1% or ± 1mc, whichever is larger.
Dispersion (Sweep Width)—200kc to 80mc. (200kc to 5mc for 10mc-100mc frequency range.
Selectivity—Adjustable I-F bandwidth, 1kc to 80kc.
Input Attenuators—100 db range for signals up to 12.1kmc. (Ext. waveguide attenuators available for 12-43kmc range.)
I-F Attenuator—0-41 db in 1 db steps. Accuracy: 0.05 db/db.
Sweep Repetition Rate—Adjustable, 1 to 60 cps.
Amplitude Scales (Switch selected) — Calibrated: Linear, 40 db Log, Power (square law).
Frequency Markers—Dispersion marker to ±40 mc. Marker Modulation; Ext source, 10kc-10mc, generates additional markers. (Spacing determined by ext. generator frequency.)
Crystal controlled markers, 0.01% accuracy, at intervals of 1, 10 and 100 mc (optional).
Dimensions—Cabinet, 20"h x 17⅜"w x 19½"d. (For optional 19" width rack mounting, specify the SPA-10/R with RM-1/10 slide tray, 3½"h)

FREQUENCY	SENSITIVITY
10 – 680 mc	– 95* to – 105 dbm
360 – 2360 mc	– 85* to – 95 dbm
2.20 – 6.04 kmc	– 90* to –100 dbm
4.64 – 12.24 kmc	– 80* to – 95 dbm
12.0 – 18.0 kmc	– 70* to – 85 dbm
18.0 – 26.5 kmc	– 60 dbm (nominal)
26.5 – 43.0 kmc	– 50 dbm (nominal)

* Guaranteed min. sensitivity throughout bands



Spectrum of pulsed microwave signal. Synchronization of sweep rate by PRF permits stationary display. Amplitude scale, linear.

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See SPA-10 in action at the N.E.C., Chicago, October 8-9-10, Booth 800

"Diffused GaAs Varactors: Epitaxial and Nonepitaxial," *John C. Irvin, Bell Telephone Laboratories, Murray Hill, N. J.*

"High Q Variable Capacitance Diodes with Extraordinary Capacitance—Voltage Dependence," *G. Edmundson, R. Berkstresser and V. Sils, Sylvania Electric Products, Inc., Woburn, Mass.*

"Semiconductor Junction Varactors with High Voltage Sensitivity," *J. J. Chang, J. H. Forster and R. M. Ryder, Bell Telephone Laboratories, Murray Hill, N. J.*

"Low Loss Ferroelectric Microwave Switch," *A. L. Stanford, Jr. and R. T. Arnold, Sperry Microwave Electronics Co., Division of Sperry Rand Corp., Clearwater, Fla.*

"The Silicon PIN Diode as a Microwave Radar Protector at Megawatt Levels: Theory," *D. Leenov, Bell Telephone Laboratories, Murray Hill, N. J.*

Session 1C—High Power Microwave Tubes

Chairman: *George Caryotakis, Eitel McCullough, Inc., San Carlos, Calif.*

Organizer: *Stanley E. Webber, General Electric Co., Palo Alto, Calif.*

"An L-Band Crossed-Field Amplifier Chain," *H. McDowell and A. Wilczek, SFD Laboratories, Inc., Union, N. J.*

"Electrostatically Focused High-Power Klystron Amplifier," *J. R. Hechtel and A. Mizuhara, Litton Industries, Electron Tube Div., San Carlos, Calif.*

"Pulsed Megawatt X-Band Klystron," *Rita Biss, Electron Tube Div., Sperry Rand Corp., Great Neck, N. Y.*

"A Grounded-Cathode L-Band Pulsed TWT Amplifier," *J. W. Sedin and K. W. Slocum, Watkins-Johnson Company, Palo Alto, Calif.*

"High Power Broad-Band Coupled-Cavity Traveling-Wave Tube," *W. Hant, M. V. Kreismanis and L. M. Winslow, Microwave Tube Div., Hughes Aircraft Co., Los Angeles, Calif.*

"A High Gain, Broad-Band Crossed-Field Amplifier for X-Band," *G. K. Farney and J. Hentschel, SFD Laboratories, Union, N. J.*

Session 1D—Integrated Devices and Microcircuits

Chairman: *Harry Knowles, Westinghouse, Baltimore, Md.*

Organizer: *Herbert M. Pollack, General Precision, Inc., Clifton, N. J.*

"Silicon Planar Integrated Circuitry—Using a Unique Combination of Multi-Layer Epitaxial and Diffusion Techniques," *F. E. Boerger, C. S. Martel and H. M. Robertson, Westinghouse Electric Corp., Newbury Park, Calif.*

"Silicon Planar Integrated Circuits—Devices and Circuits Fabricated Using Multiple Epitaxial and Diffusion Techniques," *L. O. Hill, G. B. LeBlanc and Sherman Moore, Westinghouse Electric Corp., Newbury Park, Calif.*

"Integrated Devices for Hybrid Circuits," *H. K. Dicken and D. Bailey, Motorola, Semiconductor Div., Phoenix, Ariz.*

"Methods of Isolation of Active Elements in Integrated Circuits," *D. S. King, Wilf Corrigan and Glen Madland, Motorola, Semiconductor Div., Phoenix, Ariz.*

"An Approach to Low Cost, High Performance Micro-Electronics," *E. M. Davis, W. E. Harding and R. S. Schwartz, IBM Corp., Poughkeepsie, N. Y.*

"Determination of Temperature Profiles in Microcircuits," *M. Walker, J. Roschen and E. Schlegel, Philco Corp., Lansdale, Pa.*

Analysis of the Simultaneous Fabrication of Unipolar and Bipolar Transistors in a Functional Block," *K. K. Yu, H. C. Lin and M. J. Geisler, Westinghouse Central Laboratories, Pittsburgh, Pa.*

"Nanowatt Devices," *W. W. Gaertner, CBS Laboratories, Stamford, Conn.*

"Stray Capacitance in Thin-Film Circuits," *W. W. Happ, Microsystems Electronics Dept., Lockheed Missiles and Space Co., Sunnyvale, Calif.*

"Integrated, Unijunction Transistor Oscillators," *G. Hachtel and D. O. Pederson, University of California, Berkeley, Calif.*

Friday, October 26

Session 2—Invited Papers

Chairman: *R. W. Sears, Bellcomm, Inc., Washington, D. C.*

"Avalance Multiplication Processes in Semiconductors," *Alan G. Chynoweth, Bell Telephone Laboratories, Murray Hill, N. J.*

"Comparison of Hot Electron and Related Amplifiers," *John L. Moll, Stanford University, Stanford, Calif.*

"Thermionic Energy Conversion," *Karl G. Hernquist, RCA Laboratories, Princeton, N. J.*

Session 3A—Energy Conversion Devices

Chairman: *W. H. Clingman, Texas Instruments, Inc., Dallas, Tex.*

Organizer: *Frank H. Bower, Sylvania Electric Products, Inc., Woburn, Mass.*

"Operating Characteristics of a Molten-Electrolyte Fuel Cell on Hydrocarbon and Hydrocarbon-Derived Fuels," *C. G. Peattie, I. Trachtenberg, J. K. Truitt, B. H. Barbee, J. F. Haefling, T. Hooper, K. W. Kreiselmaier and A. H. White, Texas Instruments, Inc., Dallas, Tex.*

"Solar Heated Thermionic Generators," *L. J. Lazaridis and P. J. Brosens, Thermo Electron Engineering Corp., Waltham, Mass.*

"The Effectiveness of Photovoltaic Solar Cells as an Electric Power Source for Space Vehicles," *Charles D. Fredrickson, Jet Propulsion Laboratory, Pasadena, Calif.*

"Thermo-Photo-Voltaic Conversion from High Temperature Energy Sources," *Bruce D. Wedlock, Energy Conversion Lab., Dept. of Electrical Engineering, Massachusetts Institute of Technology, Cambridge, Mass.*

"Theory of Nernst Generators and Refrigerators," *M. H. Norwood, Texas Instruments, Inc., Dallas, Tex.*

"Carrier Concentration Optimization in Semiconductor Thermoelements," *José Ma. Borrego, Centro de Investigacion y de Estudios Avanzados, Del Instituto Politecnico Nacional, Mexico City, Mexico.*

"Free Convection Heat Transfer Thermoelectric Refrigerators," *I. I. Sochard, Emerson Research Laboratories, Silver Spring, Md.*

"High Current Germanium Tunnel Diodes," *F. M. Carlson and P. D. Gardner, Radio Corporation of America, Somerville, N. J.*

Session 3B—Three-Terminal Solid-State Devices

Chairman: *Gordon Moore, Fairchild, Palo Alto, Calif.*

Organizer: *Edward Johnson, RCA Laboratories, Somerville, N. J.*

"A High-Speed, Medium-Power, Triple-Diffused Germanium NPN Switching Transistor," *W. A. Pieczonka, P. P. Castrucci and R. M. Folsom, IBM Corp., Components Div., Poughkeepsie, N. Y.*

"The Insulated-Gate Field-Effect Transistor," *F. P. Heiman and S. R. Hofstein, RCA Laboratories, Princeton, N. J.*

"A New PNPN Device," *E. A. Karcher, U. S. Army Signal Research and Development Laboratory, Fort Monmouth, N. J.; and A. G. Milnes, Carnegie Institute of Technology, Pittsburgh, Pa.*

"A Tunnel Emission Triode," *R. F. Schwarz, J. P. Spratt, E. Zimmerman and W. Follmer, Philco Scientific Laboratories, Blue Bell, Pa.*

"Low Temperature Oxidation of Silicon by Electric Discharge in Oxidizing Atmosphere," *Takuo Sugano, Hisao Nakajima and Yoshishige Takahashi, Dept. of Electronic Engineering, Faculty of Engineering, University of Tokyo, Japan.*

"The Role of Avalanching in Cold Cathodes," *K. Ramanathan, R. Johannes, and P. H. Cholet, Philco Scientific Laboratories, Blue Bell, Pa.*

Session 3C—Selected Tube Topics

Chairman: *Gordon Guernsey, Lincoln Laboratory, Lexington, Mass.*

Organizer: *Alan D. Sutherland, Electronic Tube Div., Sperry Rand, Gainesville, Fla.*

"An Investigation of Saturated RF-Currents in High Perveance Hollow Beams," *Enzio Demmel and Lee MacKenzie, Varian Associates, Palo Alto, Calif.*

"A New Nickel Matrix Cathode," *D. W. Maurer, E. J. Becker and D. E. Koontz, Bell Telephone Laboratories, Murray Hill, N. J.*

"Noise Near the Carrier in Microwave Oscillators," *Charles Mosher, Cornell University, Ithaca, N. Y.*

"A Versatile Energy Diverter (Crowbar)," *Sol Schneider, U. S. Army Signal Research and Development Laboratory, Fort Monmouth, N. J.*

"The Backward Wave Converter," *R. W. Wilmarth and R. J. Blanchard, Electron Tube Laboratory, ITT Components Division, Nutley, N. J.*

"Permanent Magnet Focusing of Multimegawatt Klystron Beams," *John Romaine and Henry Striegl, Electronic Tube Div., Sperry Gyroscope Company, Great Neck, N. Y.*

Session 3D—Diode Properties

Chairman: *William Hittinger, Bell Telephone Laboratories, Murray Hill, N. J.*

Organizer: *Howard Loar, Bell Telephone Laboratories, Allentown, Pa.*

"Study of the Effects of Fast Neutrons on Silicon Controlled Rectifiers," *Frank A. Leith, Dept. of Electrical Engineering, Massachusetts Institute of Technology, Cambridge, Mass.*

"Heterojunction High Speed Switching Diodes," *F. F. Fang and W. E. Howard, International Business Machines Corporation, Thomas J. Watson Research Center, Yorktown Heights, N. Y.*

"GaAs Junction Diodes as Cryogenic Thermometers," *B. G. Cohen, R. V. Goodman, W. B. Snow and A. R. Tretola, Bell Telephone Laboratories, Murray Hill, N. J.*

FOR PRODUCTION ECONOMY

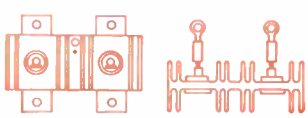
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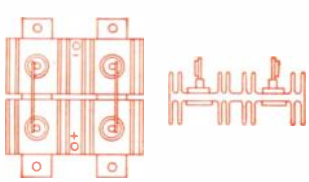
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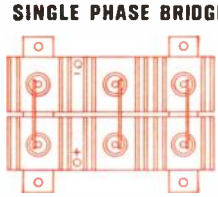
TYPICAL CONFIGURATIONS



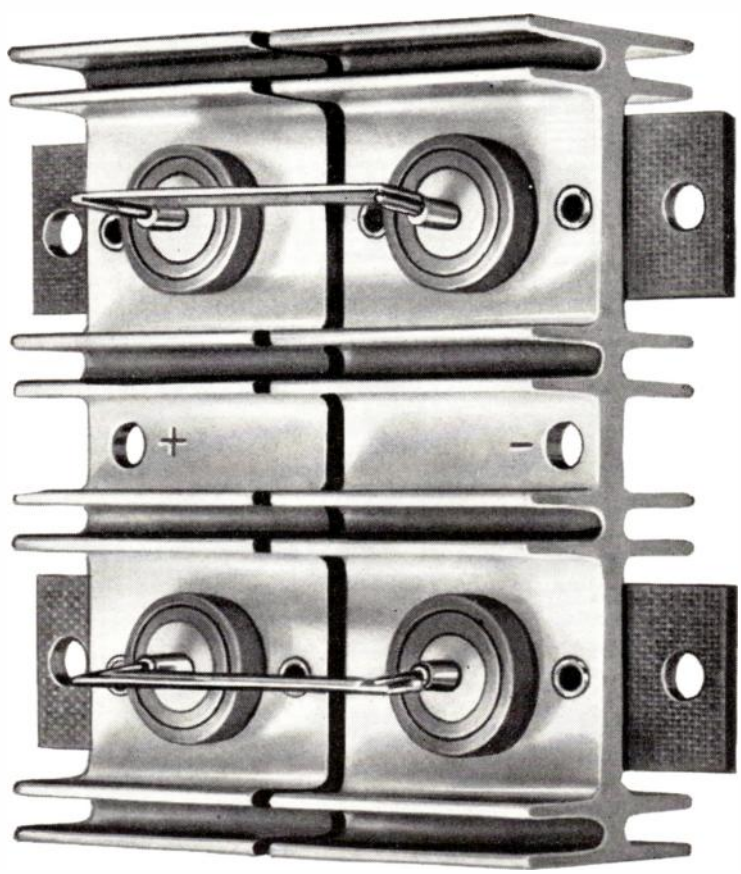
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"Tunneling Current Density as a Function of Crystallographic Polarity," *M. T. Minamoto and H. T. Malafi, Airborne Instruments Laboratory, Melville, N. Y.*

"Control of Electric Fields at the Surface of *p-n* Junctions," *Robert L. Davies and Finis E. Gentry, General Electric Company Electronic Components Div., Auburn, N. Y.*

Session 4A—CW Microwave Tubes

Chairman: *Stanley F. Kaisel, Microwave Electronics Corporation, Palo Alto, Calif.*

Organizer: *Louis W. Roberts, Metcom, Salem, Mass.*

"Millimeter-Wave Tubes Using the Coupled Cavity Circuit," *John E. Nevins, Jr., Hughes Research Laboratories, Malibu, Calif.*

"A 50-75 Gc Power Backward-Wave Oscillator," *Robert T. Schumacher, Tube Div., Varian Associates, Palo Alto, Calif.*

"A PM Focused Ultra-Low-Noise Traveling-Wave Tube for *S*-Band," *B. P. Israelsen and R. W. Peter, Watkins-Johnson Company, Palo Alto, Calif.*

"Lightweight Traveling-Wave Tubes for Space Vehicles," *B. A. Highstrete and K. P. Grabowski, Microwave Tube Div., Hughes Aircraft Company, Los Angeles, Calif.*

"Design of CW Amplitrons for Space-Borne Telemetry Applications," *W. W. Teich and W. C. Brown, Raytheon Company, Spencer Lab., Burlington, Mass.*

"Crossed-Field Backward-Wave Oscillators Above *X*-Band," *R. R. Moats and M. B. Fisher, Sylvania Microwave Devices Div., Mountain View, Calif.*

Session 4B—Other Tubes

Chairman: *George Bernstein, SFD Laboratories, Union, N. J.*

Organizer: *Lester F. Eastman, Cornell University, Ithaca, N. Y.*

"An *S*-Band Quadrupole Amplifier with 70 Degree *K* Noise Temperature," *T. J. Bridges, Services Electronics, Research Laboratory Extension, Admiralty, Harlow, Essex, England.*

"The LINEATOR—A Low-Noise Protective Device," *C. Burton Crumly and Roy A. Larsen, Zenith Radio Research Corporation, Menlo Park, Calif.*

"A Linear Phase Shifter Using Cyclotron Waves," *G. Ryan and W. Sackinger, Zenith Radio Corporation, Research Dept., Chicago, Ill.*

"Design Considerations and Performance Characteristics of Microwave Phototubes," *J. B. Gaenzle, R. T. McKenzie and B. J. McMurty, Sylvania Microwave Device Div., Mountain View, Calif.*

"Omegatron Mass Spectrometer," *C. H. Rehkopf, Chemical Development Section, Electronic Tubes Div., Sylvania Electric Products, Inc., Emporium, Pa.*

"Some Requirements for New Electron Tube Devices," *N. DeClaris, Cornell University, Ithaca, N. Y.*

Session 4C—New Solid-State Devices

Chairman: *Alan L. McWhorter, Lincoln Laboratory, Lexington, Mass.*

Organizer: *Robert Luce, Philco Laboratory, Lansdale, Pa.*

"High Efficiency Injection Electroluminescence in Gallium Arsenide," *T. M. Quist and R. J. Keyes, Lincoln Laboratory, Lexington, Mass.*

"GaAs Infrared Source," *J. R. Baird, E. L. Bonin, W. N. Carr and G. E. Pittman, Texas Instruments, Inc., Dallas, Tex.*

"Single Side-Band Suppressed Carrier Optical Modulation," *C. F. Buhrer, V. J. Fowler, L. R. Bloom, D. H. Baird and E. M. Conwell, General Telephone and Electronics Laboratories, Inc., Bayside, N. Y.*

"Superconducting Magnet for a Traveling-Wave Maser," *M. D. Bonfeld, Bell Telephone Laboratories, Allentown, Pa.*

"Semiconductor Pressure Transducers," *M. E. Sikorski, R. Edwards and P. Andreatch, Bell Telephone Laboratories, Inc., Murray Hill, N. J.*

"A New Class of Devices Based on Stressed Single and Multiple Junctions," *W. Rindner and R. Nelson, Raytheon Research Division, Waltham, Mass.*

6th National Conference on Product Engineering and Production

JACK TAR HOTEL, SAN FRANCISCO, CALIF., NOVEMBER 1–2, 1962

Session I—Components, Processes and Equipment for Electronic Production

"Use of Vacuum Deposition Techniques in an Engineering Laboratory Facility for Producing Thin-Film Integrated Circuits," *Russell D. Kraus, Solid-State Systems Div., Motorola Inc., Phoenix, Ariz.*

"Chemical Processes as Manufacturing Tools for the Electronic Industry," *L. B. Stearns, President, Chemical and Aerospace Products, Inc., Gardena, Calif.*

"Miniature Mechanical Counter-Parts to Micro-Electronic Circuitry," (tentative) *author to be announced.*

"Control of Airborne Particulate Contamination," *Fred Allison, Research Engineer, QRS Laboratories, Autonetics, Anaheim, Calif.*

Session II—Electronic Interconnection Techniques

"Assured Reliability in Soldered Connection," *Dr. L. Pessel, RCA, Camden, N. J.*

"An Approach to Miniaturized Wire Wrap," *T. J. Fox and C. N. Patterson, IBM, Data Systems Div., Poughkeepsie, N. Y.*

"A Survey of Welding Techniques," *Sam Francis, Vice President, The Sippican Corp., Marion, Mass.*

Session III—Man-Hardware Relationship

"Maintainability Demonstration Equipment," *N. Buchaca, Design Assurance Supervisor, Philco Corp., Palo Alto, Calif.*

"A Practical Concept to Maintain Product Reliability Through Integrated Packaging and Handling," *Bronson B. Baker, Manager, Packaging, Handling Engineering and Conservation, Lockheed Missiles and Space Co., Sunnyvale, Calif.*

"The Problems of Test of Miniature Components," (tentative) *author to be announced.*

"Industrial Audio-Visual Applications," *Jack M. Duer, Application Engineer, Nortronics Div. of Northrop Corp., Hawthorne, Calif.*

Session IV—Circuit Packaging

"The Design of a Digital Telemetry Unit Using Dot Components," *J. R. Goodykoontz, Space Technology Lab., Redondo Beach, Calif.*

"Circuit Packaging with Modern Conventional Components," (tentative) *author to be announced.*

"Elephantine Electronics—A New Circuit Packaging Problem," *R. L. Blessing, Senior Product Engineer, and A. M. Poiré, Production Project Engineer, Radiation at Stanford, Palo Alto, Calif.*

"Production Techniques for Integrated Electronics," *W. Dale Fuller, Senior Member, Research Staff, Lockheed Missile and Space Co., Palo Alto, Calif.*

MODEL CFI TRANSISTORIZED CALIBRATED MICROWAVE FIELD-INTENSITY (RI/FI) METER —1,000 to 10,000 MC.



Category A Air-Force Approval For MIL-I-26600 and MIL-I-6181. Transistorized, portable, compact, ruggedized.

Combines an impulse calibrator, field-intensity meter, and calibrated antenna system. Provides accurate measurements of the frequency and the absolute power level of conducted or radiated microwave energy. Choice of 12V battery or line-power operation.

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Category A Air-Force Final Approval For MIL-I-26600 and MIL-I-6181. Widest frequency range ever offered in a microwave RI/FI meter!

Combines internal CW signal generator, field-intensity meter, and calibrated antenna system. Provides accurate absolute power level and frequency measurements. The signal level indication is in microvolts, db above 1 microvolt, and db above 1 microvolt per MC.

- Only microwave Field Intensity Receiver with self-contained CW signal generator which calibrates entire system under test at any level.
- First single microwave test system to determine radiated r-f interference and susceptibility.
- Front-panel meter directly reads average, peak, slide-back peak or quasi-peak value of r-f signals.
- Six interchangeable tuning heads cover 1,000 to 21,000 mc.
- Outputs for video, audio and recorder.
- Single UNIDIAL® tuning control simultaneously tunes the receiver and signal generator.
- Calibrated antenna system includes an omni-directional broadband antenna and separate directional antennas to match each tuning unit.
- Frequency Dial Accuracy: ±1%.
- Maximum RF Input: 3 volts
- Sensitivity: −81 dbm minimum
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- Image and Spurious Response Rejection: Better than 60 db
- Attenuation: 0 to 80 db in steps of 1 db

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How To Select An RI/FI Meter

In noise-and-field-intensity measurement, Polarad has three major competitors. They are all competent. They build good equipment. Give or take an adjective, they describe and rate their equipment accurately. We respect their designs, their equipment, and their integrity as manufacturers. We believe they return that respect.

As an engineer, you know that no two design groups ever produce exactly the same instrument for a specific purpose. Experience and backgrounds differ. Approaches differ. Even basic concepts differ. Then, too, each group has its own view of the needs of the user. In a complex design, one approach will favor sensitivity over bandwidth, or, perhaps, cost over durability. We all must draw the line somewhere, in reaching each design decision. In an RI/FI meter, there are dozens of such decisions. The final "mix" of characteristics is, at best, an intelligent compromise . . . never the ultimate.

We believe that the two instruments described on this page represent by far the best "mix" of performance, economy, and versatility for the majority of applications.

You don't buy an RI/FI meter every day. Once you buy it, however, you may use it every day, for many years. Therefore, we urge you to consider the "mix" carefully. Limited range or restricted utility may seem tolerable now, but what about next month, or next year? Initial cost may dominate your thinking now, but how much does an extra man-hour a day (or one questionable result a week) cost . . . over five years?

Consider the "mix". We think you'll choose Polarad.

We can't resist listing the outstanding features of our "mixes", below — but don't decide until you have the complete technical data in front of you.

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- Digital Frequency Display
- —85 dbm Minimum Sensitivity
- 70 db Dynamic Range
- Standard Calibrated Horn Antennas
- Air Force Approved

FIM-2 "mix"

- Integral CW Generator for Accurate Calibration & Direct Substitution Measurements
- Separate Generator Output for Susceptibility Measurements
- Direct Reading without Charts
- —81 dbm Minimum Sensitivity
- Standard Calibrated Horn Antennas
- Air Force Approved

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15th Annual Conference on Engineering in Medicine and Biology

CONRAD HILTON HOTEL, CHICAGO, ILL., NOVEMBER 4–7, 1962

The 15th Annual Conference on Engineering in Medicine and Biology, sponsored by IRE, AIEE, and ISA, will be held at the Conrad Hilton Hotel, Chicago, Ill., on November 4–7, 1962. The program of the Conference is as follows:

Monday, November 5

Session I—Cardiovascular Physiology

Chairman: *A. M. Scher, University of Washington School of Medicine, Seattle, Wash.*

"The Future Direction of Cardiovascular Research," *H. R. Warner, Latter-Day Saints Hospital, Salt Lake City, Utah.; and R. F. Rushmer and A. M. Scher, University of Washington School of Medicine, Seattle, Wash.*

"A Simplified Pulsed Ultrasonic Flow. Meter for Use in Acute and Chronic Blood-flow Studies," *G. L. Mullins, F. L. Abel, and W. R. Guntheroth, University of Washington School of Medicine, Seattle, Wash.*

"A Newly Recognized Source of Zero Instability in the Sine Wave Electromagnetic Flow Meter," *A. Colvin and A. Warnick, Ford Motor Company, Dearborn, Mich.*

"Meaning of the Hydrodynamic Equations of Blood Flow," *R. S. Ledley, National Biomedical Research Foundation. Silver Spring, Md.*

"Derivation of an Electrical Delay Line as a Model of Oscillatory Flow in Arteries," *A. Noordergraaf, G. N. Jager and N. Westerhof, University of Utrecht, Holland.*

"An Electrohydraulic Model for the Investigation of Cardiovascular Hemodynamics," *P. B. Hollander, Radio Corporation of America, Princeton, N. J.*

Session II—Neurophysiology

Chairman: *Mary Brazier, University of California, Los Angeles, Calif.*

"Computer Systems in a Surgical Setting," *B. S. Rosner, West Haven V.A. Hospital and Yale University School of Medicine, New Haven, Conn.*

"Analysis of Control of Eye Movement in Man by Means of Autocorrelation and Crosscorrelation," *J. Barlow, Massachusetts General Hospital, Boston, Mass.*

"An Analog Cross-Correlator for the Analysis of Nervous Activity—Theory and Instrumentation," *J. U. Casby, R. Siminoff, and T. R. Houseknecht, Eastern Pennsylvania Psychiatric Institute, Philadelphia, Pa.*

"An Analog Cross-Correlator for the Analysis of Nervous Activity—Analysis of a Peripheral Sensory System," *R. Siminoff, J. U. Casby, and T. R. Houseknecht, Eastern Pennsylvania Psychiatric Institute, Philadelphia, Pa.*

"An Analog Cross-Correlator for the Analysis of Nervous Activity—General Applications," *T. R. Houseknecht, J. U. Casby, and R. Siminoff, Eastern Pennsylvania Psychiatric Institute, Philadelphia, Pa.*

"A Remote Control Brain Stimulator System," *H. Warner, General Electric Company.*

"Analysis and Synthesis of Evoked Cortical Voltages Using Parameter Variation in the Neuronal Population," *G. G. Furman, University of California, Berkeley, Calif.*

"Visual Information Processing in the Beetle Lixus," *J. C. Bliss, Stanford Research Institute, Menlo Park, Calif.*

"Analytical Studies of Electroencephalograms of Vitamin Deficient Chicks," *E. C. Lowenberg, F. L. Siegel, C. R. Griffin, Jr., and R. W. Hurt, University of Texas, Austin, Tex.*

Session III—Computer Applications

Chairman: *Raymond Jonnard, Prudential Insurance Company of America, Newark, N. J.*

"Heart Sound Analysis by Digital Computer," *D. S. Gerbarg and J. Hefler, IBM Corporation, Yorktown Heights, N. Y.*

"Digital Computer Simulation of Tracer Experiments with the Two-Compartment Open System," *M. L. Shore and R. Callahan, Wood V.A. Center and Marquette University School of Medicine, Milwaukee, Wis.*

"Resolution of Multicomponent Exponential Decay Curves Using Fourier Transforms," *D. G. Gardner, Illinois Institute of Technology, Chicago, Ill.*

"Accuracy in Recording of Physiological Data on Magnetic Tape," *C. A. Steinberg and Cutler Hammer, Airborne Instruments Laboratory, Deer Park, N. Y.; and C. A. Caceres, Heart Disease Control Program, U. S. Public Health Service, Washington, D. C.*

"Medical Studies on Former Radium Workers," *H. W. Fisher, N. Y. State Radiation Laboratory, West Orange, N. J.*

"A Scheme for Programming Radioactive Iron Kinetic Data," *L. Sharney, L. R. Wasserman, and L. Shwartz, Mt. Sinai Hospital, New York, N. Y. and Systems Research Group, Mineola, N. Y.*

"Automating a Hospital Information System," *L. Gross and E. Myer, System Development Co., Santa Monica, Calif.*

"Conditional Probabilities in Medical Diagnosis," *R. S. Ledley and E. Oestreich, National Biomedical Research Foundation, Silver Spring, Md.; and L. B. Lusted, University of Rochester Medical School, Rochester, N. Y.*

"Probabilistic Logic and the Algebra of Functors Applied to Symptom-Disease Complexes," *S. Swisher, University of Rochester Medical School, Rochester, N. Y.; and E. Agalides, General Dynamics Corporation, Rochester, N. Y.*

Monday Afternoon

Session IV—Cardiovascular Physiologic Measurements

Chairman: *E. H. Wood, Mayo Foundation Graduate School, University of Minnesota, Rochester, Minn.*

"Physical Methods for Measurement of Oxygen in Blood"

I—"Oximetry," *M. L. Polanyi, American Optical Company, South Bridge, Mass.*

II—"The Oxygen Electrode and Polarographic Technics," *L. C. Clark, University of Alabama Medical Center, Birmingham, Ala.*

III—"Gas Chromatographic Technics," *T. Johns, Beckman Instruments, Inc., Fullerton, Calif.*

"An Acoustical Phaseometric Method for Gas Analysis," *G. Owen and D. W. Cugell, Northwestern University, Chicago, Ill.*

"Evaluation of Mathematical Models Relating Pressure and Flow in the Ascending Aorta," *G. O. Barnett and R. L. Van Citters, University of Washington School of Medicine, Seattle, Wash.*

"A New System for Measurement of Blood Flow by Ultrasonics," *F. W. Noble, T. C. Goldsmith, P. W. Cook, and C. J. Waldspurger, National Heart Institute, Bethesda, Md.*

"Continuous Monitoring of Cardiovascular System," *G. Sullivan, G. Weltman, and D. Bredon, Spacelabs, Inc., Van Nuys, Calif.*

"Progress in the Use of Nuclear Magnetic Resonance for Measuring Blood Flow," *V. Kudravcev, National Heart Institute, Bethesda, Md.*

"Heart Sound Pickup," *D. J. Coleman, J. R. Whiteman, and C. A. Caceres, Heart Disease Control Program, U. S. Public Health Service, Washington, D. C.*

Session V—Instrumentation, General

Chairman: *M. L. Petrovick, Northwestern University Medical School, Chicago, Ill.*

"The Electrokinetic Flow Meter," *S. R. Topaz and S. D. Moulopoulos, Cleveland Clinic, Cleveland, Ohio.*

"Systolic and Diastolic Blood Pressure on a Finger," *M. Traite and L. Germain, Gulton Industries, Inc., Metuchen, N. J.; and A. W. Pratt, National Institutes of Health, Bethesda, Md.*

"An Ultra Micro Freezing Point Apparatus," *R. L. Bowman and D. J. Prager, National Heart Institute, Bethesda, Md.*

"New Velocity Transducer for Physiological Sounds," *E. Reickert, Shure Brothers, Evanston, Ill.; and M. S. Paul, Children's Memorial Hospital, Chicago, Ill.*

"An Improved Method of Registering Respiration Rate," *K. E. Robins and A. R. Marko, 6570th Aerospace Medical Research Laboratories, Wright-Patterson AFB, Ohio*

"A Linear Pneumotachometer," *R. S. Scranton, Technology Incorporated, Dayton, Ohio.*

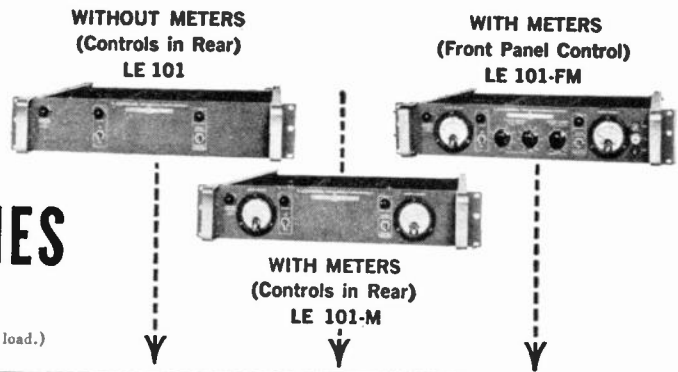
"A Photoelectric Tremometer," *E. L. Thomas, The Ontario Hospital, New Toronto and University of Toronto, Canada.*

"An Electrical Strain Gage Dynamometer for Muscle Use," *R. A. Perrin, U. S. Air Force, Wichita Falls, Tex.; and E. H. Harris and E. T. Haslam, Tulane University, New Orleans, La.*

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"Equipment for the Evaluation of Servo Mechanisms Controlling Muscle," *R. G. Black, R. S. Hayward, and J. W. Scott, University of Toronto, Toronto, Canada.*

"Detection of Glossal Motion Patterns," *M. L. Petrovick, Northwestern University Medical School, Chicago, Ill.*

Session VI—Data Storage and Retrieval

Chairman: *O. H. Schmitt, University of Minnesota, Minneapolis, Minn.*

"Data Storage and Retrieval"

Panel Members: *O. H. Schmitt, University of Minnesota, Minneapolis, Minn.; G. N. Webb, The Johns Hopkins University School of Medicine, Baltimore, Md.; C. A. Caceres, Heart Disease Control Program, U. S. Public Health Service, Washington, D. C.*

"A Poor-Man's Data System," *F. R. Sias, Jr., The Johns Hopkins University, Baltimore, Md.*

"Storage and Retrieval of Biomedical Data," *E. Ackerman, J. W. Rosevear, and J. L. Juergens, Mayo Clinic, Rochester, Minn.*

"Biomedical Information Retrieval by Means of Tabledex," *F. S. Zusman and R. S. Ledley, National Biomedical Research Foundation, Silver Spring, Md.*

"An Advanced Data Collection System for Bioastronautic Research and Development Projects," *H. T. Castillo and C. E. Hoefner, Space Labs Incorporated, Van Nuys, Calif.*

"Digital Conversion and Readout of Physiologic Data," *A. R. Kahn, Beckman Instruments, Inc., Schiller Park, Ill.*

"Data Optimizing Prior to Recording Via Analog Computer," *J. H. Trageser, Technology Incorporated, Dayton, Ohio.*

Monday Evening

Banquet

Introductory Remarks: *L. N. Katz, Director Cardiovascular Institute, Michael Reese Hospital, Chicago, Ill.; Professorial Lecturer in Physiology, University of Chicago.*

Keynote Address: *Boisfeuillet Jones, Special Assistant to the Secretary, Health and Medical Affairs, U. S. Department of Health, Education and Welfare.*

Tuesday, November 6

Session VII—Physiologic Measurements

Chairman: *R. F. Rushmer, University of Washington School of Medicine, Seattle, Wash.*

"A Comprehensive Cardiovascular Research Facility," from the Department of Physiology and Biophysics, University of Washington School of Medicine, Seattle, Wash.

I—"Objectives and Approach," (Motion Picture) *R. F. Rushmer.*

II—"Cardiovascular Recording System"

a—"Blood Flow Recording Techniques," *W. A. Schlegel and D. L. Franklin.*

b—"Dimension Recording Techniques," *N. W. Watson.*

c—"Measured Cardiac Variables and Computed Functions," *D. C. Harding.*

d—"Integrated Recording System," *D. W. Baker.*

III—"Microcirculation Recording System," *C. A. Wiederhielm.*

IV—"Rheology Research" (Motion Picture), *J. E. Meisner and R. F. Rushmer.*

V—"Training Program in Cardiovascular Research Techniques," *R. F. Rushmer.*

"FM Sonic Gauge for Continuous Measurement of Organ Dimensions in Ambulatory, Experimental Animals," *D. L. Franklin, Scripps Clinic and Research Foundation, La Jolla, Calif.; and N. W. Watson, W. A. Schlegel, and R. F. Rushmer, University of Washington School of Medicine, Seattle Wash.*

"Cinesonography as Applied to Biological Problems," *J. E. Jacobs, D. W. Cugell, and J. R. Wennemark, Northwestern University, Chicago, Ill.*

"Ultrasonic Echo Imaging in Eye Research," *R. S. Mackay, G. Eilers, J. Horowitz, and E. Marg, University of California, Berkeley, Calif.*

"Microwave Absorption Cross Section Studies of Man with Multi-Layer Phantoms," *A. Anne, O. M. Salati, and H. P. Schwan, University of Pennsylvania, Philadelphia, Pa.*

"Display Techniques for Physiological Monitoring Over Extended Periods of Time," *G. N. Webb, R. E. Rogers, and P. R. Thorne, Johns Hopkins University School of Medicine, Baltimore, Md.*

"Experimental Determination of Whole Body Animal Irradiation Distributions Using Miniature Radio-Photoluminescent Glass Rods," *S. J. Malsky, C. G. Amato, V. P. Bond, and B. Roswit, Brookhaven National Laboratory, AMF Atomics, Greenwich, Conn., V. A. Hospital, Bronx, N. Y.*

Session VIII—Functional Organization of the Central Nervous System and Sensory Receptors

Chairman: *R. W. Jones, Northwestern University Institute of Technology, Evanston, Ill.*

"Some Applications of Computer Analysis to the Electrical Activity of the Brain," *M. A. B. Brazer, University of California, Los Angeles, Calif.*

"Equipment for the Analysis of Information Transmitted by the Nerve Fibers of the Retina," *G. B. Gerace and G. Gestri, University of Pisa, Italy.*

"Signal Transmission Properties of the Cat Retina," *C. Enroth-Cugell and R. W. Jones, Northwestern University, Evanston, Ill.*

"A Neuron Model which Performs Analog Functions," *D. R. Boyle, National Bureau of Standards, Washington, D. C.*

"A Circuit for Simulating Nerve Response Based on Hodgkin-Huxley Principles," *N. S. Bromberg, Brookline, Mass.*

"Discontinuous Biological Control—The Eye Movement System," *L. R. Young and L. R. Stark, Massachusetts Institute of Technology, Cambridge, Mass.*

"Learning Phenomena During Tracking of Eye Motions," *P. Dallos, Northwestern University, Evanston, Ill.*

Session IX—Computers in Research

Chairman: *H. R. Warner, Latter-Day Saints Hospital, Salt Lake City, Utah.*

"On the Operation of a Biomedical Computer Facility"

Panel Discussion.

"A Pulse Interval Data Reduction System," *G. R. Paulsen, International Business Machines Corporation, Evanston, Ill.*

"Programmed Timer for Light Stimuli," *R. L. Schoenfeld and N. Milkman, The Rockefeller Institute, New York, N. Y.*

"The Use of Computers in Protein Bio-Chemistry," *M. O. Dayhoff and R. S. Ledley, National Biomedical Research Foundation, Silver Spring, Md.*

"Applications of a Method of Pattern Recognition to Biomedical Problems," *K. Enslein, Brooks Research, Inc., East Rochester, N. Y.*

"Digital Computer Analysis of Psycho-Physiologic Information," *H. Zimmer, The University of Georgia, Athens, Ga.*

"Techniques Employed in the Automation of the WGTA—The Wisconsin Automatic Test Apparatus (WATA)," *V. J. Polidora, University of Wisconsin, Madison, Wis.*

Tuesday Afternoon

Session X—Electrophysiology

Chairman: *B. F. Hoffman, State University of New York, Downstate Medical Center, Brooklyn, N. Y.*

"Current Status of Electrophysiology of Skeletal Muscle," *W. L. Nastuk, Columbia University, New York, N. Y.*

"Current Status of Electrophysiology of Cardiac Muscle," *A. Paes de Carvalho, Institute of Biophysics, University of Brazil.*

"Techniques for Study of Excitation-Contraction Coupling in Cardiac Muscle," *F. Kavaler, State University of New York, Downstate Medical Center, Brooklyn, N. Y.*

"Threshold Electrical Requirements for Cardiac Stimulation," *D. T. Nash, J. F. Neville, Jr., and M. A. Grossman, State University of New York, Upstate Medical Center, Syracuse, N. Y.; and L. D. Wechsler, General Electric Company, Syracuse, N. Y.*

"Microminiature Passive Telemetry for Biological Voltages," *W. Honig, Honig Laboratories, Inc., Brooklyn, N. Y.*

"Implanted Two-Way Telemetry," *I. J. Young and W. S. Naylor, Neurology Service, VARH and Northwestern University, Chicago, Ill.*

"Computer Aids to Depth Recording and Stimulation of the Human Brain," *R. G. Bickford, M. J. Marsden, and C. S. McCarthy, Mayo Foundation and International Business Machine Corporation, Rochester, Minn.*

"Phase Angle as a Factor in Pulpal Excitation," *R. C. Weiss and H. S. Brenman, University of Pennsylvania, Philadelphia, Pa.*

Session XI—Telemetry

Chairman: *L. A. Geddes, Baylor University Medical College, Houston, Tex.*

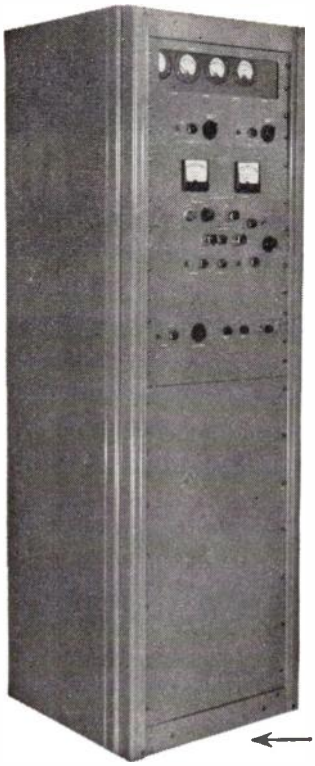
Chairman's Introduction.

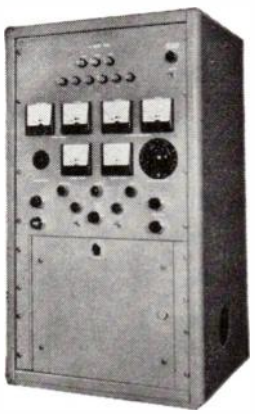
"Telemetry of Respiration and Electrocardiogram from a Single Pair of Electrodes," *H. Hanish, Bio-Com Incorporated, Los Angeles, Calif.*

"A Personalized Radio Telemetry System for Studying CNS Functions in Aerospace Flight," *D. G. Simons, School of Aerospace Medicine, Brooks AFB, Tex.*

"Telemetry in Clinical and Experimental Cardiovascular Physiology," *H. T. Jacobsmeyer, Emerson Electric Manufacturing Co., St. Louis, Mo.; and T. Cooper, St. Louis University School of Medicine, St. Louis, Mo.*

"Problems Encountered in Telemetering of Physiological Activity of Schizophrenics on an Experimental Psychiatric Ward," *B. Tursky, L. H. Turner, and L. Grinspoon, Massachusetts Mental Health Center, Boston Psychopathic Hospital, Boston, Mass.*





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"A Pulse-Rate Modulation System for Bio-Medical Radiotelemetry," *A. G. Potter, Iowa State University, Ames, Iowa.*

"A New Concept of Intensive-Care Patient Monitoring," *J. T. Burroughs and C. C. Miller, ITT Federal Laboratories, San Fernando, Calif.*

"A Hybrid Data Processing System for Respiratory Studies," *T. W. Murphy, Memorial Sloan-Kettering Cancer Center, New York, N. Y.*

Session XII—Automation in Surgery and Anaesthesia

Chairman: *F. J. Lewis, Northwestern University Medical School, Chicago, Ill.*

"Myocardial and Circulatory Assistance by Synchronized Counterpulsation," *D. E. Harken and A. A. Lefemine, Peter Bent Brigham Hospital and Harvard Medical School, Boston, Mass.*

"Total Left Heart Bypass as a Means of Extracorporeal Circulatory Support of Myocardial Infarction with Shock," *C. Dennis, S. A. Wesolowski, W. Dock, E. Carlens, M. Blumenfeld, R. R. Cappelletti, and C. Grosz, State University of New York, Downstate Medical Center, Brooklyn, N. Y., and Karolinska Institute, Stockholm, Sweden.*

"Assisted Circulation by Counter-Pulsation with an Intra-Aortic Balloon—Methods and Effects," *R. H. Clauss, P. Missier, D. A. Tice, and G. F. Reed, New York University School of Medicine, New York, N. Y.*

"Diastolic Augmentation by External Pulsed Pressure," *J. J. Osborn, M. Russi, A. Salel, M. L. Bramson, and F. Gerbode, Presbyterian Medical Center, San Francisco, Calif.*

"A Comparative Evaluation of Three Disposable Heat Exchangers for Profound Hyperthermia," *W. G. Esmond, J. Weaver, and S. Attar, University of Maryland School of Medicine, Baltimore, Md.*

"Pressure Drop for the Flow of Blood in Thin Sheets," *J. E. Kipp, R. G. Nevins, C. H. Sprague, and R. R. Berns, Kansas State University, Manhattan, Kan.; and C. F. Kittle, University of Kansas, Kansas City, Kan.*

"The Clinical Applications of Telecardiography," *D. A. Davis, T. C. Gibson, and W. Thornton, The University of North Carolina and The North Carolina Memorial Hospital, Chapel Hill, N. C.*

"Multiple-Electrode System for Measurement of *pH*, *pO2* and *pCO2* in Blood," *D. E. Judevine, University of Pennsylvania, Philadelphia, Pa.*

"The Army Artificial Heart Pump," *K. E. Woodward and T. G. Barila, Diamond Ordnance Fuze Laboratories, Washington, D. C., and Walter Reed Army Institute of Research, Washington, D. C.*

Tuesday Evening

"Autoanalysis of the Electrocardiogram and Phonocardiogram."

Moderator: *C. A. Caceres, Heart Disease Control Program, Public Health Service, U. S. Department of Health, Education and Welfare, Washington, D. C.*

"Membrane Ionics"

Moderator: *B. F. Hoffman, State University of New York, Downstate Medical Center, Brooklyn, N. Y.*

"Flow Meters"

Moderator: *P. Spencer, Bowman Gray School of Medicine, Winston-Salem, N. C.*

"Physiologic Telemetry"

Moderator: *L. A. Geddes, Baylor University Medical College, Houston, Tex.*

"Pumpoxygenators"

Moderator: *G. H. A. Clowes, Jr., Medical College of South Carolina, Charleston, S. C.*

"Monitoring Respiration—Remote and Contact Sensing Devices"

Moderator: *D. W. Cugell, Northwestern University Medical School, Chicago, Ill.*

"Electrical Activity of the Heart"

Moderator: *A. M. Scher, University of Washington School of Medicine, Seattle, Wash.*

"Biologic Applications of Ultra Sound"

Moderator: *F. J. Fry, University of Illinois, Urbana, Ill.*

"Computers in Patient Diagnosis"

Moderator: *R. S. Ledley, National Biomedical Research Foundation, Silver Spring, Md.*

"Bionics"

Moderator: *Jack Steele, Major, USAF, Wright-Patterson Air Force Base, Ohio.*

"Neurophysiology"

Moderator: *W. S. McCulloch, Massachusetts Institute of Technology, Cambridge, Mass.*

"Biomedical Training and Qualifications"

Moderator: *J. H. U. Brown, Division of General Medical Sciences, U. S. Department of Health, Bethesda, Md.*

Wednesday, November 7

Session XIII—Electrocardiography and Vectorcardiography

Chairman: *S. A. Briller, American Heart Association and University of Pennsylvania School of Medicine, Philadelphia, Pa.*

"The Status of Electrocardiography—Potential Theory," *D. B. Geselowitz, University of Pennsylvania, Philadelphia, Pa.*

"The Status of Electrocardiography—Information Theory and Computer Applications," *F. W. Stallman, Veterans Administration Hospital, Washington, D. C.*

Discussion

"Estimate of Information in an Electrocardiac Signal," *W. K. Hagan and M. Horowitz, The Magnavox Company, Fort Wayne, Ind.*

"Computer Analysis of the Electrocardiogram," *W. Thornton, T. C. Gibson, and D. A. Davis, The University of North Carolina and the North Carolina Memorial Hospital, Chapel Hill, N. C.*

"Signal Analysis by a Digital Computer Using Orthonormal Exponentials," *T. Y. Young and W. H. Huggins, The Johns Hopkins University, Baltimore, Md.*

"On the Statistical Analysis of Electrocardiograms," *T. Y. Young and W. H. Huggins, The Johns Hopkins University, Baltimore, Md.*

"A New Approach to Electrocardiography," *P. F. Stanley, Purdue University, Lafayette, Ind.*

"Cardiac Dipole Locator (More Than Vectorcardiography)," *I. Zakarias Picofarad Electronic Instrumentation, Beacon, N. Y.*

Session XIV—Biologic Control Systems

Chairman: *F. S. Grodins, Northwestern University Medical School, Chicago, Ill.*

"A Physiologist Learns About Control System Theory," *J. S. Gray, Northwestern University Medical School, Chicago, Ill.*

"A Control System Engineer Learns About Physiology," *R. W. Jones, Northwestern Institute of Technology, Evanston, Ill.*

"An Engineering Model of Motor Coordination in Man," *J. Houk, Electronic Systems Laboratory and Biology Department; and L. Stark, Massachusetts Institute of Technology, Cambridge, Mass.*

"Physiological Control System for Ocular Stabilization," *G. Melville-Jones and J. H. Milsum, McGill University, Montreal, Canada.*

"Information Theory and Bio-Rhythms," *R. Bernhard, Grumman Aircraft Co., Bethpage, N. Y.*

"Spectral Analysis of Pulmonary Ventilation in Quiescent Man," *L. Goodman, Case Institute of Technology, Cleveland, Ohio.*

"Simulation of the Chemical and Circulatory Factors in the Dynamic Analysis of the Human Respiratory Control System," *J. D. Horgan and R. L. Lange, Marquette University, Milwaukee, Wis.*

"Servoanalysis of Blood Pressure Control Mechanism," *A. M. Scher, University of Washington School of Medicine, Seattle, Wash.*

"Respiratory Regulation of Heart Rate," *N. A. Coulter, Jr. and A. Angelone, The Ohio State University, Columbus, Ohio.*

Session XV—Artificial Organs

Chairman: *W. J. Kolff, Cleveland Clinic, Cleveland, Ohio.*

"A Servomechanism to Drive an Artificial Heart Inside the Chest," *K. W. Hiller, W. Seidel, and W. J. Kolff, NASA, Lewis Research Center and the Cleveland Clinic Foundation, Cleveland, Ohio.*

"External Recording of Performance of Artificial Hearts Inside the Chest," *S. R. Topaz and W. J. Kolff, Cleveland Clinic Foundation, Cleveland, Ohio.*

"Physiology and Pathophysiology of Hearts Inside the Chest," *T. Akutsu, K. W. Hiller, V. Mirkovitch, K. Mohr, and W. J. Kolff, Cleveland Clinic Foundation, Cleveland, Ohio.*

"Artificial Heart with Antivacuum Balloon," *H. S. Norton, Thompson Ramo Wooldridge, Inc., Cleveland, Ohio; and T. Akutsu and W. J. Kolff, Cleveland Clinic Foundation, Cleveland, Ohio.*

"Engineering Aspects of the Artificial Heart," *R. G. Burney, W. S. Pierce, M. H. Boyer, R. W. Driscoll, and C. K. Kirby, University of Pennsylvania School of Medicine, Philadelphia, Pa.*

"Theoretical Limits of Efficiency for Solenoids to Drive Artificial Hearts," *J. Bindels, Cleveland Clinic Foundation, Cleveland, Ohio.*

"Excitation of a Large Coil Set for the Transport of Energy into the Chest," *J. C. Schuder, H. E. Stephenson, Jr., and B. A. Rowley, University of Missouri Medical School, Columbia, Mo.*

"A Method for the Determination of Zeta Potentials of Selected Materials in Contact with Blood," *R. E. Beck, V. Mirkovitch, P. G. Andrus, and R. I. Leininger, Battelle Memorial Institute, Columbus, Ohio, and Cleveland Clinic, Cleveland, Ohio.*

"Experimental Study on Emptying of the Cord Bladder—Transcutaneous Stimulation of Pelvic Nerves by Electromagnetic Induction Apparatus, Therapeutic Deductions," *T. H. Burghele, V. Ichim, and M. Demetrescu, Bucharest, Rumania.*

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Wednesday Afternoon

Session XVI—Neurophysiology and Instrumentation

Chairman: *C. N. Woolsey, University of Wisconsin, Madison, Wis.*

"Use of RF Powered Receiver for Evacuation of the Neurogenic Bladder," *W. E. Bradley, R. Wingrove, and S. N. Chou, University of Minnesota Hospitals, Minneapolis, Minn.*

"High Intensity Focused Ultrasound and its Action on the Central Nervous System," *F. J. Fry, University of Illinois, Urbana, Ill.*

"Electro-Encephalometer," *D. G. Aviv and B. L. Pacella, Columbia University, New York, N. Y.*

"Transmission of Brain Waves Through a Volume Conductor," *H. R. Pantek, University of Michigan, Ann Arbor, Mich.*

"Engineering Considerations in GSR Measurements," *W. H. Rickles, Jr., School of Aerospace Medicine, Aerospace Medical Division, Brooks AFB, Tex.*

"An Improved Galvanic Skin Resistance Measuring Device," *A. R. Marko, Aerospace Medical Research Laboratories, Wright-Patterson AFB, Ohio.*

"A Stimulation-Rise Control Unit for Neural Excitation—Progress Report," *R. C. Weiss and H. S. Brenman, University of Pennsylvania, Philadelphia, Pa.*

"A Transistorized Plus and Minus Pulse Pair Generator for Brain Stimulation," *J. W. Keller, Jr., Cordis Corporation, Miami, Fla.*

"Some Aspects of Decreased Motor Conduction Velocity in the Ulnar Nerve of Man," *R. G. Baginsky and J. E. Ammons, Physical Medicine and Rehabilitation Service, Veterans Administration Hospital, Bedford, Mass.*

"White Sound as an Analgesic Agent During Electrical Excitation of the Anterior-Superior Alveolar Nerve," *H. S. Brenman, University of Pennsylvania, Philadelphia, Pa.*

Session XVII—Electrocardiogram and Pacemakers

Chairman: *S. D. Larks, Marquette University, Milwaukee, Wis.*

"Electrocardiogram and Pacemakers"
Panel discussion.

"Clinical Experience with an Implantable Externally Controllable Cardiac Pacemaker," *A. Kantrowitz, College of Medicine, State University of New York, New York, N. Y.*

"An Implantable Synchronous Cardiac Pacemaker," *J. W. Keller, Cordis Corporation, Miami, Fla.; and D. A. Nathan and S. Center, Mt. Sinai Hospital, Miami, Fla.*

"Theoretical Analysis of a Micromodule Pacemaker Receiver," *J. C. Schuder, H. Stoeckle, and L. D. Muench, University of Missouri School of Medicine, Columbia, Mo.*

"Regional FM Phone Data Transmission—A New Inter-University Research Concept," *S. D. Larks, Marquette University, Milwaukee, Wis.*

"The Effect of Band Limiting on Improving the Reliability of Fetal Electrocardiotachometry," *N. P. Thompson, Palo Alto Medical Clinic, Palo Alto., Calif.*

Session XVIII—Physiologic Monitoring of Patients

Chairman: *Raymond Jonnard, Prudential Insurance Company of America, Newark, N. J.*

Panel Members: *G. Radcliffe, Columbus Hospital, Newark, N. J.; R. W. Harrington, Systems Development Company, Santa Monica, Calif.; M. Woodbury, New York University, New York, N. Y.; K. C. Rock, Minneapolis-Honeywell Company, Long Island City, N. Y.; P. Salisbury, Saint Joseph Hospital, Burbank, Calif.; N. Mageoch, General Precision Inc., Clifton, N. J.; J. Schenthal, Tulane University School of Medicine, New Orleans, La.; J. Sweeney, Tulane University, New Orleans, La.; and L. Barrer, New Jersey State Department of Health, West Orange, N. J.*

Northeast Electronics Research and Engineering Meeting

COMMONWEALTH ARMORY AND SOMERSET HOTEL, BOSTON, MASS., NOVEMBER 5–7, 1962

The Sixteenth Northeast Electronics Re-Research and Engineering Meeting (NEREM'62) will be held on November 5–7, 1962, at the Somerset Hotel and Commonwealth Armory, Boston, Mass. Over 150 scientists and engineers are represented on a 25-session technical program involving over 100 papers. Ernst Weber, President of Polytechnic Institute of Brooklyn, will deliver the keynote address on "The Shape of Things to Come." Julius P. Molnar, Executive Vice President of Bell Telephone Laboratories, Inc., will give the banquet speech entitled, "Communications Across Ocean Expanses." In addition, there will be an exhibition of the latest electronic developments.

Illustrated condensations of all the papers will be published in the *NEREM Record*, available free of charge to IRE members attending the Conference. Nonmembers at the Conference may receive a copy for $2.00. Post-Conference copies, priced at $7.50 each, will be available from: IRE Boston Section, 313 Washington St., Newton 58, Mass.

The technical program of the Conference is as follows:

Monday, November 5

Session 1—Antennas and Propagation I

Chairman: *J. Ruze, Radiation Engineering Laboratory, Maynard, Mass.*

"Electronically-Scanned Antenna Systems," *M. A. Diab and T. Maggio, Rome Air Development Center, Griffiss AFB, N. Y.*

"Multiple Beam Constrained Lens," *J. L. MacFarland, Hughes Aircraft Co., Fullerton, Calif.*

"H-Guide," *G. N. Voronoff, Dalmo Victor Co., Belmont, Calif.*

"Modulated Array for Space Communications," *E. M. Rutz and E. Kramer, Fed. Sys. Div., IBM Corp., Rockville, Md.*

"A Circularly Polarized, Omnidirectional, Helix Antenna," *C. Brown, Sylvania Electronic Systems, Waltham, Mass.*

Monday Afternoon

Session 2—Audio and Acoustics

Chairman: *D. von Recklinghausen, H. H. Scott, Inc., Maynard, Mass.*

"Voice Communication," *L. Dolansky, Northeastern University, Boston, Mass.*

"Speech Compression Studies," *C. P. Smith, Air Force Cambridge Research Labs., Bedford, Mass.*

"New Concepts in Sound System Design," *D. L. Klepper, Bolt Beranek and Newman, Inc., Cambridge, Mass.*

"Music, Acoustics and the Listener," *L. L. Beranek, Bolt Beranek and Newman, Inc., Cambridge, Mass.*

"A Full-Range Electrostatic Speaker," *A. A. Janszen, KLH Research and Development Corp., Cambridge, Mass.*

Session 3—Highlights of Army Communications and Electronics R and D

Chairman: *H. K. Ziegler, U. S. Army Signal Research and Development Lab., Fort Monmouth, N. J.*

"Small Signal Laser Amplifier Design Theory," *H. Jacobs, L. Hatkin and F. A. Brand, U. S. Army Signal Research and Development Lab., Fort Monmouth, N. J.*

"An Electronically Controlled Formant Synthesizer," *J. L. DeClerk, D. L. Phyfe and R. A. Fisch, U. S. Army Signal Research and Development Lab., Fort Monmouth, N. J.*

"New Laser Technique for Ranging Application," *R. C. Benson, R. O. Godwin and M. R. Mirarchi, U. S. Army Signal Research and Development Lab., Fort Monmouth, N. J.*

"Progress in Microwave Components," *N. Lipetz, U. S. Army Signal Research and Development Lab., Fort Monmouth, N. J.*

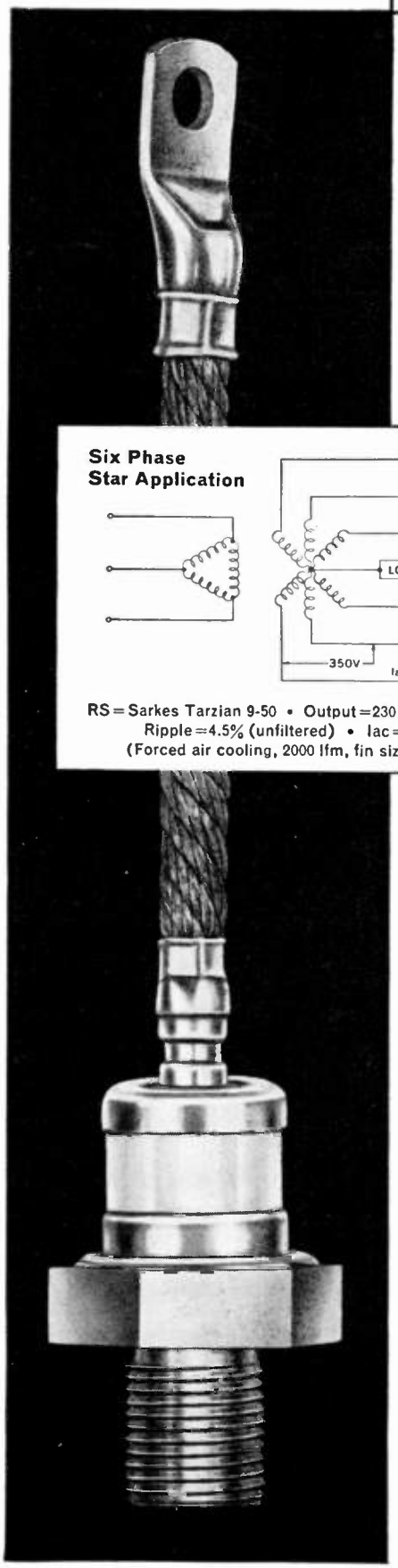
"Geophysical Factors Affecting Electromagnetic Propagation Through the Lithosphere," *S. B. Levin, U. S. Army Signal Research and Development Lab., Fort Monmouth, N. J.*

Session 4—Microwave Solid State

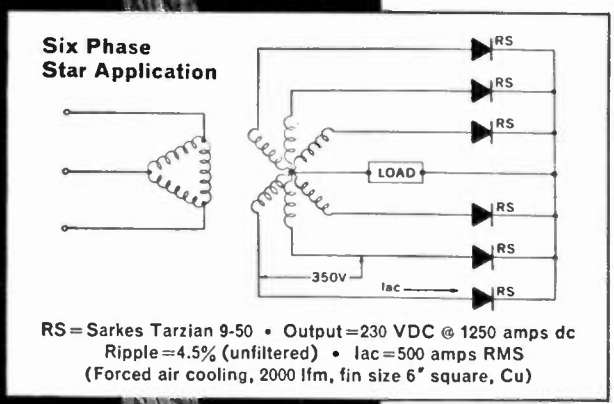
Chairman: *R. Damon, Microwave Associates, Inc., Burlington, Mass.*

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ST9-20	250	200	140	1.2	50
ST9-30	250	300	210	1.2	50
ST9-40	250	400	280	1.2	50
ST9-50	250	500	350	1.2	50
ST9-60	250	600†	420	1.2	50

*Add N for negative, P for positive, base polarity
†Other 600 PIV units available for 0.5, 0.75, 1.5, 2, 6, 10, 12, 20, 35, 50, 100, 150, and 200 ampere requirements

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ceivers and Systems," *S. Okwit, Airborne Instruments Lab., Deer Park, N. Y.*

"Ultrasonic Traveling-Wave Amplification in Piezoelectric Semiconductors," *A. R. Hutson, Bell Telephone Laboratories, Inc., Murray Hill, N. J.*

"Microwave Ultrasonic Devices and Materials," *R. T. Denton and E. G. Spencer, Bell Telephone Laboratories, Inc., Murray Hill, N. J.*

"Antiferromagnetic Materials and Their Application at Millimeter and Submillimeter Wavelengths," *G. S. Heller, M.I.T., Lincoln Laboratory, Lexington, Mass.*

Session 5—Radio Astronomy

Chairman: *S. Goldstein, Harvard College Observatory, Cambridge, Mass.*

"Radar Studies of the Lunar Surface," *J. V. Evans, M.I.T., Lincoln Laboratory, Lexington, Mass.*

"Some Practical and Theoretical Aspects of Power Spectrum Measurements," *M. J. Levin, M.I.T., Lincoln Laboratory, Lexington, Mass.*

"Advances in the Techniques of Measuring 21-cm Spectra," *V. Radhakrishnan, California Institute of Technology, Pasadena, Calif.*

"Some Applications of Optical Radar to Astronomy and Geophysics," *G. Fiocco, Research Laboratory of Electronics, M.I.T., Cambridge, Mass.*

Monday Evening

Session 6—Electronics in Outer Space

Chairman: *K. C. Black, Scientific Analysis Corp., Sudbury, Mass.*

"Tracking and Control of Near Space Satellites," *I. A. Getting, Aerospace Corp., El Segundo, Calif.*

"Active Communication Satellites," *S. H. Reiger, The RAND Corp., Santa Monica, Calif.*

"Communications and Radar for Deep Space," *W. H. Radford, M.I.T., Lincoln Laboratory, Lexington, Mass.*

"Reliability Problems of Electronics in Space," *S. Herwald, Westinghouse Electric Corp., Baltimore, Md.*

Tuesday, November 6

Session 7—Antennas and Propagation II

Chairman: *L. J. Ricardi, M.I.T., Lincoln Laboratory, Lexington, Mass.*

"Feed Design for Large Antennas," *S. J. Ajoika, Hughes Aircraft Co., Fullerton, Calif.*

"Spherical Antenna Feed Designs," *F. S. Holt, E. Bouche, A. C. Schell, E. Altshuler, and W. Mavroides, A. F. Cambridge Research Laboratory, Bedford, Mass.*

"Experimental Evaluation of a 1000-Wavelength Antenna," *V. L. Lynn, K. J. Keeping and W. D. Fitzgerald, M.I.T., Lincoln Laboratory, Lexington, Mass.*

"Field Alignment and Calibration of a Multi-Feed High-Gain Antenna System," *J. M. Moore and R. Rubin, Sylvania Electronic Systems, Waltham, Mass.*

"Structural Aspects of Large Precision Antennas," *H. Simpson, Simpson, Grumpretz and Heger, Inc., Cambridge, Mass.*

Session 8—Bio-Medical Electronics I

Chairman: *A. T. Kornfield, The Biosearch Co., Boston, Mass.*

"Bionics and Physical Analogies of Biosensor Systems," *J. Steele, Aerospace Medical Div., Wright Patterson AFB, Ohio.*

"Electromechanical Analogues in Prosthetic Systems," *R. Contini, New York University, New York, N. Y.*

"Life Support Design Bounds for Equipment vs Men," *E. Konecci, NASA, Washington, D. C.*

Session 9—New Advances in Specification of Component Reliability

Chairman: *E. F. Dortinger, Raytheon Co., Waltham, Mass.*

"Problems of High Reliability Specifications," *O. B. Moan and H. C. Werner, Lockheed Missiles and Space Co., Sunnyvale, Calif.*

"Philosophy and Guidelines for Semiconductor Reliability Assurance," *V. K. Erickson, Minneapolis-Honeywell Regulator Co., Minneapolis, Minn.*

"Lot-By-Lot and Process Control for Reliability," *F. Ruther, Defense Electronic Supply Center, Dayton, Ohio.*

"The EIA Established Reliability Specification Program and its Relation to the Armed Services and Department of Defense," *L. Podolsky, Sprague Electric Co., North Adams, Mass.*

Session 10—Quantum Electronics I

Chairman: *C. F. Luck, Jr., Raytheon Co., Waltham, Mass.*

"A Fabry-Perot Electro-Optic Modulator for Optical Maser Sources," *E. I. Gordon and J. D. Rigden, Bell Telephone Laboratories, Inc., Murray Hill, N. J.*

"Wide-Band Coherent Light Phase Modulator," *C. J. Peters, Sylvania Electronic Systems, Waltham, Mass.*

"Photon Noise," *H. A. Haus, Dept. of Electrical Engineering and Research Laboratory of Electronics, M.I.T., Cambridge, Mass.; and J. A. Mullen, Raytheon Co., Waltham, Mass.*

"An Experimental Image-Dissector Photomixer for the 2–4 Gc Region," *R. Lucy, Sylvania Electronics Systems, Waltham, Mass.*

Session 11—Computers

"Obtaining the Frequency Response of Sampled-Data Systems by Analog Computer Techniques," *G. W. Ogar, AC Spark Plug Division, General Motors Corp., Wakefield, Mass.*

"Computer Design and Control of Probabilistic Communication Networks," *R. Amara, Stanford Research Institute, Menlo Park, Calif.*

"A Practical Digital Data Transmission System for Real-Time Servicing of Branch Banks," *D. M. Bowers and D. G. Benson, Computer Control Co., Inc., Framingham, Mass.*

Tuesday Afternoon

Session 12—Antennas and Propagation II

Chairman: *J. DeBettencourt, Raytheon Co., Norwood, Mass.*

"An Electro-Mechanically Scanned Lens Antenna for Space Surveillance," *W. Rotman, E.M. Radiation Laboratory, ERD-AFCRL, Hanscom Field, Bedford, Mass.; and F. La Russa, Sylvania Electronic Systems, Waltham, Mass.*

"Millimeter Wave Antennas," *A. J. Simmons, T.R.G. Inc., East Boston, Mass.*

"ELF Signal from the Earth-Ionosphere Cavity," *C. A. Wagner and M. Balser, M.I.T., Lincoln Laboratory, Lexington, Mass.*

Session 13—Bio-Medical Electronics II

Chairman: *A. Pandiscio, Harvard University, Cambridge, Mass.*

"Problems in Bio-Medical Engineering Education," *G. Wolf, Tufts University, Boston, Mass.*

"The Growing Challenge for Engineers in Medicine," *A. Latham, A. D. Little, Inc., Cambridge, Mass.*

"Case Studies of Successful Collaboration Between Bio-Medical Scientists and Engineers," *J. B. Rewsick, Case Institute of Technology, Cleveland, Ohio.*

Session 14—Quantum Electronics II

Chairman: *N. Bloembergen, Harvard University, Cambridge, Mass.*

"Amplification of Ultrasonic Waves," *C. F. Quate, Stanford University, Stanford, Calif.*

"Dissociatively Pumped Gas Discharge Lasers," *W. R. Bennett, Jr., Sloane Physics Laboratory, Yale University, New Haven, Conn.; and W. L. Faust, R. A. McFarlane and C. K. N. Patel, Bell Telephone Laboratories, Inc., Murray Hill, N. J.*

"A Resonant Gas Maser Amplifier," *R. A. Paananen, Raytheon Co., Waltham, Mass.*

"Continuous Gas Maser Operation in the Visible," *A. D. White and J. D. Rigden, Bell Telephone Laboratories, Inc., Murray Hill, N. J.*

"Modes of a Ruby Laser," *V. Evtuhov and J. K. Neeland, Hughes Research Laboratories, Malibu, Calif.*

Session 15—Information Processing and Retrieval—A Panel Discussion

Chairman: *R. L. Blanchard, TransSonics, Inc., Burlington, Mass.*

Panel Members: *M. M. Kessler, Moderator, M.I.T., Cambridge, Mass.; L. Buckland, Inforonics, Inc., Maynard, Mass.; C. Mooers, Rockford Research Institute, Cambridge, Mass.; N. Taylor, Itek Corp., Lexington, Mass.; R. Gordon, A. D. Little, Cambridge, Mass.*

Wednesday, November 7

Session 16—Transistor Symposium

Chairman: *W. C. Dunlap, Raytheon Co., Waltham, Mass.*

"Are Transistors Approaching Their Maximum Capabilities?" *J. M. Goldey, Bell Telephone Laboratories, Inc., Murray Hill, N. J.*

"Application of Transistors in the Telephone Industry," *F. H. Blecher, Bell Telephone Laboratories, Inc., North Andover, Mass.*

"Transistors for High Power-High Frequency," *J. Hilibrand, RCA, Somerville, N. J.*

"The Transistor as a Computer Element," *R. A. Henle, IBM Corp., Poughkeepsie, N. Y.*

Session 17—General Communications and Radio Signal Technology

"On the Maximum Interference of Spherical Passive Satellites with Radio-astronomical Observations," *A. Goertz, Bureau of Standards, Boulder, Colo.*

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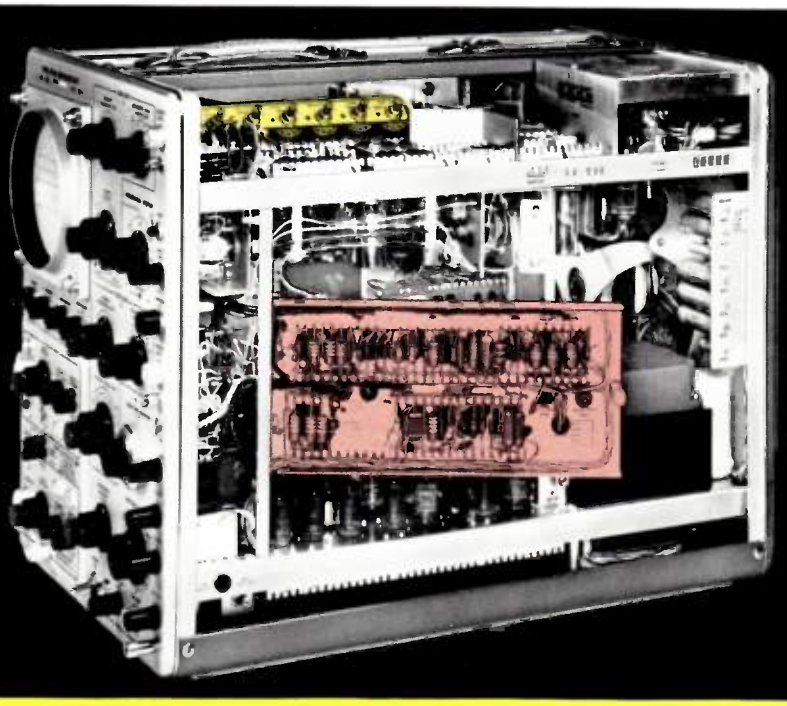
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This shows the use of A-B fixed and variable resistors in the Tektronix Type 545A high-speed DC to 30 mc laboratory oscilloscope.



A-B Type J variable resistor actual size, rated 2.25 watts at 70°C. Available in standard tapers and in standard total resistance values to 5 megohms. A-B engineers will be happy also to discuss your needs for special tapers and special, as well as, higher resistance values.



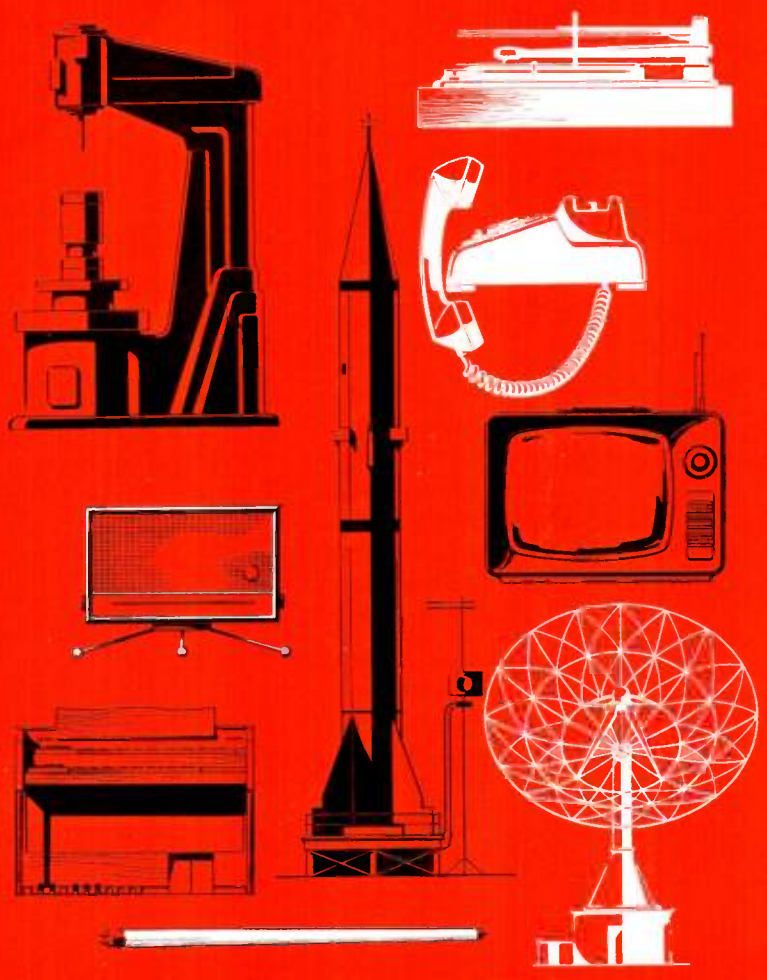
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Convergence Cores	W-01	Low residual with large gap
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TELEPHONE SYSTEMS Interstage and Matching Transformers	W-03	High permeability, low losses
H. F. FLUORESCENT LIGHTS Loading Reactors	W-07	High flux density
Transformers	W-04	High permeability, low losses, high B_{max}
ELECTRIC ORGANS AND HI-FI STEREO Oscillator Inductors	W-03	High permeability, temperature stable, linear B vs H
Output Transformers	W-04	High permeability, high B_{max}, low losses
AUTOMATIC MACHINE TOOLS Magnetic Amplifiers	R-03	Rectangular hysteresis loop, high B_{max}
Logic elements for high-power levels	R-03	Rectangular hysteresis loop, high B_{max}
Matching Transformers	W-04	High permeability, low losses, high B_{max}
MOBILE POWER SUPPLIES Static Inverters	R-03	Rectangular hysteresis loop, high B_{max}
RADAR, MISSILES Pulse Transformers	W-04 R-02 (for short pulses)	High pulse permeability, high B_{max}, low losses
PERMANENT MAGNETS Segments for D.C. Motors and Generators. Holding Magnets, Bias Magnets, Clutch Drives	M-05C M-01 M-01C	Good mechanical strength, high energy factor, high coercive force

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"Study of International Commercial Satellite Communications Systems—1965–1975," *J. S. Greenberg, S. Gubin and M. Handelsman, RCA, Princeton, N. J.*

"Instrumentation Considerations for a High-Powered Tracking System," *W. M. Humphrey, Sylvania Electronic Systems, Waltham, Mass.*

"Antenna Temperature Gradients of Atmosphere Oxygen Emission at Wavelengths Near 6 mm," *A. C. Anway, Collins Radio Co., Cedar Rapids, Iowa.*

Session 18—Information Technology I

Chairman: *D. B. Brick, Sylvania Electronic Systems, Waltham, Mass.*

"Automatic Speech Recognition," *F. C. Frick, M.I.T., Lincoln Laboratory, Lexington, Mass.*

"General Recognition Processes," *T. Marill, Bolt Beranek and Newman, Inc., Cambridge, Mass.*

"Spectral Speech Signal Representations," *H. J. Manley, Sylvania Electronic Systems, Waltham, Mass.*

"Syntactic Structure and Content Analysis," *A. G. Oettinger, The Computation Laboratory of Harvard University, Cambridge, Mass.*

Session 19—Waveguides and Transmission Lines I

Chairman: *R. Rivers, Aircom, Inc., Boston, Mass.*

"T-Septate Lunar Line," *A. Y. Hu, The Boeing Co., Renton, Wash.; and A. Ishimaru, University of Washington, Seattle, Wash.*

"Circular Electric Waveguides for Communication," *S. E. Miller, Bell Telephone Laboratories, Inc., Holmdel, N. J.*

"Failure Mechanisms in High Power Microwave Transmission Lines," *M. Gilden, Microwave Associates, Inc., Burlington, Mass.*

"Quasi-Optical Transmission Methods for Millimeter Waves," *F. Sobel, Electronic Communications, Inc., Timonium, Md.*

Session 20—Microwave Plasma Interactions I

Chairman: *G. St. John, Microwave Associates, Inc., Burlington, Mass.*

"New Microwave Applications for Plasma," *G. S. Kino, Stanford University, Stanford, Calif.*

"Electron-Beam Generated Plasmas," *L. Smullin, M.I.T., Cambridge, Mass.*

"Microwave Plasma Interactions," *A. W. Trivelpiece, University of California, Berkeley, Calif.*

Wednesday Afternoon

Session 21—Thin-Film Amplifiers

Chairman: *J. M. Lavine, Raytheon Co., Waltham, Mass.*

"Characteristics of the Insulated-Gate Thin-Film Transistor," *H. Borkan and P. K. Weimer, RCA Laboratories, Princeton, N. J.*

"Silicon Thin-Film Diodes and Transistors," *E. Rasmanis, Sylvania Electronic Systems, Waltham, Mass.*

"The Hot Electron Triode with Semiconductor-Metal Emitter," *M. M. Atalla, HP Associates, Palo Alto, Calif.*

"The Metal-Interface Amplifier," *J. M. Lavine and A. A. Iannini, Raytheon Co., Waltham, Mass.*

Session 22—Position Fixing on Earth and in Space

Chairman: *R. H. Woodward, Pickard and Burns, Inc., Waltham, Mass.*

"Omega," *J. A. Pierce, Cruft Laboratory, Harvard University, Cambridge, Mass.*

"Design Criteria for Star Trackers," *W. Markey and J. Searcy, Experimental Astronomy Laboratory, M.I.T., Cambridge, Mass.*

"Use of an Inertial System for Land Navigation," *R. Galle and J. P. Sputz, General Precision, Inc., Little Falls, N. J.*

"Optical Tracking of Satellites for Geodetic Purposes," *J. Rolff, Smithsonian Astrophysical Observatory, Cambridge, Mass.*

Session 23—Information Technology II

Chairman: *G. A. Rath, Raytheon Co., Waltham, Mass.*

"The Computer-Based Teaching Machine," *D. D. Bushnell, System Development Corp., Santa Monica, Calif.*

"Decision Theory Models for Learning," *J. A. Swets, Psychology Section, M.I.T., Cambridge, Mass.*

"Time-Shared Operation of a Small Computer," *J. B. Dennis, M.I.T., Cambridge, Mass.*

"The Theory of Sensory Perception," *H. Yilmaz, A. D. Little, Inc., Cambridge, Mass.*

Session 24—Waveguides and Transmission Lines II

Chairman: *R. Michelson, Raytheon Co., Waltham, Mass.*

"Strip Transmission Line for High Density Microwave Packaging," *W. R. Lowe, Jr., Sanders Associates, Inc., Nashua, N. H.*

"YIG Electronically Tunable Filters," *Robert Blau, Loral Electronics Corp., New York, N. Y.*

"Practical Aspects of the Use of Circular Waveguide," *R. M. Collins, Sylvania Electronic Systems, Waltham, Mass.*

Session 25—Microwave Plasma Interactions II

Chairman: *C. S. Brown, M.I.T., Cambridge, Mass.*

"Amplification by Longitudinal Waves in Beam Plasma Systems," *H. Groendijk, Philips Research Laboratories, Eindhoven, Netherlands.*

"Beam Plasma Interaction Experiments in the *X*-Band," *M. A. Allen, P. Chorney and G. E. St. John, Microwave Associates, Inc., Burlington, Mass.*

"The Self-Shielded Microwave Discharge," *C. E. Muehe, Jr., M.I.T., Lincoln Laboratory, Lexington, Mass.*

1962 Radio Fall Meeting

KING EDWARD HOTEL, TORONTO, ONTARIO, CAN., NOVEMBER 12–14, 1962

The 1962 Radio Fall Meeting, sponsored by the EIA Engineering Department and the IRE Professional Groups Committee, will be held at the King Edward Hotel, Toronto, Ontario, Can., on November 12–14, 1962. The program is as follows:

Monday, November 12

EIA Engineering Department Session

Moderator: *J. R. Whitney, Western Electric Co.*

"Legal Guides for Conducting Engineering Standardization Programs," *J. B. Olverson, General Counsel, Electronic Industries Assoc.*

"An Editor Looks at Electronics," *B. F. Osbahr, Editor, Electronic Industries.*

"EIA Established Reliability Program and Its Relation to the Armed Services and DOD," *Leon Podolsky, Chairman, EIA Parts Panel, and J. H. Schumacher, EIA Staff Enginer.*

Monday Afternoon

Reliability and Quality Control Session

Arranged by the IRE Professional Group on Reliability and Quality Control.

Moderator: *Dr. L. S. Gephart, Lockheed Missile and Space Co.*

"Reliability Assessment of Component Parts by Means of Drift Index Evluation," *Fred Decker, Minneapolis-Honeywell Regulator Co.*

"Reliability Requirements of Industrial Electronics and Mechanical Equipments," *Ray Strauss, Lockheed Missile and Space Co.*

"Reliability of Multifunction Devices—A Progress Report," *J. A. Davies and W. R. Leib, General Electric Co.*

"Reliability Program Planning Approaches," *E. A. Reeve, International Business Machines Corp.*

"A Computer Reliability Program," *S. E. Basara, Radio Corporation of America*

Tuesday, November 13

Electron Tube Session

Arranged by the IRE Professional Group on Electron Devices.

Moderator: *R. E. Moe, General Electric Co.*

"Novel Tube and Circuit Designs for Automatic Gain Controlled Video IF Amplifiers," *M. E. Ulug, Canadian General Electric Co., Ltd.*

"A New Approach to the Problems of Vertical Deflection Applications," *D. E. Steel and H. J. Fromell, Sylvania Electric Products Inc.*

"Construction and Performance of IF Frame Grid Tubes," *A. Overstrom and E. R. Campagna, Westinghouse Electric Corp.*

"Industry Standardization of Noise Figure Measurements of RF Amplifier Tubes," *T. E. Gausman, Sylvania Electric Products Inc.*

"New Applications of Multicontrol Grid Electron Tubes," *W. M. Austin and J. A. Dean, RCA.*

Tuesday Afternoon

Semiconductor Device Session

Arranged by the IRE Professional Group on Electron Devices.

Moderator: *C. R. Gray, Philco Corp., Lansdale Division.*

"A Practical Four-Stage AC-DC AM Receiver," *S. M. Humerick and H. M. Kleinman, Radio Corporation of America.*

"The Design and Application of a Germanium Field Effect Transistor," *Floyd Garrett and Dick Alford, Texas Instruments Inc.*

"Microelectronics in Consumer Products"

A panel discussion by representatives of both semiconductor manufacturers and end equipment manufacturers.

Wednesday, November 14

Radio and TV Receiver Session I

Arranged by the IRE Professional Group on Broadcast and Television Receivers.

Moderator: *B. F. Tyson, General Telephone and Electronics Labs.*

"Interim Report on Receiver Installations, New York City UHF-TV Project," *Jules Dietz, Federal Communications Commission.*

"DC Restoration—A New Approach," *S. P. Ronzheimer, Hazeltine Research, Inc.*

"Characterization of RF Transistors for AM/FM Automobile Radio Applications," *Robrecht Bosselaers and Selig Gertzis, Amperex Electronic Corp.*

"Present and Future Unsolved Radio Designers' Problems," *E. D. Hasty, The Bendix Corp.*

Wednesday Afternoon

Radio and TV Receiver Session II

Arranged by the IRE Professional Group on Broadcast and Television Receivers.

Moderator: A. Cotsworth, *Zenith Radio Corp.*

"Surveys of Adjacent Channel Interference," *W. L. Hughes, Oklahoma State University.*

"Equipment and Techniques for Simplified Testing of Color TV Receivers," *R. F. Tschannen and E. J. Duffy, Hazeltine Research, Inc.*

"UHF Translators for Extending Television Coverage," *S. P. Lapin, Adler Electronics, Inc.*

"A High Performance FM/AM Automobile Receiver," *P. S. Manson and T. H. Powell, The Bendix Corp.*

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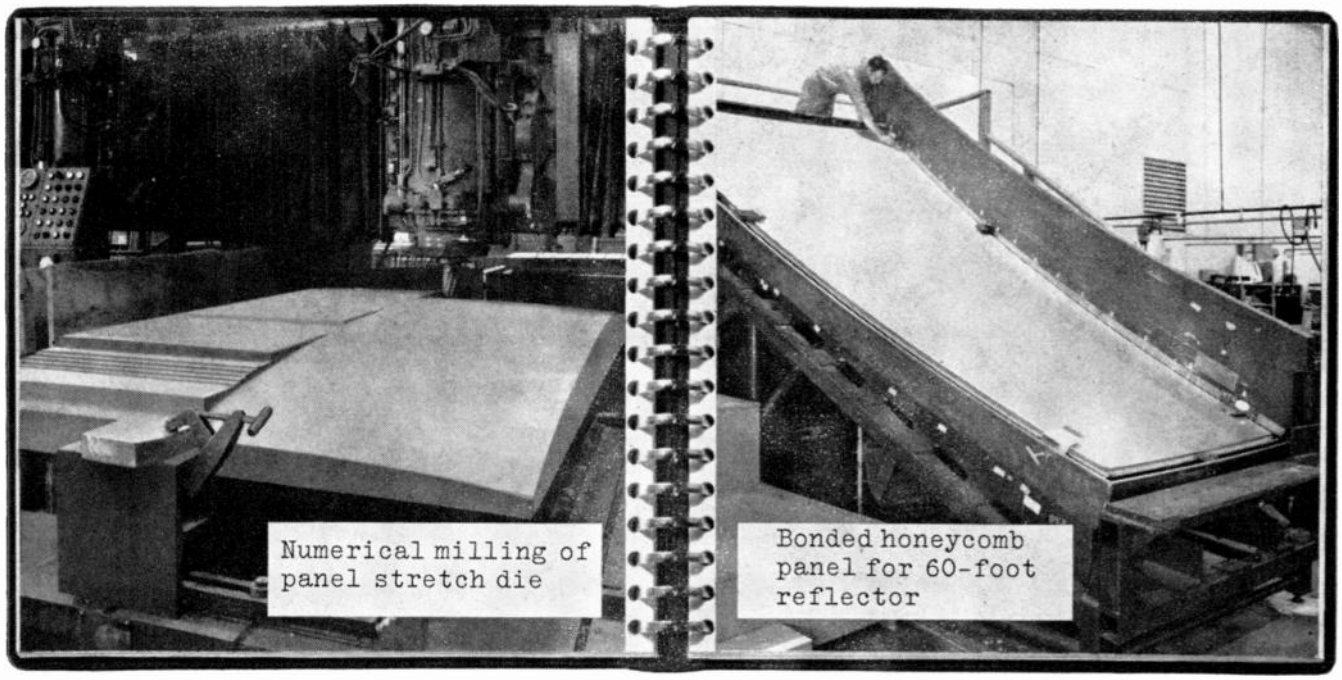
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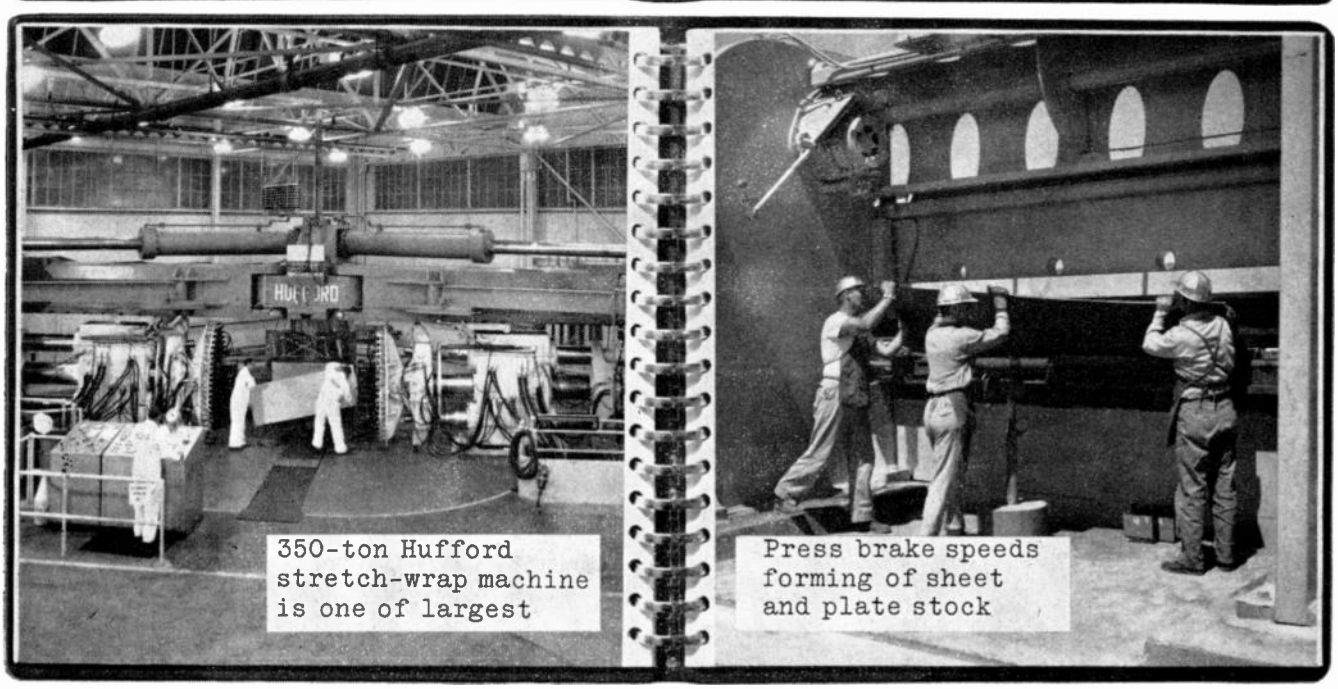
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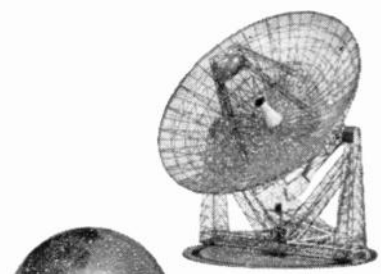
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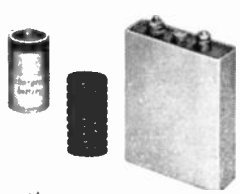
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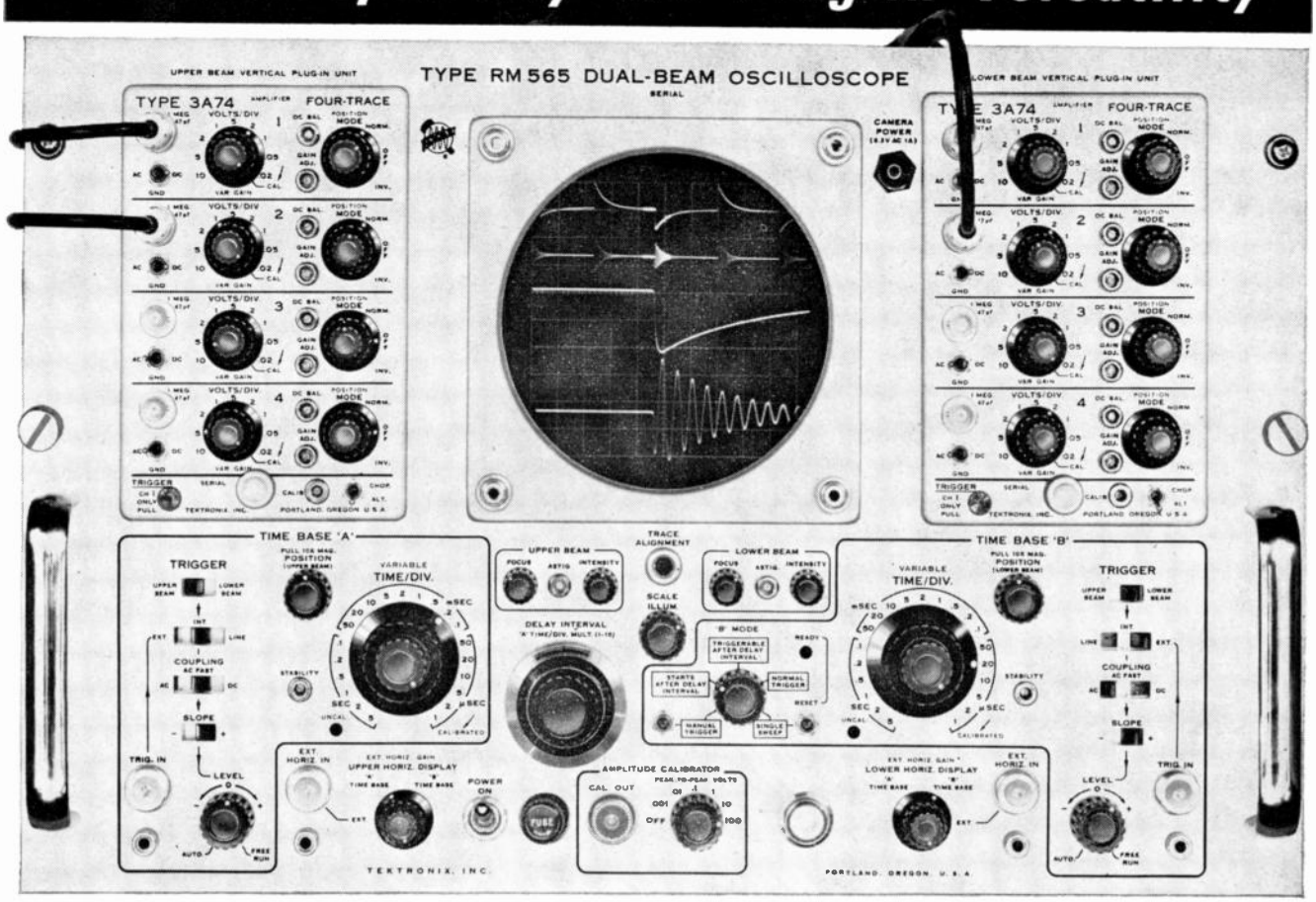
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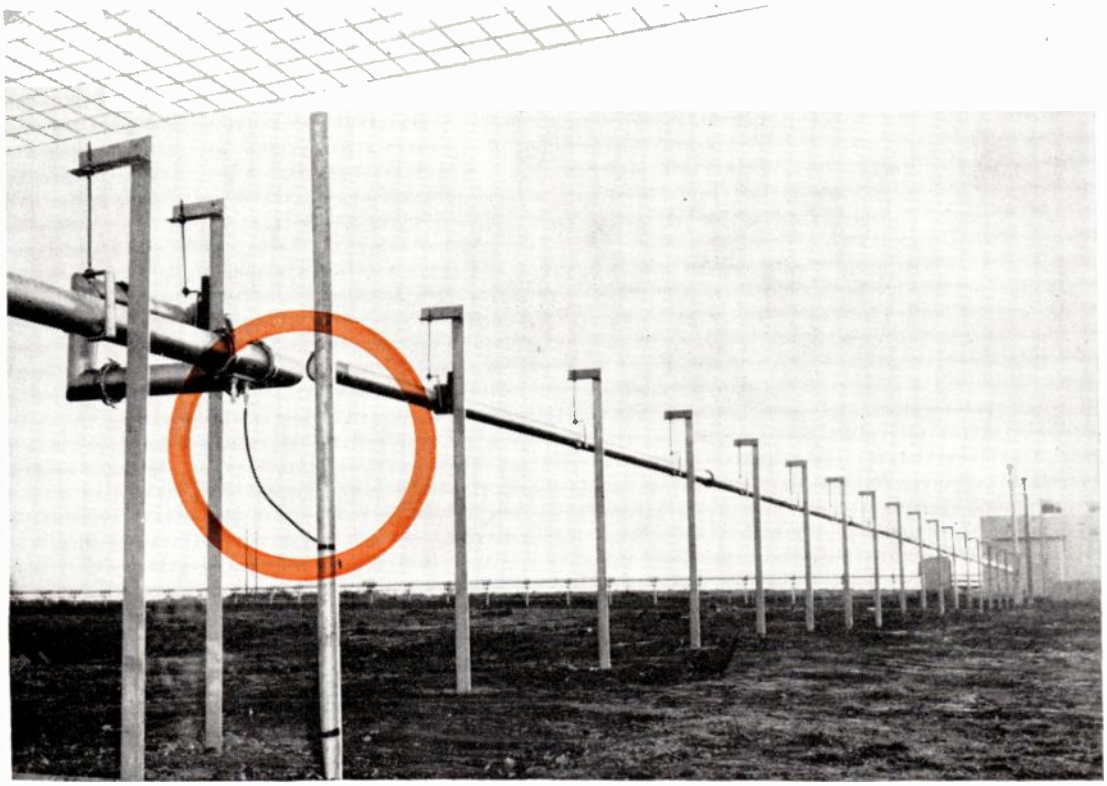
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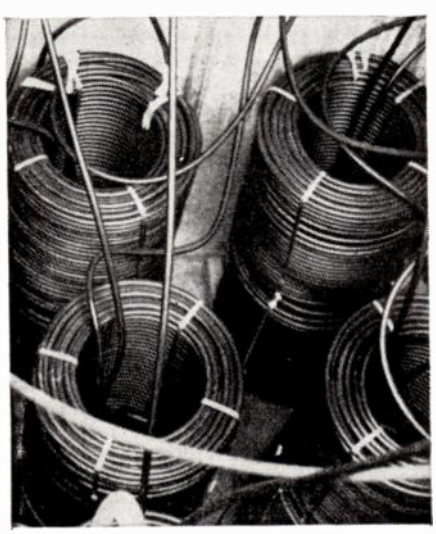


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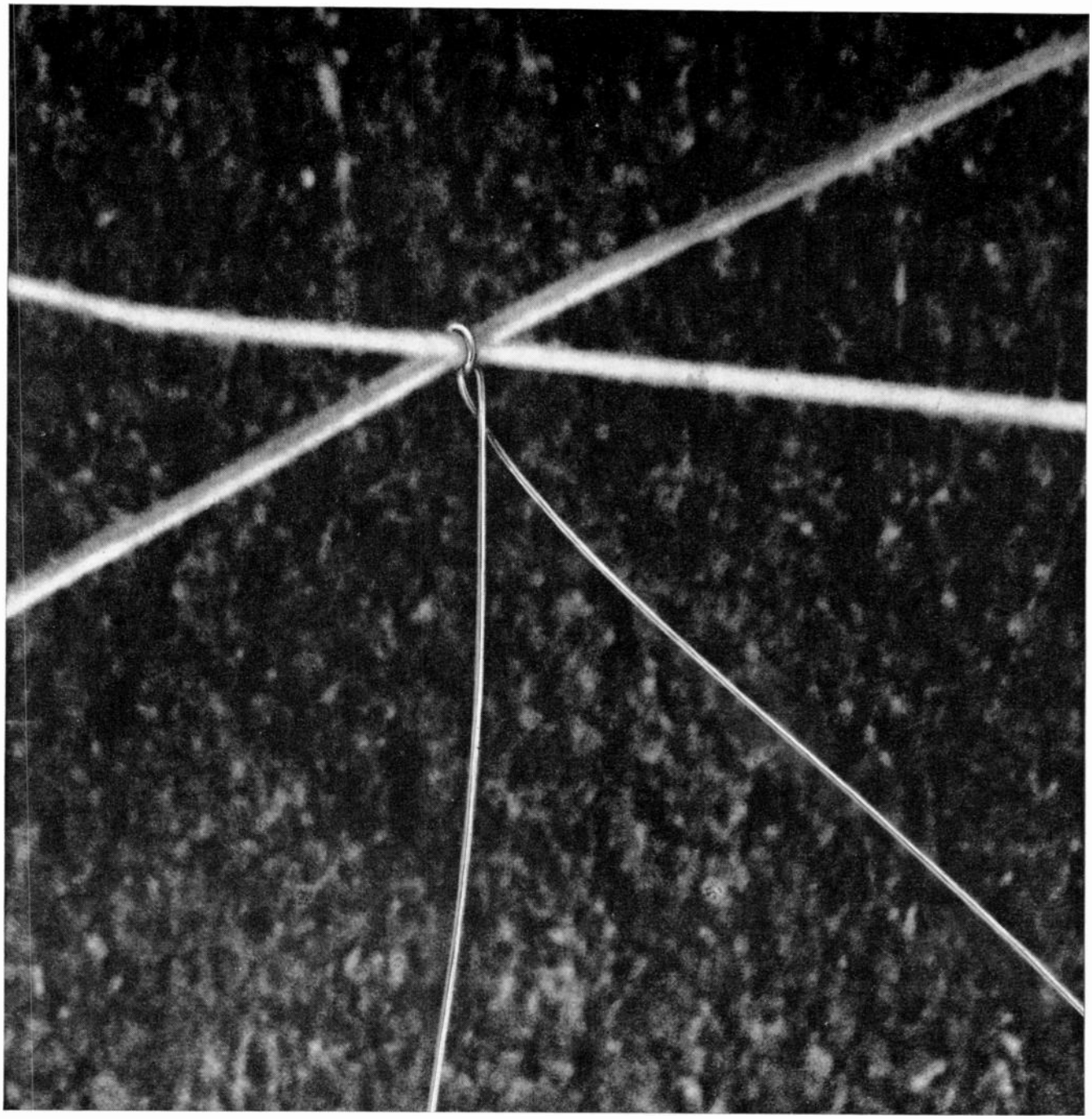
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(with nickel alloy wire, ⅓ the thickness of a human hair)

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28.4.2 Methods of Test for Transistors for Linear CW Transmission Service

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28.4.3 Definitions and Letter Symbols of Semiconductors

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28.4.9 Semiconductor Diode Definitions

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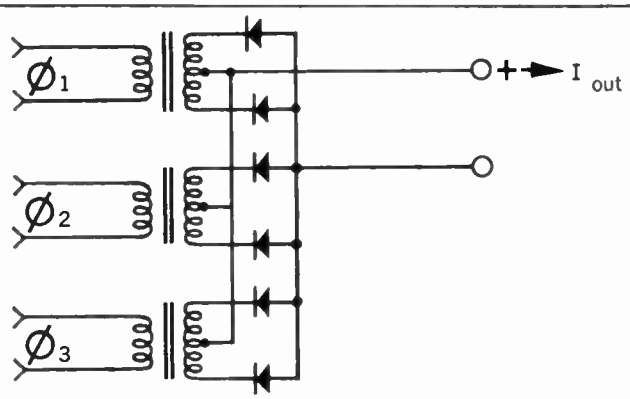
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USA 1N673	400	10	1.0	0.4	2.0	320	0.4	0.4	3.0
USA 1N947	600	10	1.0	0.4	2.0	480	0.4	0.4	3.0
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6-PHASE POWER SUPPLY OUTPUT AT 65°C.

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USA 1N947	300V	2A
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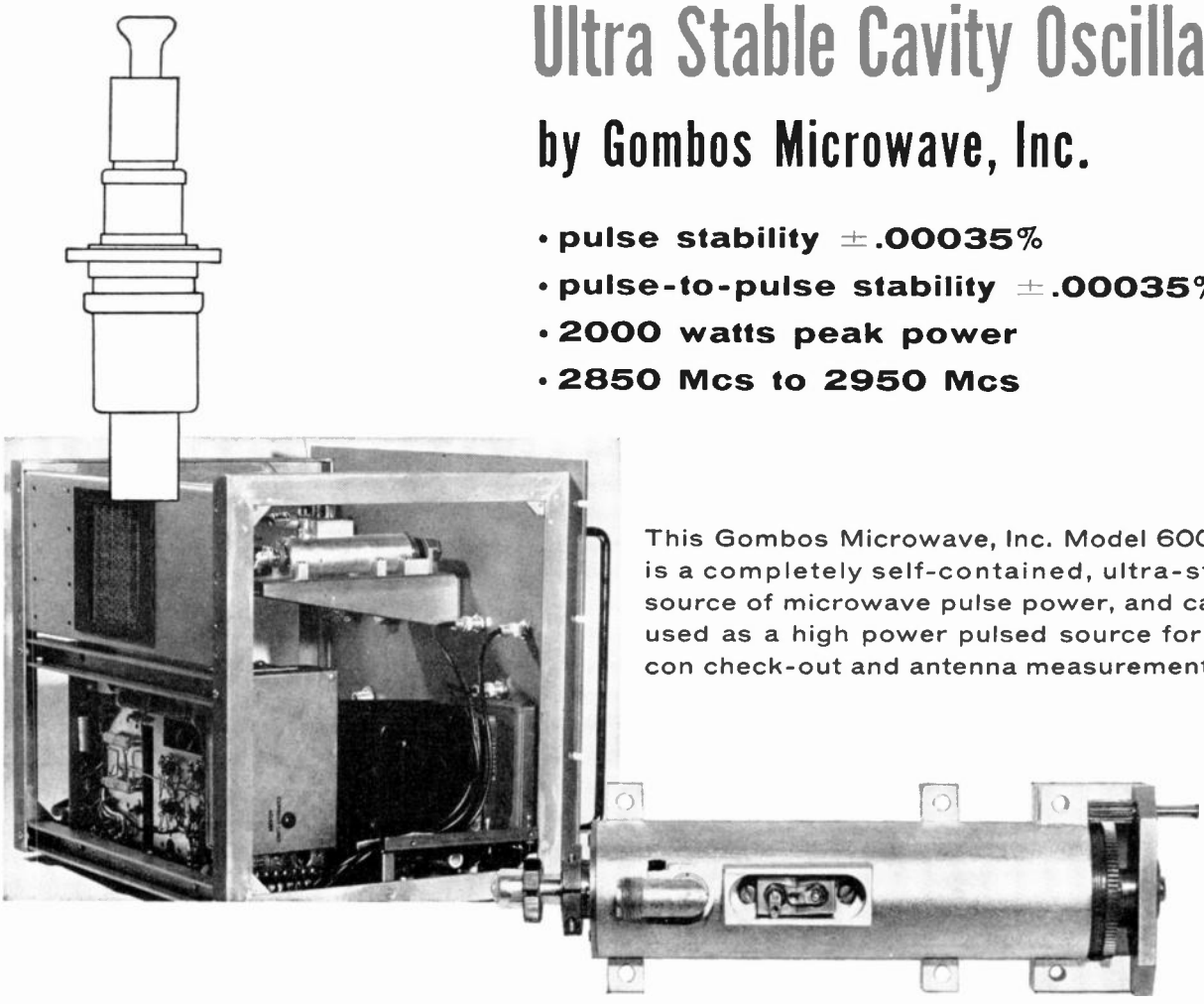
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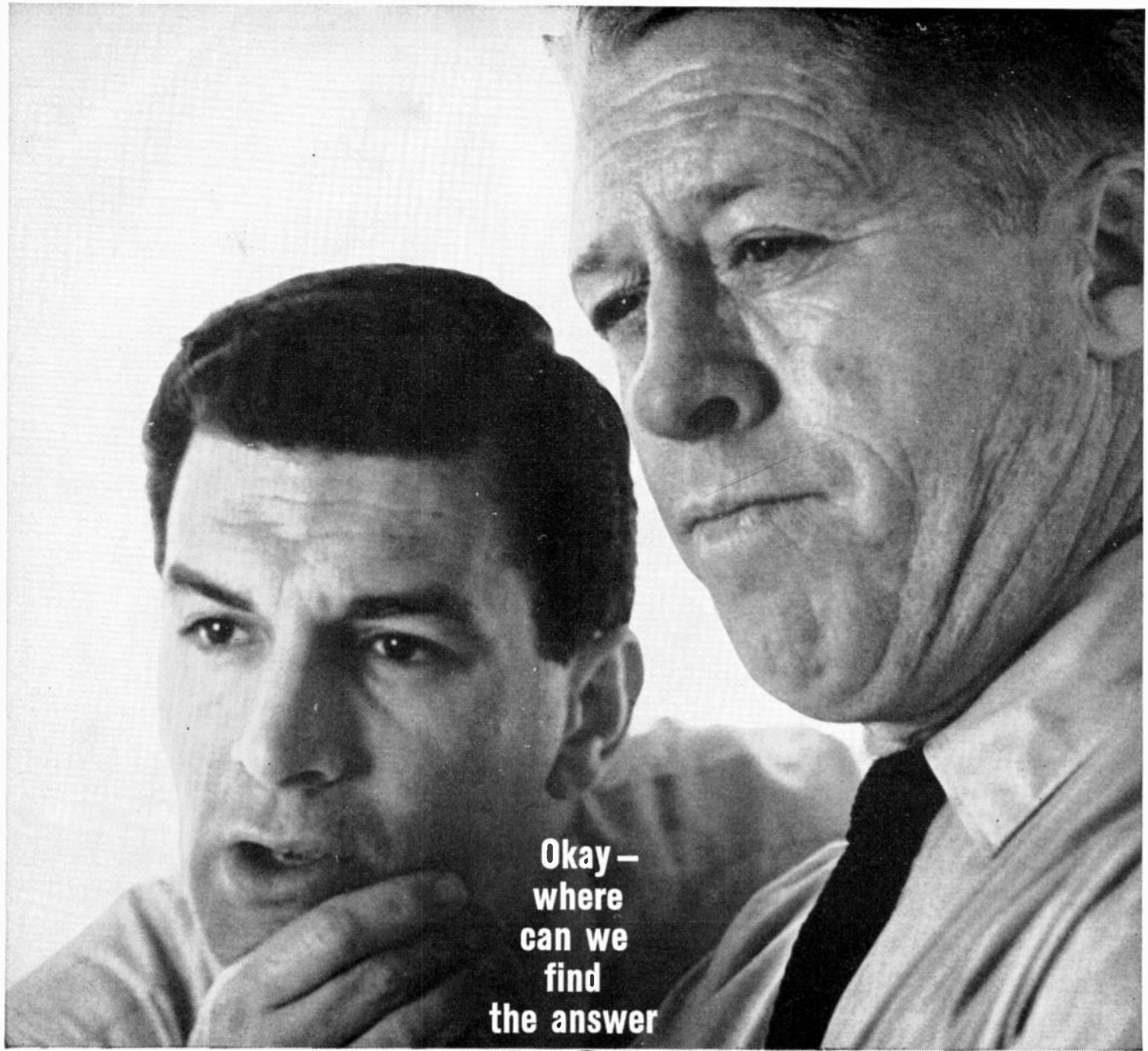
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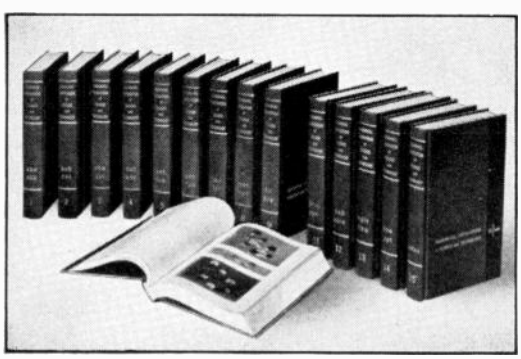
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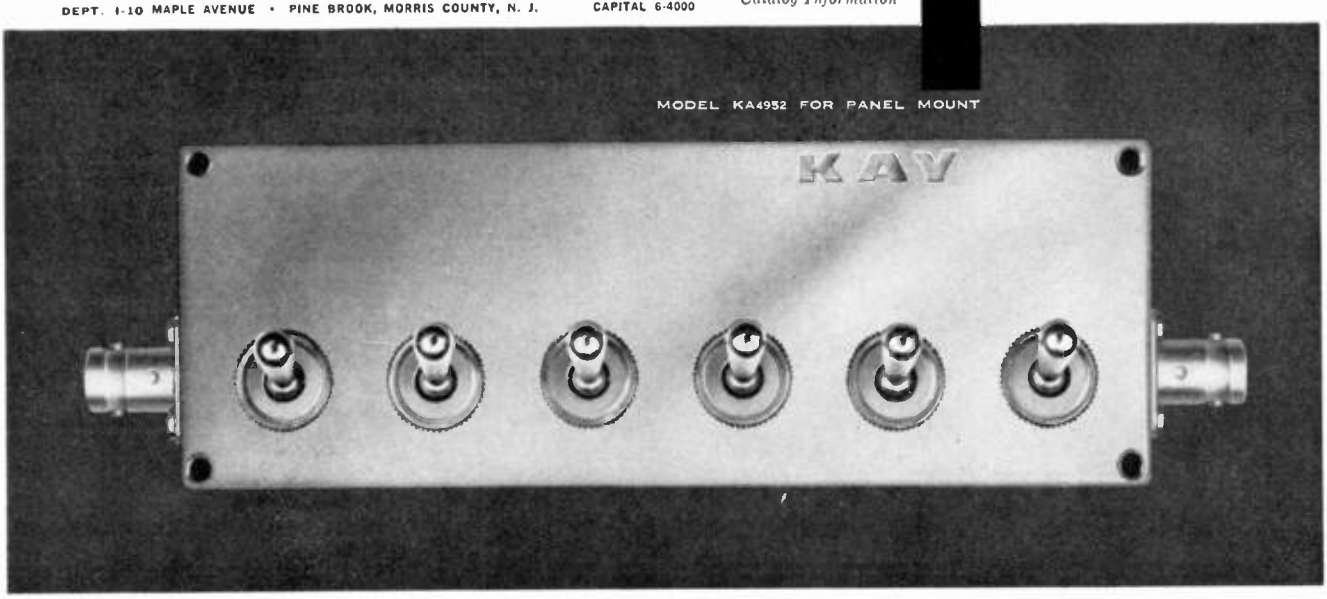
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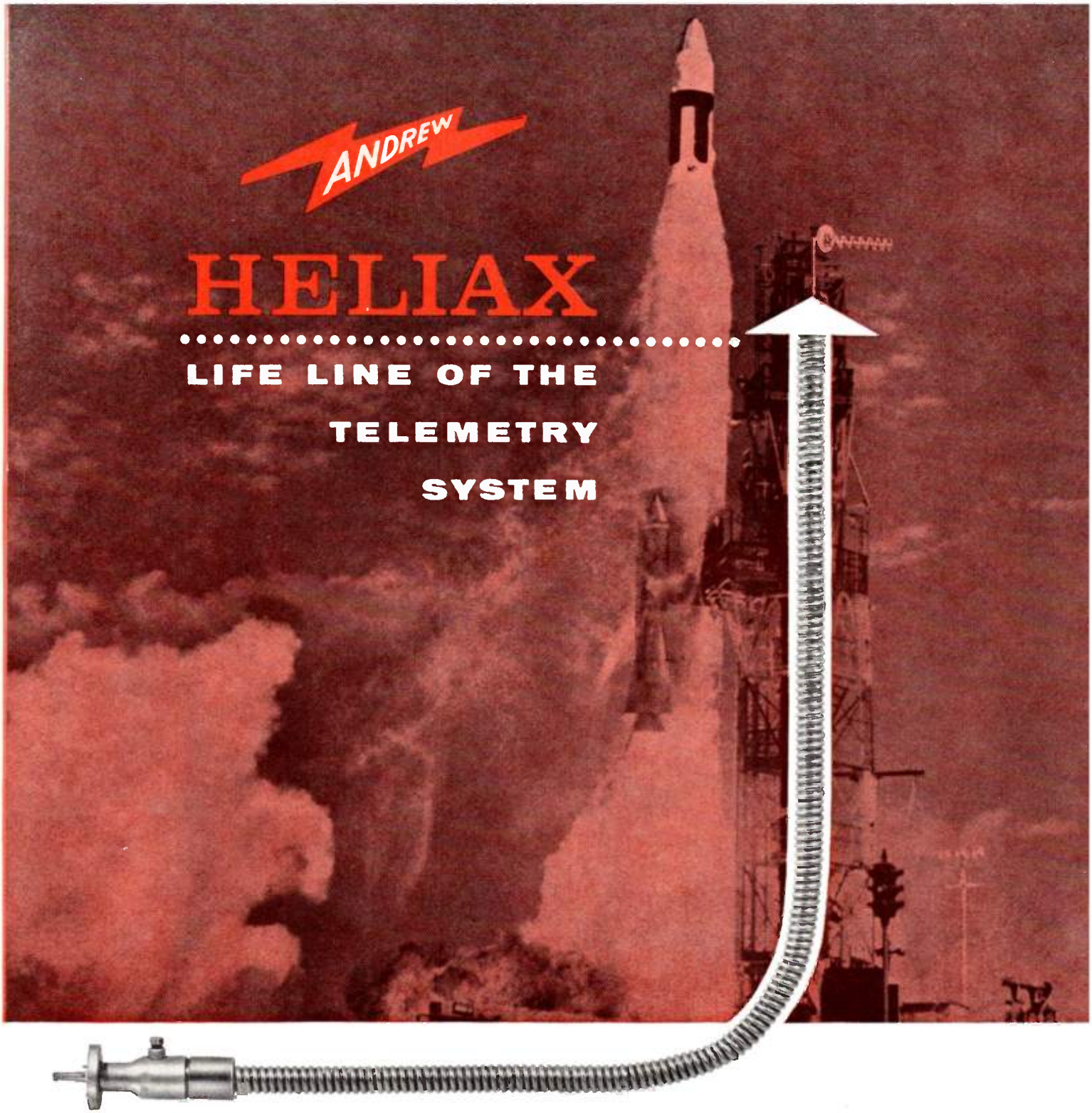
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Industrial Engineering Notes*

Association Activities

The EIA last week informed the Federal Communications Commission that a poll of all television set manufacturers revealed a preference for July, 1964 as the cutoff date for the shift from VHF to all-channel receiver production. Every TV set manufacturer responded to the EIA inquiry, and a summary of all 21 views was transmitted to the FCC. Dates recommended ranged from August, 1963 to June, 1965. Four domestic tuner producers also participated. The time needed by tuner manufacturers to increase their production capacity up to 15 times and the lead-time required by set manufacturers to change production and distribution schedules were cited by EIA spokesmen as the chief reasons why the industry needs two years to convert to all-channel production. Only about 7 per cent of the TV tuners manufactured in recent years has been of the all-channel variety, Mr. Secrest explained, so that a substantial increase in production will be required to supply the requirements of TV set manufacturers now producing at the annual rate of about 6 million units. Set manufacturers also need time to prepare for the change-over and to clear their own inventories and distribution outlets, he added.

Military and Space

The Army Materiel Command (AMC) went into operation, taking over materiel functions formerly handled by the Army's seven technical services, including research and development, procurement and production, storage and distribution, inventory management, and maintenance and disposal. At the same time, seven AMC subordinate commands, including the U. S. Army Electronics Command, became operational. The new Electronics Command, headed by Maj. Gen. Stuart S. Hoff, is headquartered at Fort Monmouth, N. J. It assumed responsibility for Army communications systems, electronic warfare systems, combat surveillance systems, automatic data processing equipment, and radar. The Electronics Command supervises Fort Monmouth, the Army Signal Supply Agency at Philadelphia, the Air Defense Engineering Agency at Fort Meade, Md., the Material Support Agency at Fort Monmouth, and related research and development laboratories and activities. The unified AMC, a result of the first major Army reorganization in nearly 10 years, will be headed by Lt. Gen. Frank S. Besson, Jr. Headquarters are located in Washington, D. C.

. . . New organizations for radio frequency management and applications engineering have been established in the Systems Research and Development Service of the Federal Aviation Agency. The Frequency Management Division will be responsible for establishing the FAA requirements in the radio frequency spectrum and obtaining radio frequency channels for FAA use from the Office of the Director, Telecommunications Management, which makes frequency assignments to Government agencies. It also will be responsible for assuring that FAA is complying with conditions of the assignments. The division was formerly under the FAA Aviation Facilities Service, which has been split into a Systems Maintenance Service and an Installation and Materiel Service. Under the recent reorganization plan, it was decided that the frequency management function would better serve FAA from within the Systems Research and Development Service and would be in line with SRD's planning responsibilities. The newly formed Applications Engineering Division will be responsible for final engineering design, development of prototypes, and specification for production of all equipment systems used by the agency for air traffic control and navigation. The organizational change is designed to speed up improvements in air traffic control and navigation systems. It will permit early system decisions based upon engineering models, followed by the final prototype designs, evaluation, and procurement.

Industry Marketing Data

June production of TV sets totaled 620,653, compared with 474,647 the month before. Cumulative production this year totaled 3,295,501, against last year's 2,801,136. Radio output for June totaled 1,721,873, compared with 1,444,074 in May. Cumulative production of radios stood at 9,264,445 this year and 7,537,290 in 1961. . . . **Japanese electronics factory output totaled $1.4 billion in 1961, a 20 per cent increase over the 1960 total of $1.2 billion, the Electronics Division of the Commerce Department's Business and Defense Services Administration reported.** During the same period, U. S. electronics output increased from $9.8 billion to $10.2 billion. BDSA said according to information from the Japanese Ministry of International Trade and Industry, Japanese television receiver production accounted for over one-third the total output in 1961 and increased by some 24 per cent above the 1960 level. Over 4.5 million television receivers were produced in 1961, more than 95 per cent of which were 14-inch types. However, a trend toward larger screen sizes was said to be evident. Production of sound recorders and reproduc-

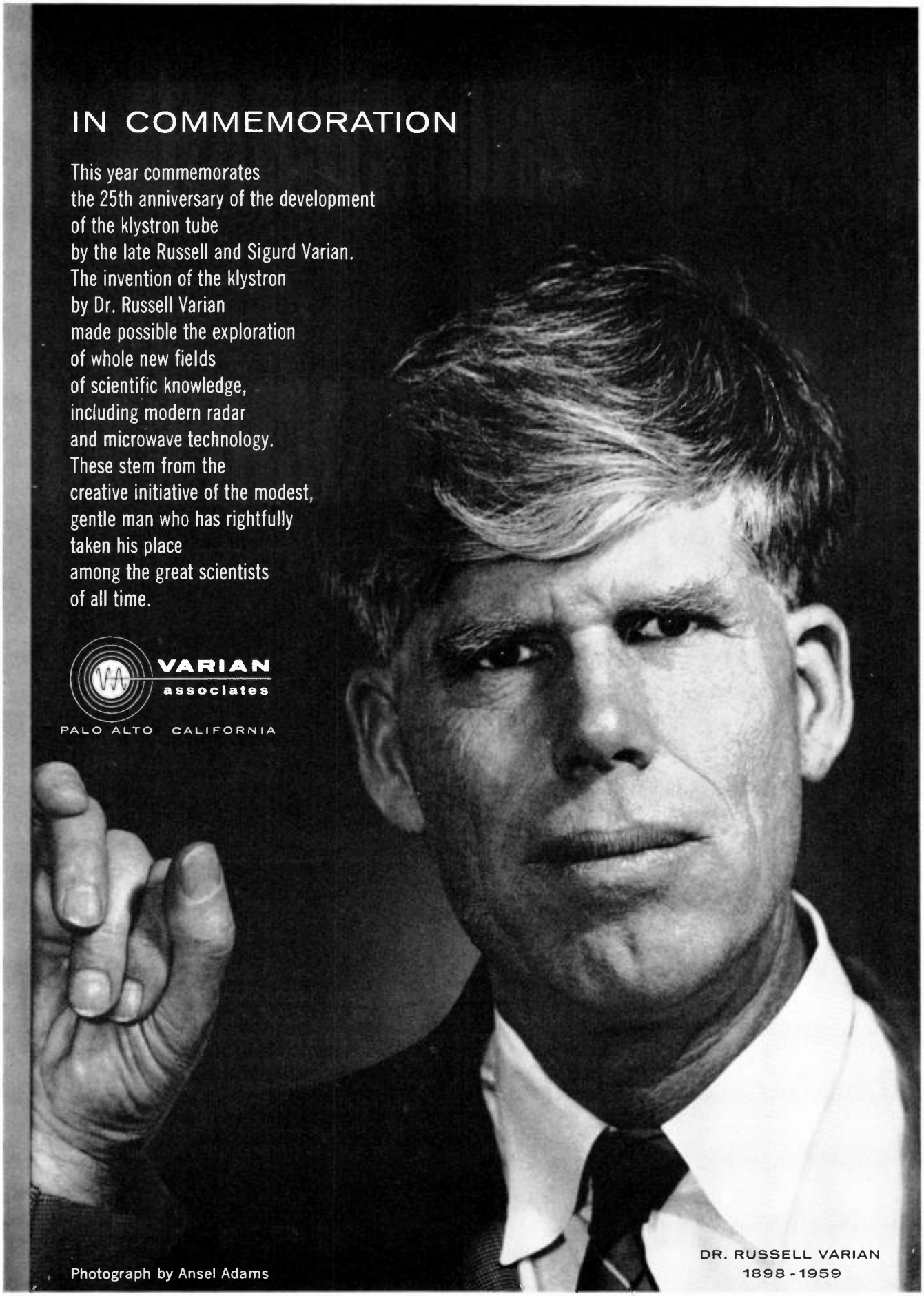
(Continued on page 60A)

* The data on which these NOTES are based were selected by permission from *Weekly Reports*, issues of August 6, 13, and 20, 1962, published by the Electronic Industries Association, whose helpfulness is gratefully acknowledged.

IN COMMEMORATION

This year commemorates
the 25th anniversary of the development
of the klystron tube
by the late Russell and Sigurd Varian.
The invention of the klystron
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made possible the exploration
of whole new fields
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These stem from the
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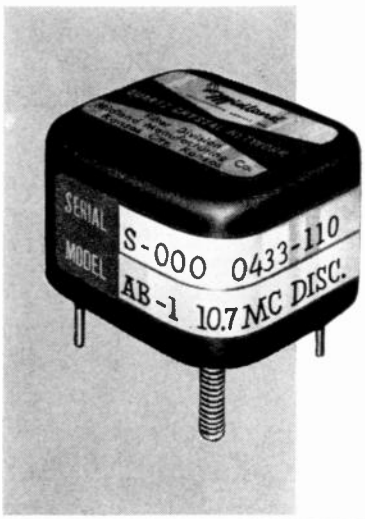


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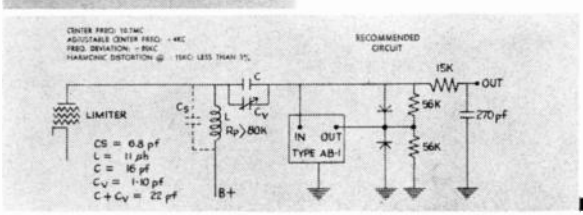
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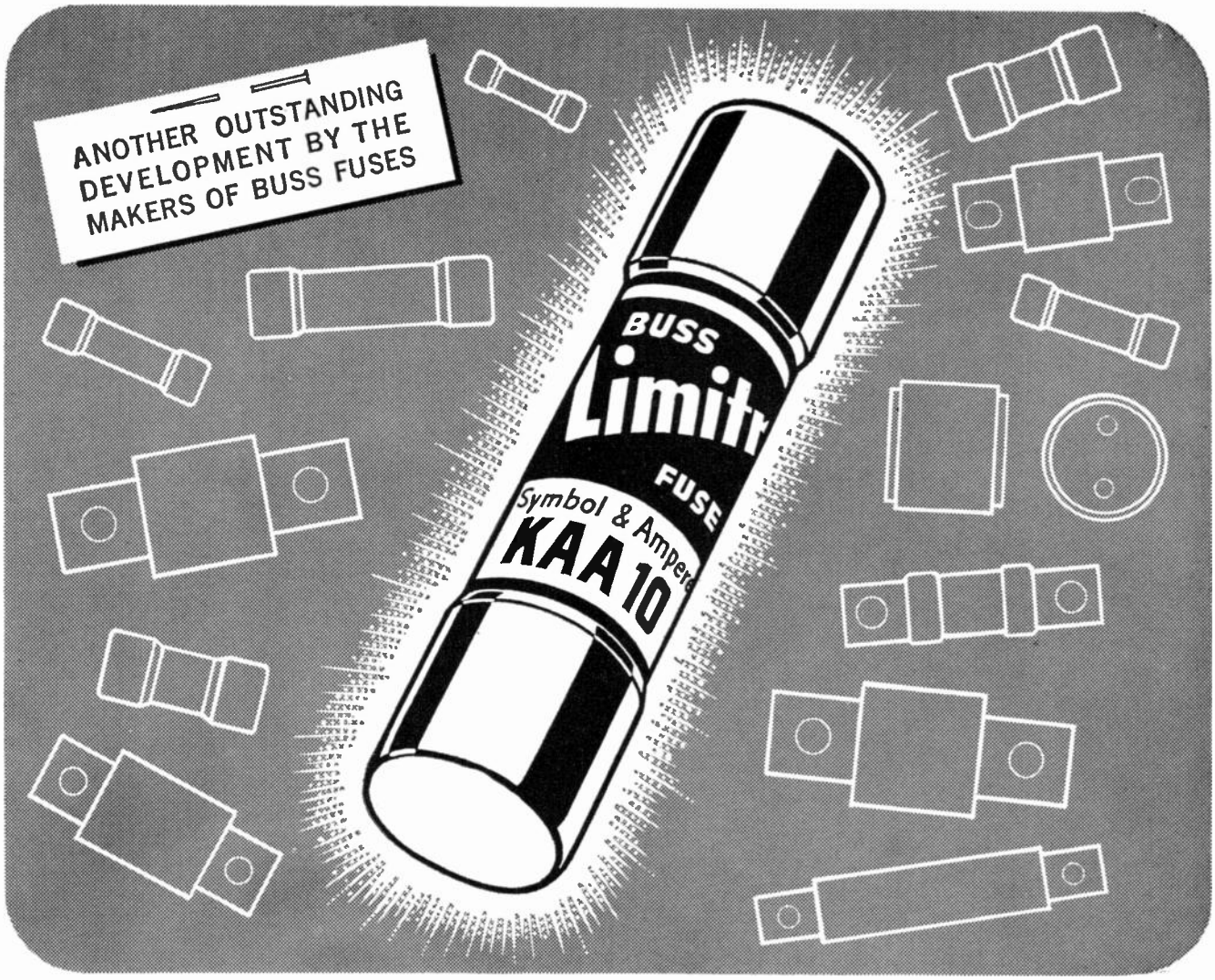


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Industrial Engineering Notes

(Continued from page 58A)

ing equipment, predominently magnetic tape recorders, registered a 60 per cent increase—from $26.3 million in 1960 to $42.3 million in 1961, BDSA reported. The output of tape recorders in 1960 totaled 447,000 units valued at $24.6 million, while the 1961 output totaled 883,000 units valued at $40.2 million. An increasing number of these recorders were transistorized. Radio-phonograph production rose to $40 million in 1961, an increase of 55 per cent over 1960. The value of hi-fi amplifier output increased by 120 per cent. Production of tube-type radios in 1961 dropped 15 per cent by volume and 26 per cent by value from the 1960 levels, while output of radios with 3 or more transistors continued to rise, reaching a total of 12.2 million units of which 10.7 million were portable, according to the report. Official statistics on "toy" radios of one or two transistors are not maintained, but Japanese trade circles estimated the 1961 output at about 6 million units. Output of commercial and industrial electronic equipment rose 35 per cent in 1961 over the previous year, but still accounted for only 14 per cent of the total electronics production. Unit production of receiving tubes increased, while reported value of production declined sharply in 1961 from the 1960 levels. According to Japanese trade sources and the Ministry of International Trade and Industry, this apparent steep drop in the average unit price was only partly the result of actual price reductions, which averaged less than 10 per cent. The major cause was attributed to a statistical overvaluation during the first nine months of 1960, occasioned by the use of a "standard price" computed several years earlier and now out of date. The "standard price" was recomputed for use in calculating the value of tube output beginning the last quarter of 1960. Other electronic components showing marked increases in unit production in 1961 were: Transistors up 29 per cent; diodes, 40 per cent; silicon rectifiers, 170 per cent; resistors, 42 per cent; capacitors, 35 per cent; and transformers, 95 per cent. Both the Japanese Government and Japanese industry circles are reported cautiously optimistic on prospects for 1962, but believe that expansion will be at a slower pace than in recent years, BDSA said. An 8 per cent increase over 1961 was seen probable. . . . **Shipments of electrical measuring instruments by U. S. manufacturers in the first quarter of 1962 were valued at approximately $35 million compared with $32.5 million in the first quarter of 1961, the Business and Defense Services Administration, U. S. Department of Commerce reported last week.** BDSA's Scientific, Photographic and Business Equipment Division said that military shipments amounted to about $10 million, while nonmilitary purchases accounted for $25 million. Shipments of laboratory standards during 1961 were not reported in the totals on a quarterly basis, but are included in the portable category in 1962. However, during the entire year of 1961, shipments of laboratory standards amounted to less than $1 million. The broad category of recording instruments continued to increase rapidly, BDSA said, indicating that the trend toward more sophisticated instrumentation persists. Reported shipments of this type doubled, from $10 million in 1960's first quarter to $20 million in the first quarter of this year. The 1961 first quarter value was $17 million. . . . **Manufacturers' shipments of home television receivers in 1961 totaled 6 million sets valued at $805 million a 6 per cent increase in quantity and a 4 per cent rise in value over 1960, according to a Census Bureau report released last week.** Census reported that shipments of radios and radio-phono combinations, excluding auto sets, totaled 11.2 million worth $301 million in 1961, compared with 10.4 million sets valued at $271 million in 1960. . . **Transistor sales at the factory during June were up by a moderate 558,118 units over the May total, but total value increased by a healthy $1,422,618, the EIA Marketing Services Department's latest monthly report disclosed.** The month-end statistics showed 21,776,037 transistors worth $26,374,356 sold in June, against 21,217,919 valued at $24,951,738 traded the month before. Cumulative totals for this year stayed far ahead of 1961 in sales, far behind in revenue accrued. Through June 123,025,915 transistors worth $150,237,229 were sold. During the same period last year, the totals were 88,682,882 units valued at $157,120,149



For Protection of Semi-Conductor Rectifiers— BUSS designed special LIMITRON Fuses

BUSS Limitron fuses are especially designed for the protection of semi-conductor rectifiers. They provide extremely fast opening on overload and fault currents, with a high degree of restriction of the let-thru current.

If each diode is protected by the proper size BUSS Limitron fuse, the fuse will open very quickly when the current drawn exceeds the rating of the diode.

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You can save time and trouble by relying on BUSS as your one source for fuses of unquestioned high quality. There is a complete line of BUSS fuses in sizes from 1/500 amperes up . . . plus a companion line of fuse clips, blocks, and holders.

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precision attenuator dissipates up to 20 watts continuously

In an active length of approximately 3¼ inches, Weinschel Engineering's new Model 870 Precision Attenuator is capable of dissipating up to 20 watts continuously without significant change in attenuation.

These high power performance characteristics result from a new patented* design principle developed by Weinschel Engineering. The design employs a resistive outer conductor, closely coupled thermally to the heat dissipating shell. This exclusive Weinschel feature overcomes the limitation in heat dissipation characteristic of resistive inner conductors and lossy dielectric materials previously used in microwave coaxial attenuators.

The Weinschel technique for building extremely stable resistors has been adapted to the dissipative coating on the outer conductor, and these coatings are artificially aged and stabilized to give maximum stability under peak pulse power and extreme temperature and humidity cycling.

The Model 870 covers the frequency range from 7 to 12.4 KMC. A Certificate of Calibration showing insertion loss and VSWR test data at 7.0, 9.0 and 11.0 KMC is supplied with each attenuator. We invite you to write for complete specifications and prices.

* U.S. Patent #3,002,166

For your less precise applications we recommend BlueLine General Utility Components by Weinschel.

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IRE People



Dr. Leslie S. Radnay (A'60) has been appointed Project Engineer and Consultant for High Frequency Communications of Manson Laboratories, Inc., Stanford, Conn.

He has been engaged for many years in the design, development and supervision of communication devices both in his native Hungary as well as in this hemisphere. He has been responsible for the development of such equipment as SSB Transmitters, Variable Frequency Oscillators, Linear Amplifiers and Low-Power Wide-Band Transformers. In addition, he has published extensively on oscillators, antenna systems and is the author of "Design of Small Radio Transmitters." Prior to joining Manson, he was employed by Reeves Instrument Company as Chief Engineer. He also has been associated with the Westrex Corporation and R.C.A. of Canada.

Dr. Radnay received the E.E. degree from the University of Budapest in 1934 and the Ph.D. degree in natural sciences from the University of Szeged in 1935. He is a member of the Armed Forces Communications and Electronics Association and the ARRL.

❖



Richard W. Rhyins (M'54) has joined the staff of Measurement Systems, Inc., Norwalk, Conn., as Manager of Advanced Development Engineering. His first assignment will involve a new optical missile tracking system accurate to 5 seconds of arc, being developed for the National Aeronautics and Space Administration.

He was formerly with the Barnes Engineering Company, Stamford, Conn., where he was responsible for the basic design of a number of infrared instruments and systems, including cameras, scanners, tracking spectrometers and a sun tracker employed in the Thor missile. Before joining Barnes in 1958, he was a Senior Engineer with Electro-Mechanical Research, Inc., Ridgefield, Conn.

Mr. Rhyins received the B.S. degree in physics from Rennselaer Polytechnic Institute in 1950. He is a member of the Optical Society of America, and the American Association of Physics Teachers.

❖



John W. Scheck (M'60) formerly Marketing Manager of the Berkeley Division of Beckman Instruments, Inc., Richmond, Calif. has been named Marketing Manager for Electronic Instruments at Allen B. DuMont Laboratories, Divisions of Fairchild Camera and Instrument Corporation. In addition to his background in general marketing administration, he brings to Du Mont specific experience in planned product diversification, other company and product acquisitions, as well as continuous broadening of existing product lines.

Before his association with Beckman Instruments, he was with a manufacturer's representative firm in the Midwest, where he represented Tektronix, Incorporated, Ballantine Laboratories, Beckman Instrument, Kepco Labs, Wave Line and others.

He is an electrical engineering graduate of Illinois Institute of Technology and has taken additional courses in sales management and management development at Northwestern and Santa Clara University. He is a member of the Board of Directors of the Western Electronic Manufacturer's Association.

❖



Dr. Leonard S. Sheingold (S'47–A'51–M'55–SM'55) Vice President of Research for Sylvania Electronic Systems, a division of Sylvania Electric Products Inc., received an Exceptional Civilian Service Citation from General Curtis E. LeMay, Air Force Chief of Staff. Dr. Sheingold has returned to Sylvania following a 12-month leave to serve as Chief Scientist to the U. S. Air Force. The citation, signed by Secretary of the Air Force Eugene M. Zuckert, cited his "outstanding ability as a scientist, his resolute efforts to improve the scientific and technical effectiveness of the Air Force, and his complete devotion to the nation's welfare. . . ." Prior to his service as Chief Scientist, he was Director of Sylvania's Applied Research Laboratory in Waltham, Mass.

❖

(Continued on page 64A)

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SILICON CHOPPERS



Sprague offers this Wide Selection of Silicon Chopper Transistors in Volume Production now!	
2N2162 2N2165	Guaranteed 30 volt rating. Typical f_T of 20 Mc and low offset voltage make these transistors ideal where high voltage is required
2N2163 2N2166	Have 15 volt rating and same high frequency performance and low offset voltage as 2N2162
2N2164 2N2167	Highest frequency P-N-P Silicon Choppers available as standard types
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2N2274	Similar to 2N2185 but has lower inverted dynamic saturation resistance
2N2276	Low-cost version of 2N2274
2N2278	Very low offset voltage of 1.75 mV at I_B = 1 mA
2N2187	Matched pair of 2N2185 with ΔV_{OFF} = 50μV max. from +25 C to +85 C
2N2275	Matched pair of 2N2274 with ΔV_{OFF} = 100μV max. from +25 C to +65 C
2N2277	Matched pair of 2N2276 with ΔV_{OFF} = 100μV max. from +25 C to +65 C
2N2279	Matched pair of 2N2278 with ΔV_{OFF} = 50μV max. from +25 C to +85 C





Specifically designed for use as low-level choppers, Sprague Silicon Precision Alloy Transistors are inherently stable. Every Sprague chopper transistor undergoes a rigid production conditioning of 40 temperature cycles from −55 C to +140 C, a 200-hour bake at +140 C, and a 2-hour 125 mW operational burn-in!



For additional information, write Technical Literature Service, Sprague Electric Co., 235 Marshall St., North Adams, Mass., indicating the types in which you are interested.

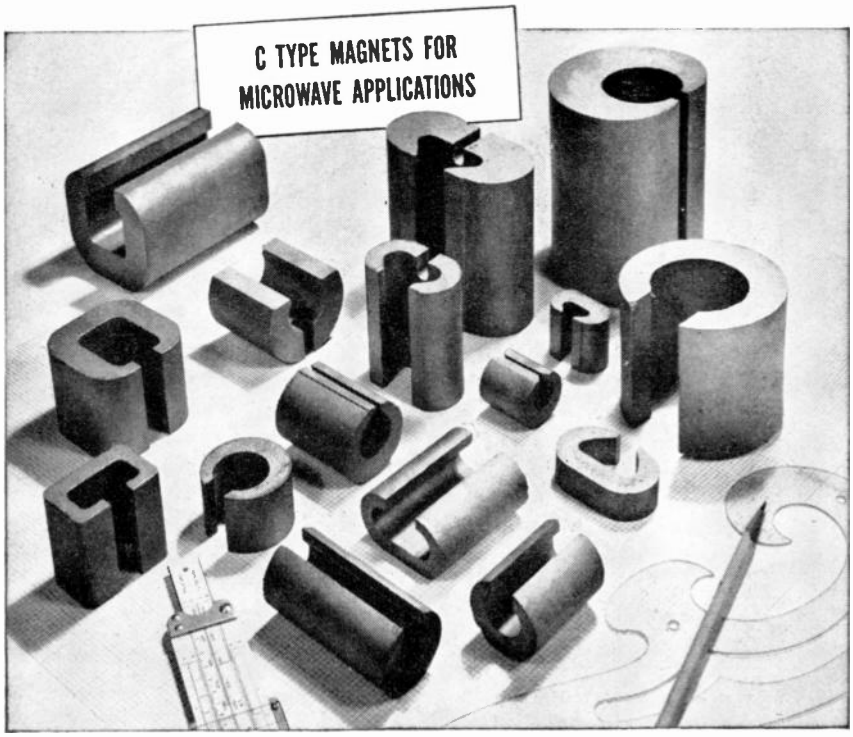
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Arnold C-type Alnico Magnets are available in a wide selection of gap densities ranging from 1,000 to over 7,500 gausses. There are six different basic configurations with a wide range of stock sizes in each group.

The over-all size and gap density requirements of many prototype designs can be met with stock sizes of Arnold C Magnets on an immediate-shipment basis, or readily supplied in production quantities.

When used in transverse field isolators, Arnold C Magnets supply the magnetizing field to bias the ferrite into the region of resonance, thus preventing interaction between microwave networks and isolating the receiver from the transmitter. These magnets are also used in differential phase shifters and duplexers, and Arnold is prepared to design and supply tubular magnets to provide axial fields in circular wave guides.

A feature of all Arnold C Magnets is the excellent field uniformity along the length of the magnet. Versatility in design may be realized by using multiple lengths of the same size magnet stacked to accomplish the needs of your magnetic structure.

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Telephone: ENterprise 6374

IRE People

(Continued from page 62A)



Bruno O. Weinschel (A'45–M'47–SM'53) has been appointed to the Advisory Council of George Washington University's Center for Measurement Science. The primary objective of the center is to improve, expand, and disseminate knowledge in the field of measurement science. He is President of Weinschel Engineering, Kensington, Md., developers and producers of precision microwave instruments and components. Prominent in the precise microwave measurements field, he is a member of Commission I of URSI and serves on the Public Advisory Committee on Calibration, USAF, as well as on the Calibration and Measurements Advisory Committee, NBS.

❖



Bruce B. Barrow (S'52–M'56–SM'59) has been appointed Senior Engineering Specialist at the Applied Research Laboratory of Sylvania Electronic Systems, a division of Sylvania Electric Products Inc.

He will be associated with the Communications Research Department, which has been carrying out investigations into advanced communications techniques. He will initially be concerned with studies leading towards improved tropospheric-scatter systems.

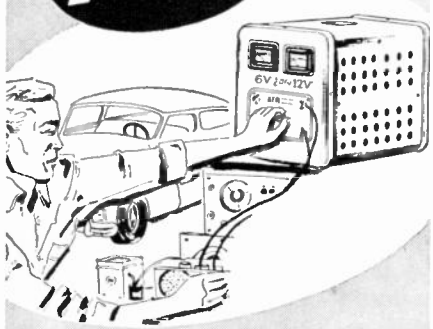
Prior to joining Sylvania, he was with the SHAPE Air Defence Technical Centre in the Hague, Netherlands, where he worked for the past several years in the Communications Group. He recently received the degree of Doctor in Technical Science, from the Technische Hogeschool at Delft, Netherlands, for his research on data transmission over fading radio paths. Before going to the Netherlands, he worked for Hermes Electronics, Inc., Cambridge, Mass. and Paris, France.

Dr. Barrow received the B.S. and M.S. degrees in 1950 from Carnegie Institute of Technology, and the advanced degree of Electrical Engineering from Massachusetts Institute of Technology, where he studied on a Tau Beta Pi Fellowship, in 1956. He spent a year (1952–1953) at Delft on a Fulbright Scholarship. He was one of the founders of the IRE Benelex Section, was its first Secretary-Treasurer, and was one of the organizers of the new IRE Region 9. He is also a member of Sigma Xi, Tau Beta Pi, and the Nederlands Radiogenootschap.

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(Continued on page 92A)

ATR



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For Demonstrating and Testing Auto Radios

TRANSISTOR OR VIBRATOR OPERATED
6 Volt or 12 Volt!

New Models . . . Designed for testing D.C. Electrical Apparatus on Regular A.C. Lines. Equipped with Full-Wave Dry Disc Type — Rectifier, Assuring Noiseless, Interference-Free Operation and Extreme Long Life and Reliability.

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		-or-				
		12	6	12		
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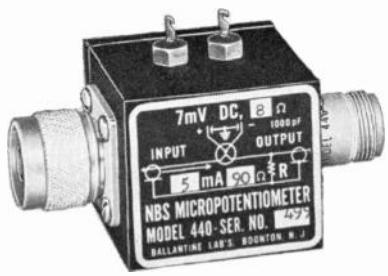
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For the calibration of ac voltmeters, signal generator outputs, or the frequency response of 'scopes

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Price: $175 per resistor, plus $75 per thermocouple housing assembly.

Acts as a low impedance source of accurately known voltage at frequencies from 0 to 900 Mc. It consists essentially of a UHF thermocouple whose heater is in series with a special radial resistor. When connected to an external signal source, the voltage drop across the resistor can be held to a known value over a range of frequencies from 0 to 900 Mc by monitoring the dc output of the thermocouple. Each thermocouple-resistor combination can be operated over a voltage range of 4 to 1 selected between the limits of 15 microvolts and 1 volt. The device is ideal for calibration of ac voltmeters, 'scopes, or signal generators. There is probably no device better adapted to these tasks, principally because it is so simple to use over such a wide range of frequencies. It is based on designs by Mr. Myron C. Selby of the National Bureau of Standards.

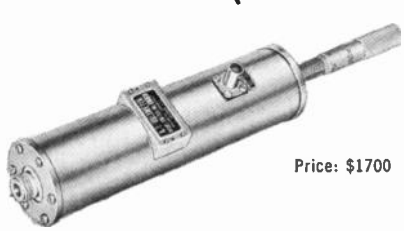
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Price: $950 with 5 probes, 1 to 50 volts

A device for the precision measurement of ac voltages of 1 to 50 volts at frequencies from 25 cps to 30 Mc. Measurements are made by equating an unknown ac voltage to an accurately measurable dc voltage using a resistor-thermocouple probe. Accuracy is better than 0.1% up to 10 Mc, and better than 0.5% to 30 Mc even without application of calibration data. It is based on designs by Mr. F. L. Hermach of N. B. S.

MODEL 390 A-T (ATTENUATOR-THERMOELEMENT) VOLTMETER



Price: $1700

A laboratory standard device for measurement of voltages from 1 volt to 300 volts (depending on frequency) from 10 Mc to 1000 Mc. NBS calibration is available to 400 Mc at the present time, but facilities are planned to 1000 Mc. The Model 390 is based on a design by Messrs. Selby and Behrent of N. B. S.

1932 30 years 1962

Ask for 4-page brochure on any of these instruments.

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Boonton, New Jersey

CHECK WITH BALLANTINE FIRST FOR LABORATORY AC VACUUM TUBE VOLTMETERS, REGARDLESS OF YOUR REQUIREMENTS FOR AMPLITUDE, FREQUENCY, OR WAVEFORM. WE HAVE A LARGE LINE, WITH ADDITIONS EACH YEAR. ALSO AC/DC AND DC/AC INVERTERS, CALIBRATORS, CALIBRATED WIDE BAND AF AMPLIFIER, DIRECT-READING CAPACITANCE METER, OTHER ACCESSORIES.

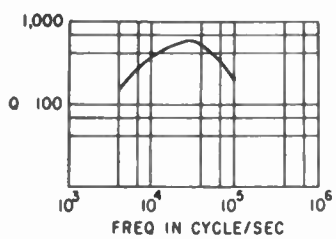
Ferrite improvements extend core use, save assembly costs

Significant advances in ferrite materials by Ferroxcube Corporation of America make possible cost and size reductions in inductive components. Such components include transformers, filters, delay lines, loading coils, modulation reactors, chokes, accelerators, pulse networks and transformers, amplifiers, oscillators, r-f tuners and many others operating at frequencies from a few hundred cycles to several hundred Mc/sec.

In the past few years, temperature stability of Ferroxcube ferrites have been radically improved while permeabilities have been increased from 500 to as high as 5000. Ferrites offer material resistivity from 10^6 to 10^{11} higher than other metal cores, reducing eddy current losses even at frequencies approaching microwave regions.

The lower losses of ferrite also give rise to higher Q components. In recording heads, and especially in erase heads, less driving power is required. Ferrites are also extremely well suited to low-power transistor applications.

Special consideration should be given to the wide frequency range over which permeability of the ferrite remains constant. Reasonably high Q values at relatively low frequencies are possible when ferrite cores are employed, (see Figure).



Typical Q curve for a particular pot curve configuration.

When selecting material for inductance or magnetic applications requiring magnetic cores, check ferrite advantages in: core costs, efficiency, ease of installation, required accuracy in the duplication of parts, size of core, possible core configurations, power requirements, frequency of operation, power losses, environmental effects, mechanical requirements, cost of assembly, overall initial and long range cost of the component.

Complete details on the various core shapes and materials are available from Ferroxcube upon request.

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These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your IRE affiliation.

Servo-Meter-Potentiometer



A new approach to the miniaturization of servo-meter-potentiometer combinations, resulting in reduced weight, size and power without sacrificing system accuracy or speed of response, has been developed by **Computer Instruments Corp.**, 92 Madison Ave., Hempstead, N. Y. The new principle makes possible accuracies to 0.25% and response speed to 10–50 ms, without trains, conventional motors and ganged pots.

Weighing 6 oz., the servo-meter-pot combinations measure 1½" deep, including feedback and follow-up pots, and require 0.5 watts maximum. Because of their performance, reduced size and long life, they are suitable for such applications as servo-multiplication, function generation, analog storage and position indication.

Improvements in both physical and performance characteristics are made possible by a newly designed galvanometer drive mechanism—a meter movement utilizing the firm's carbon-film elements. The virtually infinite resolution of these elements is responsible for the high system accuracy. Fast speed of response is aided by the low inertia of the galvanometer coil and low torque wipers, which result in a higher torque-to-inertia ratio than is normally possible with conventional galvanometer design.

Pressure Transducers

Capabilities of the new Type 4-313 pressure transducers are described in a two-page bulletin available now from **Consolidated Electrodynamics Corp.**, 360 Sierra Madre Villa, Pasadena, Calif., a subsidiary of Bell & Howell.

Absolute, gage, and unidirectional differential models of the 4-313 can measure pressures from 0–15 to 0–5000 psi. Bidirectional differential models measure from ±5 psi through ±50 psi. Superior performance is maintained under vibration, shock, and in an operable temperature range of −320°F to +300°F.

Copies of Bulletin 4-313 can be obtained from CEC.

Solid State Supply



A 100% solid-state transistorized power supply designed for 0.01% regulation, and supplying power from an extremely low impedance source, is announced by **Power Devices, Inc.**, 8710 Darby Ave., Northridge, Calif. Range is DC to mc, reducing external decoupling requirements. This compact, high efficiency component offers superior stability with less than 10 microsecond response. Several series are available and can be supplied for use in computer systems, military ground support equipment and test instrumentation. Output voltage and current are continuously variable and individually metered. Input is 105–125 volts RMS, 55–65 CPS. Prices for 8 different models range from $395 to $625. All are available from stock.

Sub-Miniature Terminations

The series TG Terminations by **Microlab**, 570 W. Mt. Pleasant Ave., Livingston, N. J., offer the design engineer a new flexibility in microwave systems where size and weight are critical. Aluminum housing plus standardization of miniature parts provide a substantial size—weight reduction together with a new low cost.



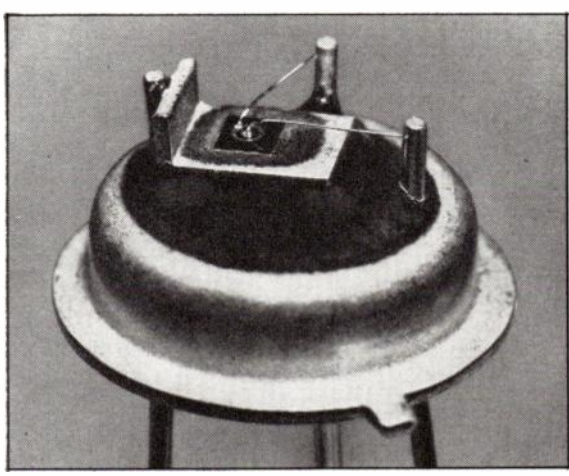
The TG Terminations are provided with connectors to mate with most of the sub-miniature coaxial connectors now available. Normally supplied with a male-female combination, they may also be supplied with other combinations of the same or different series. Special connectors may be adapted to the Series TG line upon request.

Typical frequency range is DC to 2500 MC with VSWR of 1.3:1 depending on the connectors. Power rating is 1 watt with a peak rating of 1 Kw. Temperature coefficient is 0.00008db/db/°C and temperature range is −55°C to +125°C. Finish is silver plate per QQ-S-365 with clear Irridite. Meets MIL-E-5252 environmental test.

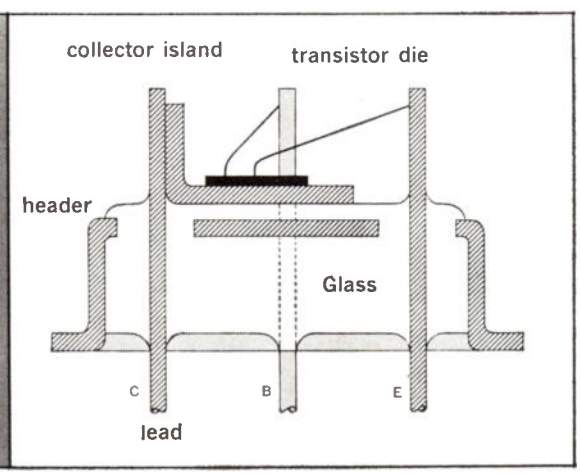
(Continued on page 70A)

From Fairchild

PLANAR* TRANSISTORS ELECTRICALLY ISOLATED FROM PACKAGE



Enlarged photo of header with can off. 7 times actual size.



Cross-section of header and isolated transistor.

Gold-plated Kovar island is electrically welded to collector lead and fused onto glass insulator completely isolating transistor from can.

Fairchild now offers Planar transistors with collectors electrically isolated from their TO-5 packages. The entire transistor die is isolated; it is bonded to a gold-plated Kovar "collector island" which is fused onto a layer of glass, providing positive insulation from the rest of the package. All three leads are glass-insulated as they pass through the header platform. This isolation means you may now design using Planar transistors which are independent of their conductive packages in the circuit. Electrically isolated transistors eliminate the cost of special design and/or assembly necessary for transistor insulation. This type of construction also offers excellent resistance to thermal shock, mechanical shock and vibration. Mounting of the silicon transistor die and attachment of leadwires inside are accomplished by standard Fairchild bonding operations.

2N1893
2N911
2N910

These types offered in TO-5 Isolated Collector packages

2N1893 Type – SP8400 **2N910 Type – SP8401**
2N911 Type – SP8402

Data sheets available on all items.

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29 Fields of Special Interest-

The 29 Professional Groups are listed below, together with a brief definition of each, the name of

Aerospace and Navigational Electronics

Annual fee: $3.

The application of electronics to operation and traffic control of aircraft and to navigation of all craft.

Mr. S. H. Dodington, Chairman, ITT Labs., 500 Washington Ave., Nutley, N.J.

44 Transactions, *9, Vol. 2, No. 3; Vol. 4, No. 1, 2, 3; Vol. 5, No. 2, 3, 4; Vol. 6, No. 1, 2, 4; Vol. 7, No. 1, 2, 3, 4; Vol. 8, No. 1, 2, 3, 4; Vol. 9, No. 1, 2.

Antennas and Propagation

Annual fee: $6.

Technical advances in antennas and wave propagation theory and the utilization of techniques or products of this field.

Dr. S. A. Bowhill, Chairman, Dept. of E. E., Univ. of Illinois, Urbana, Ill.

51 Transactions, *Vol. AP-2, No. 2; AP-4, No. 4; AP-5, No. 1-4; AP-6, No. 1, 2, 3, 4; AP-7, No. 1, 2, 3, 4; AP-8, No. 1, 2, 3, 4, 5, 6; AP-9, No. 1, 2, 3, 4, 5, 6; AP-10, No. 1, 2, 3.

Audio

Annual fee: $2.

Technology of communication at audio frequencies and of the audio portion of radio frequency systems, including acoustic terminations, recording and reproduction.

Dr. Robert W. Benson, Chairman, Robert W. Benson & Assoc., Inc., Nashville 5, Tenn.

65 Transactions, *Vol. AU-2, No. 4; Vol. AU-3, No. 1, 3, 5; Vol. AU-4, No. 1, 6; Vol. AU-5, No. 1, 2, 3, 4, 5, 6; AU-6, No. 1, 2, 3, 4, 5, 6; AU-7, No. 1, 2, 3, 4, 5, 6; AU-8, No. 1, 2, 3, 4, 5, 6; AU-9, No. 1, 2, 3, 4, 5, 6; AU-10, No. 1, 2, 3.

Automatic Control

Annual fee: $4.

The theory and application of automatic control techniques including feedback control systems.

Mr. Louis B. Wadel, Chairman, 3546 Caruth Blvd., Dallas 25, Tex.

20 Transactions, PGAC-3-4-5-6, AC-4, No. 1, 2, 3; AC-5, No. 1, 2, 4; AC-6, No. 1, 2, 3; AC-7, No. 1, 2, 3, 4.

Bio-Medical Electronics

Annual fee: $4.

The use of electronic theory and techniques in problems of medicine and biology.

Dr. O. H. Schmitt, Chairman, 1912 Como Ave., Minneapolis 14, Minn.

25 Transactions, 8, 9, 11; ME-6, No. 3, 4; ME-7, No. 2, 4; BME-8, No. 2, 3, 4; BME-9, No. 1, 2.

Broadcast & Television Receivers

Annual fee: $4.

The design and manufacture of broadcast and television receivers and components and activities related thereto.

Mr. John F. Bell, Chairman, Zenith Radio Corp., 6001 W. Dickens Ave., Chicago 39, Ill.

32 Transactions, *7, 8; BTR-1, No. 1-3, BTR-2, No. 1-2-3; BTR-3, No. 1-2; BTR-4, No. 2, 3-4; BTR-5, No. 1, 2; BTR-6, No. 1, 2, 3; BTR-7, No. 1, 2, 3; BTR-8, No. 1, 2.

Broadcasting

Annual fee: $2.

Broadcast transmission systems engineering, including the design and utilization of broadcast equipment.

Mr. Raymond F. Guy, Chairman, 264 Franklin St., Haworth, N.J.

22 Transactions, No. 10, 11, 12, 13, 14; BC-6, No. 1, 2, 3; BC-7, No. 1, 2, 3, 4, BC-8, No. 1.

Circuit Theory

Annual fee: $4.

Design and theory of operation of circuits for use in radio and electronic equipment.

Dr. James H. Mulligan, Jr., Chairman, College of Eng., New York University, New York 53, N.Y.

38 Transactions, CT-4, No. 3-4; CT-5, No. 1, 2, 3; CT-6, No. 2, 3, 4; CT-7, No. 1, 2, 3, 4; CT-8, No. 2; CT-9, No. 1, 2.

Communications Systems

Annual fee: $4.

Radio and wire telephone, telegraph and facsimile in marine, aeronautical, radio-relay, coaxial cable and fixed station services.

Mr. Ralph L. Marks, Chairman, Rome Air Dev. Center, Griffiss AFB, N.Y.

25 Transactions, CS-5, No. 2, 3; CS-6, No. 1, 2; CS-7, No. 1, 3, 4; CS-8, No. 1, 2, 3, 4; CS-9, No. 1, 2, 3, 4; CS-10, No. 1.

Component Parts

Annual fee: $3.

The characteristics, limitation, applications, development, performance and reliability of component parts.

Dr. Louis Kahn, Chairman, Astron Corp., 255 Grant Ave., East Newark, N.J.

28 Transactions, CP-4, No. 1, 2, 3, 4; CP-5, No. 2, 3, 4; CP-6, No. 1, 2, 3, 4; CP-7, No. 1, 2, 3, 4; CP-8, No. 2, 3, 4; CP-9, No. 1, 2.

Education

Annual fee: $3.

To foster improved relations between the electronic and affiliated industries and schools, colleges, and universities.

Mr. George E. Moore, Chairman, Westinghouse Electric Corp., East Pittsburgh, Pa.

17 Transactions, Vol. E-1, No. 3, 4; E-2, No. 1, 2, 3, 4; E-3, No. 1, 2, 3, 4; E-4, No. 1, 2, 3, 4; E-5, No. 1.

Electron Devices

Annual fee: $5.

Electron devices, including particularly electron tubes and solid state devices.

Mr. R. W. Sears, Chairman, Bellcomm Inc., 1737 L St. N.W., Washington, D.C.

41 Transactions, *Vol. ED-1, No. 4; ED-3, No. 2, ED-4, No. 3, 4; ED-5, No. 2, 3, 4; ED-6, No. 3; ED-7, No. 2, 3, 4; ED-8, No. 1, 3, 4, 5, 6; ED-9, No. 1, 2, 3, 4, 5.

Electronic Computers

Annual fee: $4.

Design and operation of electronic computers.

Mr. Walter L. Anderson, Chairman, General Kinetics Inc., So. Arlington, Va.

42 Transactions, EC-6, No. 2, 3; EC-7, No. 1, 2, 3, 4; EC-8, No. 1, 2, 3, 4; EC-9, No. 1, 2, 3, 4; EC-10, No. 2, 3, 4; EC-11, No. 1, 2, 3.

Engineering Management

Annual fee: $3.

Engineering management and administration as applied to technical, industrial and educational activities in the field of electronics.

Mr. Edward S. White, Chairman, Warwick Mfg. Co., 7300 N. Lehigh Ave., Chicago 31, Ill.

26 Transactions, EM-4, No. 1, 3, 4; EM-5, No. 1-4; EM-6, No. 1, 2; EM-7, No. 1, 2, 3; EM-8, No. 1, 2, 3, 4; EM-9, No. 1, 2.

Engineering Writing and Speech

Annual fee: $3.

The promotion, study, development, and improvement of the techniques of preparation, organization, processing, editing, and delivery of any form of information in the electronic-engineering and related fields by and to individuals and groups by means of direct or derived methods of communication.

Mr. Chester W. Sall, Chairman, 121 Grover Ave., Princeton, N.J.

11 Transactions, Vol. EWS-1, No. 2; EWS-2, No. 1, 2, 3; EWS-3, No. 1; EWS-4, No. 2, 3; EWS-5, No. 1.

THE INSTITUTE OF RADIO

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the group chairman, and publications to date.

*** Indicates publications still available**

Geoscience Electronics

Annual fee: $2.

Research and development in electronic instrumentation for geophysics and geochemistry, especially gravity measurements, seismic measurements, magnetics, well-logging, space exploration, meteorology, oceanography, etc.

Mr. Robert W. Olson, Texas Instruments Co., Dallas, Tex.

Human Factors in Electronics

Annual fee: $3.

Development and application of human factors and knowledge germane to the design of electronic equipment.

Dr. Albert S. Jackson, Chairman, Control Technology, Inc., 1232 Belmont Ave., Long Beach 4, Calif.

5 Transactions, HFE-1, No. 1, 2; HFE-2, No. 2; HFE-3, No. 1.

Industrial Electronics

Annual fee: $3.

Electronics pertaining to control, treatment and measurement, specifically, in industrial processes.

Mr. J. E. Eiselein, Chairman, RCA Victor Div., Camden, N. J.

17 Transactions, *PGIE 3, 5, 6, 7, 8, 9, 10 11; IE-7, No. 1, 2, 3; IE-8, No. 1, 2; IE-9, No. 1.

Information Theory

Annual fee: $4.

The theoretical and experimental aspects of information transmission, processing and utilization.

Dr. David Van Meter, Chairman, 1716 Cambridge St., Cambridge 38, Mass.

36 Transactions, PGIT-4, IT-1, No. 3; IT-3, No. 1, 2, 3; IT-4, No. 1, 3, 4; IT-5, No. 1, 2, 3, 4; IT-6, No. 1, 4, 5; IT-7, No. 1, 3, 4; IT-8, No. 1, 2, 3, 4, 5.

Instrumentation

Annual fee: $3.

Measurements and instrumentation utilizing electronic techniques.

Mr. L. D. McLellan, Chairman, Hughes Aircraft Co., Culver City, Calif.

22 Transactions, PGI-4, Vol. 1-6, No. 2, 3, 4; Vol. 1-7, No. 1, 2; Vol. 1-8, No. 1, 2; Vol. 1-9, No. 1, 2, 3; Vol. 1-10, No. 1, 2; Vol. 1-11, No. 1.

Microwave Theory and Techniques

Annual fee: $4.

Microwave theory, microwave circuitry and techniques, microwave measurements and the generation and amplification of microwaves.

Dr. Seymour B. Cohn, Chairman, Rantec Corporation, 23999 Ventura Blvd., Calabasas, Calif.

44 Transactions, MTT-4, No. 3; MTT-5, No. 3, 4; MTT-6, No. 1, 2, 3, 4; MTT-7, No. 2, 3; MTT-8, No. 1, 3, 5, 6; MTT-9, No. 1, 3, 4, 5, 6; MTT-10, No. 1, 2, 3, 4.

Military Electronics

Annual fee: $4.

The electronics sciences, systems, activities and services germane to the requirements of the military. Aids other Professional Groups in liaison with the military.

Mr. Edward N. Dingley, Jr., Chairman, Electronic Communications, Inc., 1501-72 St. N., Saint Petersburg 10, Fla.

17 Transactions, MIL-1, No. 1; MIL-2, No. 1; MIL-3, No. 2, 3, 4; MIL-4, No. 2-3, 4; MIL-5, No. 1, 3, 4; MIL-6, No. 1, 2, 3.

Nuclear Science

Annual fee: $3.

Application of electronic techniques and devices to the nuclear field.

Dr. W. A. Higinbotham, Brookhaven Natl. Lab., Upton, L.I., N.Y.

26 Transactions, NS-1, No. 1; NS-4, No. 2; NS-5, No. 1, 2, 3; NS-6, No. 1, 2, 3, 4; NS-7, No. 1, 2-3, 4; NS-8, No. 2, 3, 4; NS-9, No. 2, 3, 4.

Product Engineering & Production

Annual fee: $2.

New advances and materials applications for the improvement of production techniques, including automation techniques.

Mr. John W. Trinkaus, Chairman, Sperry Gyroscope Co., Mail Station 1A36, Great Neck, L.I., N.Y.

12 Transactions, No. 3, 4, 5, 6; PEP-5, No. 1, 2, 3, 4; PEP-6, No. 2.

Radio Frequency Interference

Annual fee: $2.

Origin, effect, control and measurement of radio frequency interference.

Mr. Herman Garlan, Chairman, Office of Chief Engineer, Federal Communications Comm., Washington, D.C.

5 Transactions, RFI-1, No. 1, RFI-2, No. 1, RFI-3, No. 1; RFI-4, No. 1, 2.

Reliability and Quality Control

Annual fee: $3.

Principles and practices used in reliability and quality engineering of electronic products.

Mr. L. J. Paddison, Chairman, Sandia Corp., Sandia Base, Albuquerque, N.M.

24 Transactions, 11, 12, 13, 14, 15; RQC-9, No. 1, 2, 3; RQC-10, No. 2, 3; RQC-11, No. 1, 2.

Space Electronics and Telemetry

Annual fee: $3.

The control of devices and the measurement and recording of data from a remote point by radio.

Mr. H. B. Riblet, Chairman, 3300 Plyrrs Mill Rd., Kensington, Md.

22 Transactions, TRC-1, No. 2-3; TRC-2, No. 1; TRC-3, No. 2, 3; TRC-4, No. 1; SET-5, No. 1, 2, 3, 4; SET-6, No. 1, 2, 3-4; SET-7, No. 1, 2, 3, 4; SET-8, No. 1, 2.

Ultrasonics Engineering

Annual fee: $2.

Ultrasonic measurements and communications, including underwater sound, ultrasonic delay lines, and various chemical and industrial ultrasonic devices.

Mr. Allen H. Meitzler, Chairman, 12 Harding Road, Morristown, N.J.

11 Transactions, PGUE, 5, 6, 7; UE-7, No. 1, 2; UE-8, No. 1, UE-9, No. 1.

Vehicular Communications

Annual fee: $3.

Communications problems in the field of land and mobile radio services, such as public safety, public utilities, railroads, commercial and transportation, etc.

Mr. S. C. Bartlett, Chairman, Amer. Elec. Power Serv. Corp., 2 Broadway, New York 8, N.Y.

19 Transactions, 5, 8, 9, 10, 12, 13; Vol. VC-9, No. 1, 2, 3; Vol. 10, No. 1, 2; Vol. 11, No. 1.

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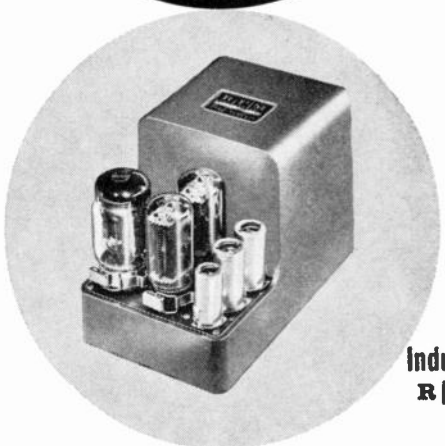
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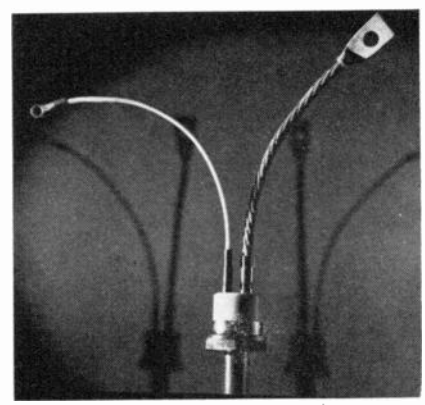
NEWS New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your IRE affiliation.

(Continued from page 66A)

Controlled Rectifiers

A 70-ampere "Rock-Top" Trinistor controlled-rectifier line (JEDEC 2N1792 and 2N1909 Series) with forward blocking voltage to 400 volts and peak reverse voltage to 480 volts is offered by the **Westinghouse Electric Corp., Semiconductor Div.,** Youngwood, Pa. This three-terminal, four-layer PNPN silicon controlled rectifier features a turnoff time of 10 to 20 microseconds. Turnon time is 0.6 to 3.0 microseconds and the maximum dc current is 110 amperes.



Typical applications include high-power inverters, light dimmers, motor control, magnetic amplifier replacement, ignitron firing, frequency changers and dc power regulators.

The increased voltage capabilities permit savings in equipment size and assembly cost, because there no longer is a need to connect two or more devices in series to achieve proper voltage protection.

Additional features include hard soldered construction that eliminates thermal fatigue problems due to heat cycling, low thermal impedance, glazed ceramic headers and hermetically weld-sealed cases to insure dependable operation under severe application conditions.

The rectifiers are available from stock at authorized distributors. Prices of the units vary from $28.70 to $170 in quantities of one to 24, and in quantities of 25 to 99, the price range is $21.50 to $134.

For more information, write to the firm.

Digital Test Equipment

A new class of digital test equipment has been introduced by **Consolidated Avionics Corp.,** 800 Shames Drive, Westbury, L. I., N. Y., a subsidiary of Consolidated Diesel Electric Corp.

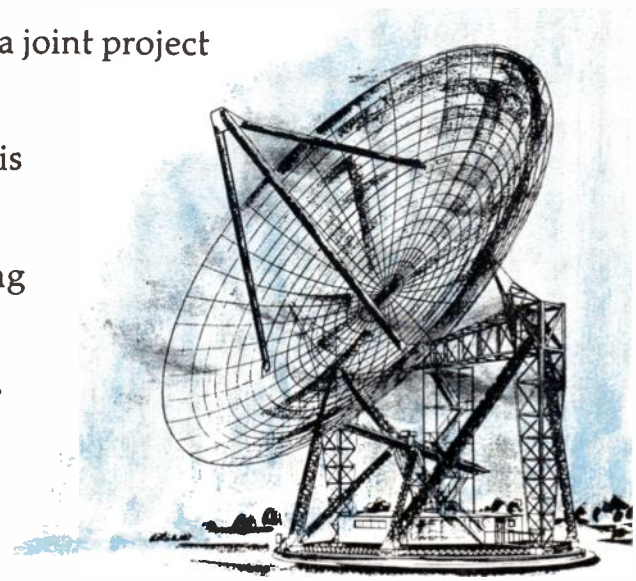


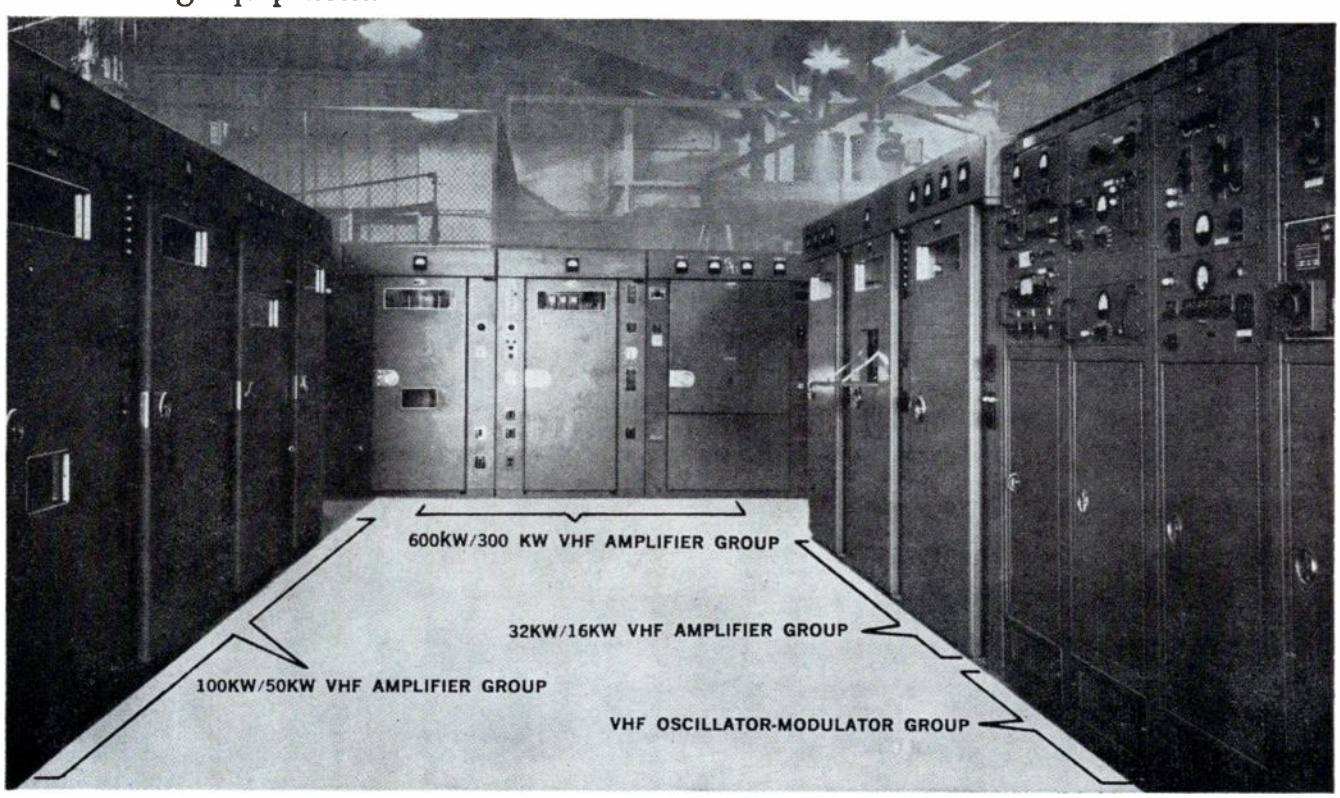
Designed to perform various timing, pattern generation, pulse generation and control functions, the new device, selling for $2350, is claimed to replace racks full of special equipment which would be needed to perform the same operations and to eliminate hundreds of manhours required to set up the operations from currently-available components. The firm claims that the savings in equipment and labor will permit the performance of tests now

(Continued on page 72A)

STANFORD'S RADAR TELESCOPE FOR SPACE RESEARCH...

Standing atop the ridge of hills at the back of the Stanford University campus is a 15-story dish antenna, a joint project of Stanford Research Institute and the University. The 150-foot parabolic reflector is illuminated by a Continental Electronics high power, very high frequency transmitting equipment, and is part of one of the finest radar astronomy establishments in the world. Another example of Continental's continuing leadership in super power transmitting equipment.





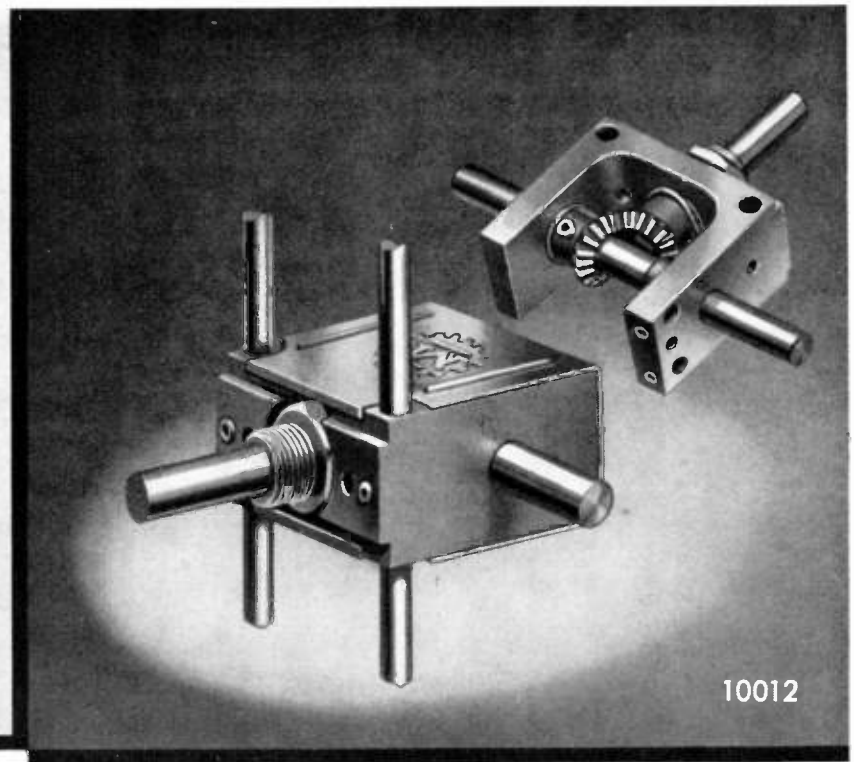
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The new Interstate Model LF 400 VLF Receiver/Correlator tunes any of three VLF worldwide transmissions, and electronically compares the signal phase to a local oscillator for calibration of local equipment. Accuracy of 3 parts in 10 billion over a few hours is easy to attain. In addition, the rugged solid state unit, with 1 microvolt sensitivity, provides an output timing pulse synchronized to the incoming VLF signal. Write for brochure.

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NATIONWIDE REPRESENTATIVES

ANOTHER Interstate SOLID-state INSTRUMENT

NEWS New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your IRE affiliation.

(Continued from page 70A)

considered impractical from a time and cost standpoint.

Known as Vari-Bit Model 8003, the new equipment provides several functions at the turn of a single selector switch.

1. It can function as a continuously variable delay line with delays ranging from 10 to 1000 μs. Pulse repetition rates can be as high as 250 kc.

2. As a digital function generator, it can provide pulse patterns containing as many as 248 pulses, provided there are at least four microseconds between pulses. Rate control can be either internal or external.

3. As a variable frequency pulse generator with fixed pulse width, the unit provides pulses at frequencies between 1 and 100 kc.

4. As a variable frequency pulse generator with variable pulse widths, the unit provides frequencies between 1 and 100 kc with pulse widths ranging from four microseconds upwards to the limit imposed by periodicity of output.

5. It can divide frequencies as high as 500 kc by factors from 2 to 500.

6. For frequency multiplication, it can multiply any frequency pulse train between 500 cps and 125 kc by factors from 2 to 248. The upper limit on the output frequency is 250 kc.

7. Set up as a variable capacity serial memory, the unit acts as a dynamic recirculating storage register with built-in reclocking and erasure controls.

8. Other special functions can be provided by the insertion of additional logic cards for which capacity has been provided.

Functioning of the Vari-Bit Model 8003 is based on the Con

(Continued on page 74A)

the three sides to sell in RADIO AND ELECTRONICS



The fellows who write the big orders in this business know their market. And they know that hitting that market with year-in, year-out consistency is a three-way proposition. The reason is simple. The radio and electronics engineers of America believe in and wholeheartedly support these three *Institute of Radio Engineers'* services.

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Here's the engineer's monthly fact sheet—with more pages, more editorial features, and more readers than any publication in its field. Over 80,000 radio-electronics engineers read *Proceedings* because they want to know what's going on and they've discovered *Proceedings* is the place to find out. It's on top of the latest developments in this fast-moving, fast-expanding field.

IRE Directory

More than 1,000 pages long, with circulation well above the 75,000 mark, this unique annual covers the radio-electronics industry from stem to stern. In the Directory you'll find a pinpointed classification of 5,500 manufacturers and 2,000 distributors of radio and electronic equipment along with the products they handle. To help you clinch the sale the *IRE Directory* lists business phone numbers. No wonder they call the *IRE Directory* the "Bluebook" of the radio-electronics industry.

Radio Engineering Show

The radio-electronics "spectacular" held annually at New York's mammoth Coliseum. Here's your personal introduction to 70,000 radio-electronics engineers—the men who use the products you sell and manufacture. They're eager to see your new equipment, to listen to your new ideas. This year, don't pass up this golden opportunity to meet them on their home grounds.

For all the details:

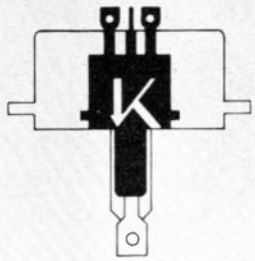
The Institute of Radio Engineers

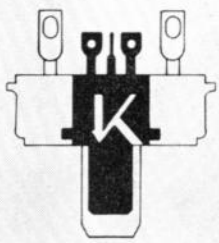
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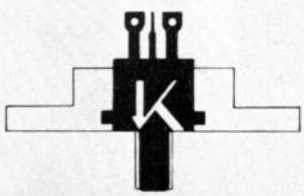
KEARFOTT

KEARFOTT power transistors average 30% lighter, 50% smaller and occupy 50% less mounting surface. **NEW** modified TO-10 hermetically sealed case and improved internal construction provide miniaturized **GERMANIUM PNP POWER** devices with ratings up to **120V @ 15 AMPS** (our #KPG2000). We are also a **DEPENDABLE SOURCE** for the following devices with ratings up to **200V@ 3AMPS:** 2N538 through 2N540A, 2N1202, 2N1203, 2N1326, 2N1438, 2N1466, 2N1501, 2N1502, 2N1504/10. Also available in TO-13 cases: 2N143/13, 2N156, 2N158, 2N158A, 2N1437, 2N1465.

The following **DISTRIBUTORS** stock these devices: Carter Assoc. Inc., Scottsdale, Arizona; Cramer Electronics, Inc., Newton, Mass.; Hollywood Radio & Electronics, Hollywood, California; Solid State Specialist, Mountain View, California; Terminal-Hudson Electronics, New York City; Valley Electronics, Inc., Towson, Maryland. Write Kearfott Division, General Precision, Inc., Little Falls, New Jersey or 437 Cherry Street, West Newton, Massachusetts. These devices are designed, manufactured and life tested by KEARFOTT SEMICONDUCTOR CORP., West Newton, Mass.


TO-36


MT-7


TO-3


TO-10


RO-28-1

Note how little volume and space are required by Kearfott's Transistors

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NEWS New Products

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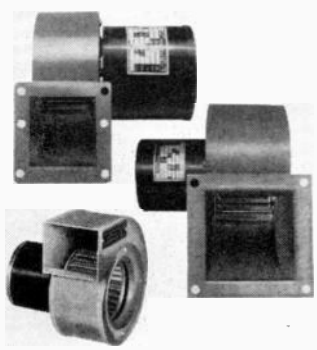
(Continued from page 72A)

Avionics "Torsmag" principle introduced last fall. The basic phenomenon involved in the "Torsmag" principle is that the fields of permanent magnets spaced along a torsional delay line generate pulses in the line as a wave traveling the length of the line passes through the magnetic fields. Pulse timing is determined by the spacing of the permanent magnets along the line.

The Vari-Bit Model 8003 consists of two interconnected units—a "Torsmag" line and a control unit. The latter contains the logic circuitry and switching control required for the various modes of operation and the necessary power supplies.

Input power requirement, aside from various signals required when the unit is functioning in conjunction with external equipment, is 115-volt, 60-cycle ac.

MIL-SPEC Blower Catalog



A line of MIL-SPEC packaged blowers for electronic rack cooling is presented in a new catalog published by **McLean Engineering Laboratories,** P. O. Box 228, Princeton, N. J. The line includes packaged blowers in a wide variety of specifications and performances, ring fans, panel fans, single and dual centrifugal blowers, control panels, and special cooling equipment. The processes, finishes, materials, components and motors in each of the modular series are covered by military specifications. All MIL-SPEC motors and blowers are custom designed and built by McLean specially for each application, assuring against the possibility of mismatched items. A special section of the book lists the principal MIL-SPECS that apply to cooling equipment for electronics. Another section describes housing variations and blower discharge variations of the rack units. McLean's Air Flow Test Chamber is also described. Copies of the catalog may be had by writing to the firm.

(Continued on page 76A)

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BIPCO® LOGIC is low cost and reliable

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Typical 96 diode BIPCO Module.
(Measures 5/16" x 1 1/8" x 15/16")

BIPCO Logic represents an entirely new approach to microelectronics. Batches of identical components are "Built-In-Place" at one time in patterns determined by the logical function to be performed. Individual component handling is eliminated, interconnections are simplified, and electrical uniformity is achieved. For the first time, the cost barrier inherent in other approaches has been eliminated so that reliability *and* low cost can be obtained.

Functional diode logic is now available in individual BIPCO Modules or in printed circuit packages including: Decimal Counters, Binary and Alpha-Numeric Decoders, Encoders, and Distributors. BIPCO Transistor, Resistor, and Capacitor Modules will be available soon.

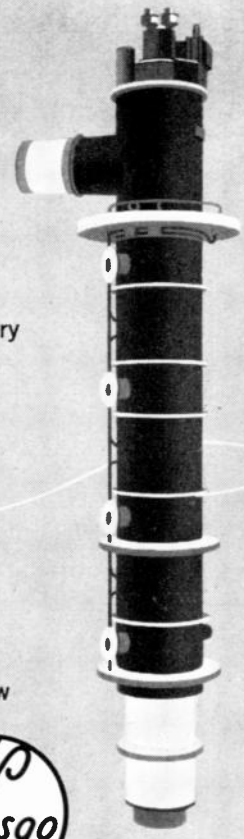
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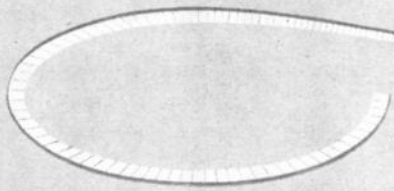
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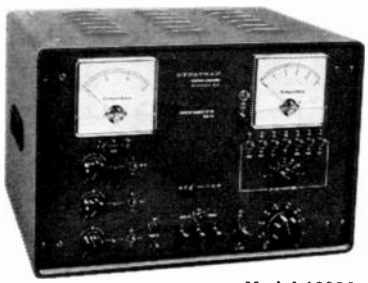
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NEWS New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your IRE affiliation.

(Continued from page 74A)

Thermistors Bulletin

A new 16-page booklet on negative temperature coefficient resistors (thermistors, or NTC's) is available from **Ferroxcube Corp. of America,** Saugerties, N. Y. Included is information on several standard shapes manufactured by Ferroxcube; cylindrical rods, discs, thermometer types, thermistors with heating elements and beads.

Applications of NTC's are discussed in detail, as is the determination of their requirements for selecting the right type and value. Complete diagrams and graphs of the characteristics of each type are provided—including special types.

The booklet is free on request.

(Continued on page 119A)

Section Meetings

BENELUX

"Error Probabilities For Data Transmission Over Fading Radio Paths," B. B. Barrow, SHAPE. 5/22/62.

"Simulators For Teaching Navy Operators and For Wireless Controlled Operations," J. P. J. Nabben, N. V. van der Heem; "A Single Side Band Communication System," S. J. Noteboom, N. V. van der Heem. 6/27/62.

BUENOS AIRES

"Some Experiences in Installation and Operation of Radioelectric Networks," O. R. Canese. 4/12/62.

"IRE Standards For T.V. Receivers," C. Santoalla and P. J. Martinez; Beginning of Electronics Week. 5/14/62.

"Automatic Control Systems—New Concepts and Applications," J. H. Decanini; Tour of Power House installations of Buenos Aires Electric Supply Co. 5/15/62.

"Electronics Contribution to Guided Missiles," T. Hasduk; Visit to IBM World Trade Installations. 5/16/62.

Visit to Esso Oil Co. installations and Cracking Plant. 5/17/62.

Visit to Philips Factory. Formal Electronics Week Dinner. 5/18/62.

CANAVERAL

"The Technological Revolution," Dr. P. Haggerty, IRE President; Installation of Section officers. 6/15/62.

HAWAII

"Laser Light—Its Hardware, Packaging and Applications," N. Najita, Univ. of Hawaii. 8/8/62.

HUNTSVILLE

"Resolvers for Missile Guidance," L. S. Kahn, Bendix Corp. 3/30/62.

"The Saturn Space Booster," O. H. Lange, NASA. Tour of NASA facilities and presentation of Student Awards. 5/4/62.

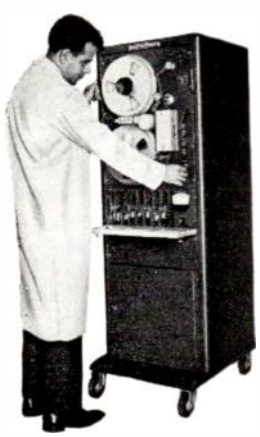
"Computers of the Future," D. Metzger, IBM. 6/1/62.

Business meeting and election of officers. 6/29/62.

(Continued on page 78A)



SANBORN FLEXIBILITY — 7 channels — fm, direct record or any combination/plug-in, all solid-state circuitry/record-reproduce amplifiers on same card/4 speeds — 3¾ to 30 ips; 1⅞ to 15 ips and 7½ to 60 ips optional / 7" high electronics available separately/optional extras include voice channel amplifier, digital input circuit, push-pull input coupler, precision footage indicator, loop adapter and remote control unit. AT AN UNMATCHED $7200.



This new Sanborn/Ampex Model 2007 system conforms to accepted IRIG instrumentation standards, provides 1% system accuracy and bandwidths to 100,000 cps with direct recording, 10,000 cps with FM amplifiers. Max. error due to non-linearity is only ±0.5% on DC, ±1% on AC.

Basic system features include quickly interchanged, readily accessible printed circuit plug-in modules . . . flutter compensation by using one channel to compensate all others . . . alignment of all FM channels with built-in meter and selector switch, eliminating need for electronic counters . . . automatic squelch circuit . . . entire system in only 31" of rack panel space . . . packaging in either mobile console shown or portable cases for tape transport and electronics.

System price of $7200 includes 7-channel tape transport, transfer chassis, playback preamplifiers, power supply and 7 channels of FM Record/Reproduce electronics, housed in metal mobile cabinet. All prices F.O.B. Waltham, Mass., and subject to change without notice.

Get the complete specifications on this new Tape System — as well as 3 new types of Sanborn Data Amplifiers, 17" Multi-Trace Scope and other related instrumentation — from your local Sanborn Industrial Sales-Engineering Representative. Ask him for your copy of the complete Industrial Catalog.

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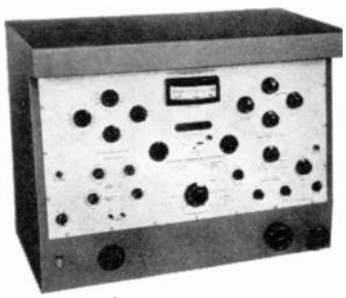
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Section Meetings

(Continued from page 76A)

INDIANAPOLIS

"Aspects of the IRE-AIEE Merger," W. Bruene, Collins Radio Co. Annual banquet and installation of officers. 6/28/62.

ISRAEL

"Circuit Design of Digital Computers," S. Ruhman, Univ. of California. 2/13/62.

"Medical Electronics," All-day session with series of lecturers followed by discussions. 5/22/62.

LONG ISLAND

"The Basic Nature of Stereo"—Lecture Series, C. J. Hirsch, RCA. 2/8/62.

"Principles & Problems of 'Unconventional' Energy Conversion," P. H. Klein, Sperry Rand Research Center; Film—"1961 Indianapolis '500' Auto Race." 2/13/62.

"Studio Stereo Recording"—Lecture Series, D. L. Richter, RCA. 2/15/62.

"F.M. Stereophonic Broadcasting," C. G. Eilers, Zenith Radio Corp. 3/1/62.

"Home Stereo Reproduction"—Lecture Series, D. R. Von Recklinghausen, H. H. Scott, Inc. 3/8/62.

"Loud Speakers & Room Acoustics," Lecture Series, A. B. Cohen, Acoustic Developments, Inc. 3/15/62.

Fellow Awards Presentations and Annual Cocktail party. 3/25/62.

"The Orbiting Astronomical Observatory—A Space Laboratory," D. J. Shergalis, Grumman Engineering Corp.; Film—"Circle of Confidence." 4/10/62.

"The Outlook for Electronics"—Panel discussion. Panelists: A. J. Catapano, Merrill Lynch, Pierce, Fenner & Smith, Inc.; P. Clark, Burnham & Co.; P. Leserman, Leserman, Stenbuck, Ltd.; R. C. Sellers, R. C. Sellers & Associates. 5/8/62.

Section field trip to Sperry Rand Polaris Facility. Films—"X Marks the Spot" and "Count Down for Polaris." S. Noss, Sperry Rand. 6/16/62.

NORTH CAROLINA

"Nuclear Power," A. C. Thies, Duke Power Co. 6/15/62.

TOLEDO

Conducted Tour of Dana Corporation. 4/12/62

"Proposed Merger of IRE and AIEE," A. Rauth, AIEE. 5/17/62.

TWIN CITIES

"Radio Astronomy at the National Radio Astronomy Observatory," J. W. Findley, NRAO 2/7/62.

"IRE Plus AIEE Equals IEEE," Dr. P. Haggerty, President IRE. Tour of new Control Data Plant. 5/2/62.

SUBSECTIONS

BUENAVENTURA

"Navy Space Objectives," R. F. Freitag, BuWeps, Washington; "Project Transit," Dr. R. B. Kershner, Johns Hopkins Univ. 5/11/62.

"RF Preamplification," Dr. W. E. Ayer, Applied Technology, Inc., 6/13/62.

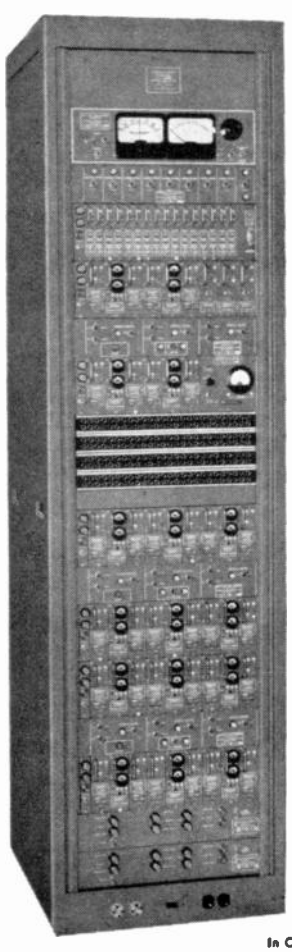
MID-HUDSON

"Target Catching by Bats," D. Cahlander, MIT Lincoln Labs. 3/20/62.

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"Air Traffic Control," R. Varese, FAA. Election of officers. 4 17/62.

Annual Dinner Meeting and Dance. 6 9 62.

ORANGE BELT

"New Electronic Frontiers in Oceanography," T. Higgins, Jr., Lockheed. Tour of Lockheed Aircraft Service facilities. 6 12 62.

PANAMA CITY

"Digital Computers, Past, Present and Future," C. M. Callahan, US Navy Mine Defense Laboratory; Computer demonstration, H. Lawton. 7 31/62.

SANTA BARBARA

Annual Dinner Dance. 6/16/62.

WESTERN NORTH CAROLINA

"Data Transmission," J. Wolf, Southern Bell Telephone Co. 4/27/62.

"Meter Repair and Calibration," M. Little, Electro-Tech. 5/26/62.

"Atomic Power Plant at Parr, South Carolina," A. C. Thies, Duke Power Co. 6/15/62.

Professional Group Meetings

ANTENNAS AND PROPAGATION

Los Angeles—November 9

"A Self-Focusing or Active Adaptive Antenna Array," Dr. A. F. Gangi, Space General Corp., Los Angeles, Calif.

Los Angeles—March 8

"The Importance of Being Coherent," W. A. Snyder, Hughes Aircraft Co., Malibu, Calif.

San Francisco—April 11

"HF Communications Frequency Management Techniques and Frequency Independent Antenna Applications," R. W. Berg, Granger Associates, Palo Alto, Calif.

San Francisco—May 9

"Theory and Design of H-Guide Antennas," G. N. Voronoff, Dalmo Victor Co.

AUDIO

Twin Cities—April 11

"Wide Band Recorders," Vic Mohrlandt, Minnesota Minning and Manufacturing, St. Paul, Minn.

AUTOMATIC CONTROL

Long Island—January 30

"Status of Numerical Controls in the U. S. Today," James Childs, TRW Computers.

(Continued on page 80A)

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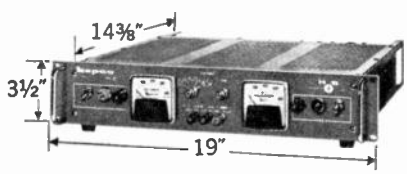
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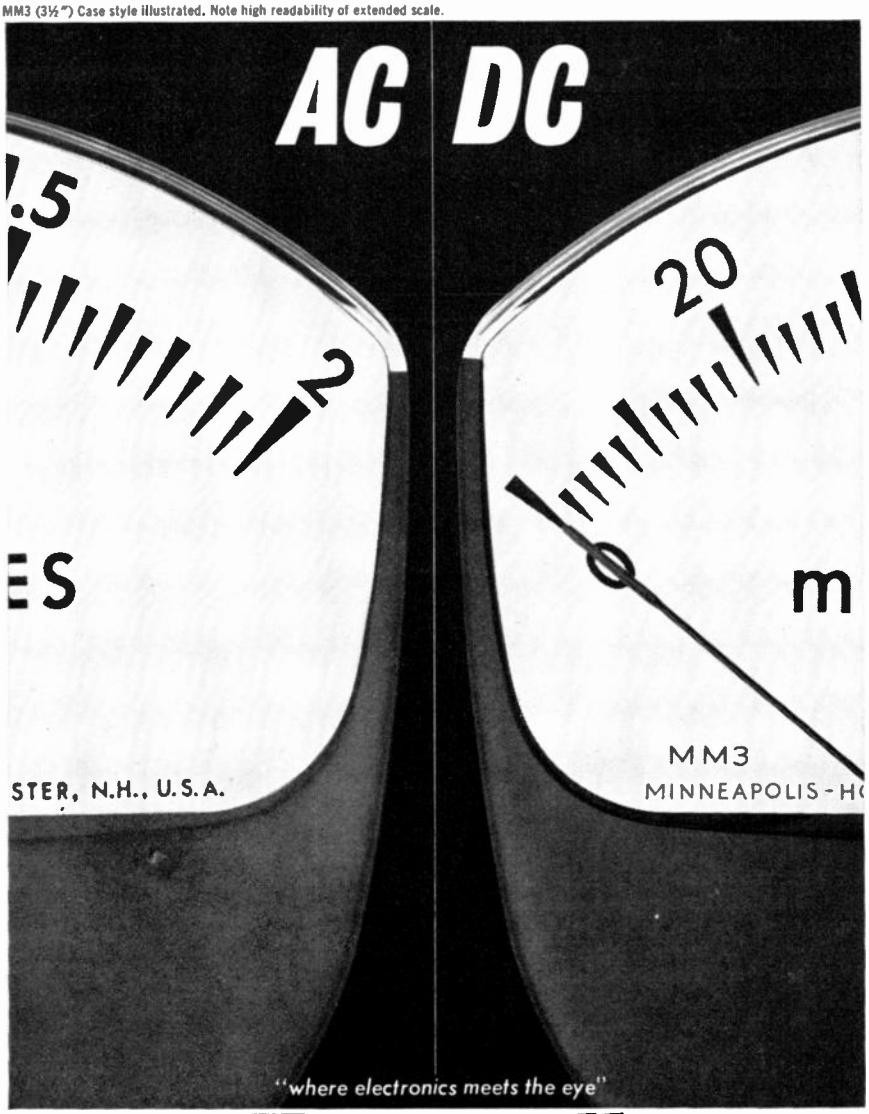
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MM3 (3½") Case style illustrated. Note high readability of extended scale.



Honeywell

Precision Meters

Professional Group Meetings

(Continued from page 79A)

Long Island—March 6

"Unsolved Problems in Control Theory," Dr. Ludwig Braun, Polytechnic Institute of Brooklyn.

Long Island—June 12

"A Guided Tour Through FAA Control Center, Idlewild Airport," Lincoln Marvin, FAA.

Twin Cities—February 14

"Adaptive Autopilot for Aerospace Application," Dave Mellen, Minneapolis Honeywell, Minneapolis, Minn.

Twin Cities—January 23

"Magnetic and Solid State Control Systems," D. F. Marks, Telex-Lumen Corp.

Electron Devices

Boston—May 22

Panel Discussion: "Microelectronics as of 5/22/62," D. A. McLean, Bell Labs.; S. M. Stubebard, P. R. Mallory, Inc.; G. J. Selvin, Sylvania Company; E. B. Sussman, Texas Instruments.

Twin Cities—January 17

"Magnetic Film Memories and Devices," Arthur Pohm, Iowa State University, Ames, Iowa.

Electronic Computers

San Francisco—May 22

"Error Correcting Codes and a Particular Physical Realization," Dr. Bernard Elspas and W. K. English, SRI, Menlo Park, Calif.

San Francisco—June 26

"Opto Electronic Components and Their Applications to Data Processing Machines," Frank Litz, Opto Electronic Devices Inc.

Engineering Management

Twin Cities—April 19

"Coming Issues in Management," Dr. P. V. Grambsch, School of Business, University of Minnesota, Minneapolis, Minn.

Engineering Writing and Speech

Twin Cities—May 23

"Semantics," Harold Alford, University of Minnesota, Minneapolis, Minn.

(Continued on page 82A)

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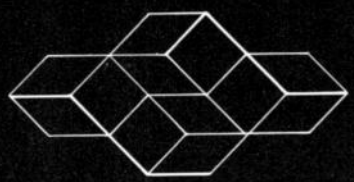
Find out how readily and inexpensively your most exacting S, C, X, and L Band coupler requirements can be satisfied. Write us at 185 N. Main St., Bristol, Connecticut, outlining your specifications in terms of frequency range, power level, coupling attenuation and type of waveguide.

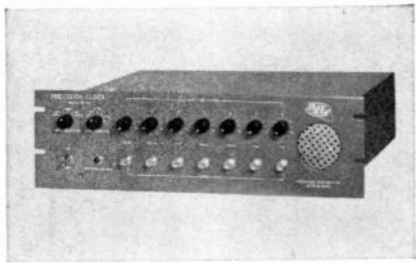
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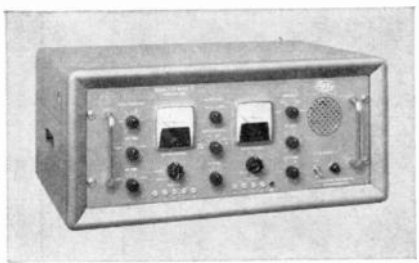
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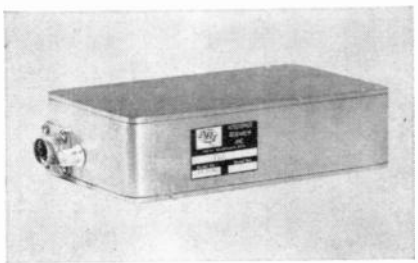




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Professional Group Meetings

(Continued from page 80A)

Instrumentation

San Francisco—June 27

"Coupling Systems to Computers," Norman Dawirs, Astrodata, Inc., Los Angeles, Calif.

San Francisco—May 23

"Information Storage in Data Systems, Bob Flanders, Ampex, Redwood City, Calif.

San Francisco—April 25

"Information Processing," William Gross, Dymec, Palo Alto, Calif.

Product Engineering and Engineering

San Francisco—May 22

"Brazing Problems in Vacuum Tube Manufacturing," David K. Davis, Varian Associates, Palo Alto, Calif.

San Francisco—April 24

Panel Discussion: "Packaging," Bronson Balcer, LMSC; R. M. Bracamente, Bracamente Co.; H. Wood, Hewlett Packard Co.; T. E. Scatchard, Backman.

Space Electronics and Telemetry

Los Angeles—June 12

"Systems Design Concepts for Satellite Magnetic Core Memory,"

"Prediction Recording of Telemetry Data," Alex Egger; Space Tech. Labs., Los Angeles, Calif.; Finn Jargensen, Sr., Minnesota Mining and Manufacturing Co., Los Angeles, Calif.

San Francisco—June 19

"The Mariner *R* Telemetering System," Charles C. Kirsten, Jet Propulsion Lab., Pasadena, Calif.

Vehicular Communications

Detroit—May 23

"Inspection Tour of TV Control Center for Freeways," Alger F. Malo, Director of Streets and Traffic, City of Detroit.

Los Angeles—June 18

"Current Status of Noise Blanking Techniques as Applied to Vehicular Communication," Verne Stineman, Communication Products Dept., General Electric Co., Los Angeles, Calif.

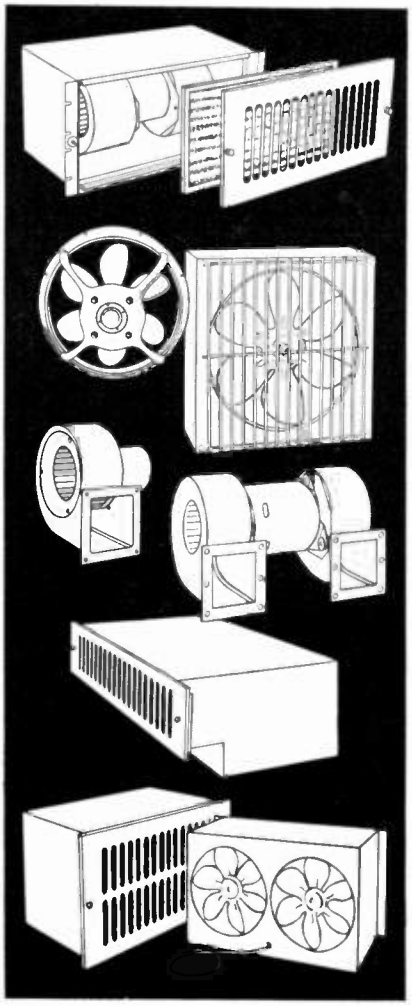
Twin Cities—April 30

"Nike Sites," Lieutenant Aslan, Twin Cities Air Defense Command.

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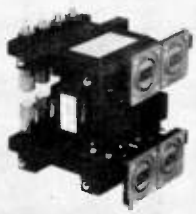
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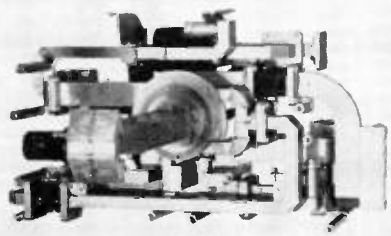
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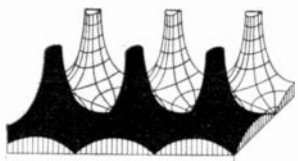
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WRH

October, 1962 *Vol. 50* *No. 10*

Proceedings of the IRE

Poles and Zeros



Another Golden Anniversary. This month marks the 50th anniversary of the licensing of amateur radio as set up and recognized by the Congressional Radio Act of 1912. No other activity in modern technology owes so much to "non-profit" amateur activity as does modern radio communications. The challenge and excitement of radio, without hope or concern for monetary rewards, have directed thousands of men and women toward intrinsically useful activities that have profoundly influenced our profession in particular and civilization in general.

In 1912, when amateurs were relegated to the "worthless" wave lengths below 200 meters, they obediently moved, and by cut and try and a goodly amount of science, proved the vagaries and merits of the "high frequencies." They pioneered portable and mobile communications and established the statistics for general acceptance of same. Single-sideband communication, "skip" measurements, and VHF communication capability have all been developed and proven out principally by amateur effort.

Radio in its infancy has been so exciting and all-absorbing that many have worked professionally by day and became amateurs by night. The value of amateur activity as reflected into professional creativity via this route has never been evaluated, but must be very great indeed. Furthermore, many of today's radio scientists began as amateurs and, motivated by the exciting, unforgettable experiences thereof, became effective professional researchers, developers, and administrators.

When World War II broke upon us, the world of amateur radio contributed its all, loyally, to the various governments of the world. The United States was indeed fortunate to have 40,000 men and women who knew the language and techniques which were to prove so decisive in the great conflict.

On this, the 50th anniversary of amateur radio, it would be interesting to know how many of the more than 100,000 IRE'ers have known the thrill, the ecstasy, the challenge, the fulfillment that could *only* have been provided by amateur radio. Happy anniversary!

Cedar Rapids and "The Career Brochure." Three years ago the Cedar Rapids Section adopted a concept of service by appointing a task force to develop a document that would assist youngsters in evaluating and planning careers in electronics. In 1960 the IRE Board of Directors supported publication of 10,000 copies which have subsequently been distributed. In the meantime, a plan has been instrumented for publication and distribution of career guidance information. The September issue of the STUDENT QUARTERLY AND EE DIGEST contains a Career Guidance section adapted from the Cedar Rapids publication. 40,000 reprints of the section will be run for distribution to high school guidance counselors and libraries. Would you like to assist in the project? Write Mr. Reed Crone at HQ for copies for high schools in your area.

Already Merged. In the discussions of merger, educators and students were the groups who stood virtually 100 per cent in favor. The competition since World War II between the AIEE and the IRE for student members, established a schism in our profession which served to confuse youth and to divide the faculty. The era of competition and confusion signalled its end in mid-September. At that time, the STUDENT QUARTERLY (IRE student magazine) and the EE DIGEST (AIEE student newspaper) joined together in a single, effective publication and thereby became the first positive, functional result of merger—88 pages, no advertising!

A Remarkable Slate. The 14-man merger committee designated by the Boards of IRE and AIEE is to be congratulated on the remarkable slate of officers which they presented last month. Dr. Ernst Weber, presidential nominee, is a Fellow of both societies (AIEE '34; IRE '51), V. P. nominee B. R. Teare, Jr., likewise (AIEE '42; IRE '51), and V. P. Bernie Oliver is an IRE Fellow ('54). The 25 nominees for the Board of Directors approximately balance the interests of the two societies and most have strong roots in both. As a direct consequence of the merger, many men, at all organizational levels, are having to step aside into supporting roles—there are only about one-half as many leadership posts in the merged society as in the two constituent societies. As the merger progresses, we will get many reassuring glimpses of the greatness of our colleagues as they subjugate immediate personal interests to the development of a strong and effective resultant society.

Credit Where Credit Is Due. The Editor acknowledges the following communication from Managing Editor E. K. Gannett: "In your August *Poles and Zeros* item on historical papers, you very graciously credited me with originating the suggestion for the series of historical papers. Actually, the suggestion orginated with Major General Francis F. Uhrhane, Commanding General of Fort Huachuca, Arizona, and a Senior Member of IRE, in a letter which he wrote to IRE sometime last fall. I don't know how this got by me, but in the hustle and bustle of inserting the merger news at deadline time, it did!"

The re-publication of Marconi's paper in the August issue has elicited a multitude of comments from members, which indicates that General Uhrhane's recommendation is one of the best and most fruitful suggestions we have received in a long time. Our apologies, General, and thanks.—T.F.J.



F. K. Willenbrock

Director, 1962–1963

F. Karl Willenbrock (S'46–A'51–M'55–SM'61) graduated from Brown University, Providence, R. I. in 1942, with a B.S. degree in electrical engineering. While still an undergraduate, he worked at the Bell Telephone Laboratories, Inc., New York, N. Y. After graduation he was employed at the Naval Ordnance Laboratory, Washington, D. C., where he worked with a research group concerned with underwater acoustics. He was commissioned as an Ensign in the U. S. Naval Reserve in 1943, and was on active duty during the remainder of World War II. Promoted to the rank of Lieutenant, he served as a specialist in electronic underwater ordnance equipment used in anti-submarine warfare.

At the termination of World War II, he began his graduate work at Harvard University, Cambridge, Mass. He received the M.S. degree in applied physics in 1947 and the Ph.D. degree in electron physics in 1950. His doctoral dissertation, which was concerned with the interaction of an electron beam with the oscillating electromagnetic field in a microwave cavity, was under the supervision of Professor E. L. Chaffee.

After completion of the Ph.D., he was appointed a Research Fellow and Lecturer in Applied Physics at Harvard. In 1955 he became Associate Director of the Laboratories of the Division of Engineering and Applied Physics. In 1960 he was appointed to his present position of Associate Dean and Director of Laboratories of Engineering and Applied Physics. In these capacities he is concerned with both the undergraduate and graduate instruction programs in these fields, and also with the pre-doctoral and post-doctoral research programs. His major areas of technical interest, in which he has published several articles, are microwave physics and solid-state electronics.

Dr. Willenbrock has been active as a consultant for both government and industry. At present he is a Senior Consultant and Acting Director of Research for the Sperry Rand Research Laboratory, Sudbury, Mass. In addition he is a Consultant to the Research Director of the Office of Naval Research, Washington, D. C. He is a member of Sigma Xi, Tau Beta Pi, the American Physical Society, and the American Society for Engineering Education. In 1959–1960 he was Chairman of the Boston Section of the IRE. He is currently Chairman of the Technical Program Committee of the Northeast Electronics Research and Development Meeting (NEREM).

Scanning the Issue

Popov and the Beginnings of Radiotelegraphy (Süsskind, p. 2036)—On June 1, 1894, in commemoration of the untimely death of Hertz at the age of 36, Dr. Oliver Lodge gave a lecture at the Royal Institution in London in which he described and demonstrated various methods of detecting electromagnetic waves which had been developed by Hertz and several successors. Lodge's lecture was widely reported, stimulated several important investigations, and was soon to have far-reaching consequences. For at least as an indirect result, within two years two men, Popov and Marconi, succeeded in transmitting intelligence by radio waves. The question of who was first has been with us in some degree ever since. However, it was not until Soviet authorities formally declared in 1945 that Popov was the "inventor of radio" that the question became a subject of world-wide controversy. Curiously enough, although a number of articles have since appeared which discuss various facets of the subject, a comprehensive examination and evaluation of all the available historical evidence is still lacking. It is especially appropriate that the PROCEEDINGS should present, in the golden anniversary year of the IRE, what is believed to be the most serious and thorough investigation of the origins of radio that has yet been published. As a document of historical research, it makes a major contribution to the history of the most important decade in radio. As a commentary, it brings fresh light on a topical controversy. But most important of all, it brings to a new generation a fuller appreciation of the achievements of all those who contributed to the birth of radio, and does so in a highly readable and interesting fashion.

Neuristor—A Novel Device and System Concept (Crane, p. 2048)—This paper introduces a new concept of major significance—an active transmission line along which signals can be propagated without attenuation by a process of continual regeneration. The action is analogous to that of a burning zone moving along a chemical fuse or an ionic discharge propagating along a nerve axon. A fuse does not recover after such propagation, and can be used only once. A neuristor line, however, like the axon, recovers after a brief interval following the passage of a signal. The fact that the region of the line immediately behind the signal goes temporarily "dead" during the recovery period leads to some novel and useful effects. For instance, two signals traveling towards one another will be annihilated upon collision because of the "dead" zones behind each. For the same reason, reflections cannot occur when the signal encounters discontinuities in the line. These and other unusual properties of neuristor lines extend their potentialities considerably beyond the prospect of using them merely as a means of interconnecting conventional devices and circuits. As the author shows, networks of such lines can perform logic and memory functions, opening the door to a new and versatile family of devices.

An Active Pulse Transmission Line Simulating Nerve Axon (Nagumo, *et al.*, p. 2061)—The preceding paper proposed a novel device, in the form of an active transmission line, in which the mode of signal propagation is somewhat analogous to that which occurs along a nerve axon. The present paper presents the physical realization of a similar type of line, using tunnel diodes distributed along the line. The result is a development which is extremely interesting not only because it simulates certain pulse transmission properties of the animal nerve axon but also because these same properties make for a highly reliable and useful electronic transmission system. This paper will therefore find a wide audience, including those in bionics, medical electronics, semiconductors, microcircuitry and data transmission.

The Analytic Signal Representation of Modulated Waveforms (Bedrosian, p. 2071)—In this day of multiphase communications (*e.g.*, stereo FM, color TV) it is appropriate to re-evaluate our methods of modulation systems analysis. The author of this paper finds a convenient and useful way of representing modulated waveforms which in essence is simply a formalized version of the familiar rotating vector or phasor technique. By comparing his method of analysis with the conventional method each step of the way, for each of the familiar types of modulation, he succeeds in throwing new light on the phasor concept. This method of analysis will be of wide interest to workers in circuit theory, telemetry, communication theory, systems development, and indeed to anyone who is involved in rotating concepts or multiphase concepts. Equally interesting are the results of the analysis which reveal the existence of a novel form of modulation—single sideband FM—which can be derived from a conventional phase-modulated signal by an additional amplitude modulation.

Observations of Satellite-Related Ionization Effects Between 1958 and 1960 (Kraus and Tiuri, p. 2076)—Late in 1957 the radio observatory at Ohio State University found that the 20 Mc transmissions of WWV often underwent large increases in signal strength at the time of passage of a satellite. The initial reports of this phenomenon were published in previous issues of the PROCEEDINGS. Since then, observations have continued at Ohio State and at other places. This paper brings together the results of these observations, which were carried on for a two-year period. In addition to providing substantiating evidence that a detectable amount of ionization is generated in the ionosphere by the passage of a satellite, the results bring to light some interesting details concerning various long-range and short-range effects which are produced. These effects will be of interest to all engineers concerned with satellite communications, radio propagation, and satellite tracking. They should also be important to those attempting to refine our knowledge of the ionosphere and of the fields surrounding the earth.

The Dielectric Microwave Resonator (Okaya and Barash, p. 2081)—Because of recent interest in quantum electronics, due largely to developments in microwave and optical masers, it becomes increasingly important to understand electromagnetic wave coupling with high dielectric constant resonators. This paper develops the theoretical background on mode computation and presents experimental verification in rutile crystals. Of particular interest is the fact that at microwave frequencies, rutile is found to have a Q as high as that of a metal-walled cavity. Thus, this work will be of value to those interested in using high-Q crystals as microwave filter components as well as those working on maser devices.

Generalized Resolution in Radar Systems (Urkowitz, *et al.*, p. 2093)—This paper provides an excellent discussion of, and contribution to, a very important topic in radar, namely, resolution. Previous work on determining the fundamental limitations of a radar system in resolving targets dealt with a two-dimensional problem, *i.e.*, targets differing in range and in range rate. This work has now been extended to four dimensions by the addition of azimuth and elevation as considerations. The result is a basic paper on practical radar design which will be of additional interest to those engaged in information theory and antenna design.

Scanning the Transactions appears on page 2153.

Popov and the Beginnings of Radiotelegraphy*

CHARLES SÜSSKIND†, SENIOR MEMBER, IRE

Summary—**Popov's rôle in the development of radiotelegraphy is evaluated in relation to the contributions of other pioneers, notably Lodge and Marconi. The official Soviet position on the matter is critically reviewed.**

SINCE 1945, Soviet authorities have vociferously claimed the title of "inventor of radio" for Aleksandr Stepanovich Popov (1859–1905; Fig. 1), bringing forward a great deal of historical evidence to support their claim. The task of critically examining that evidence has long been incumbent upon non-Soviet historians of technology, but has been hitherto tackled only in a very patchy way. The present contribution is the result of a careful study of the original Russian sources and other pertinent technical literature of the period, an examination of most of the material published in the Soviet Union afterwards, and an evaluation of some of the comments that the controversy has elicited in other countries. The author approached his task with an entirely open mind and drew his conclusions not in any effort to discredit Popov (who was one of the pioneers in the art of adapting Hertz's experiments to practical applications), but rather with a view to presenting an impartial account, free from the strongly partisan attitude characteristic of the material published on Popov by his fellow countrymen.



Fig. 1—A. S. Popov (1859–1905).

The Post-Hertzian Period

During the first five or six years after Hertz had proved the validity of Maxwell's wave theory of electromagnetism by a series of brilliant experiments published in 1888–1889, they were carried further and elaborated by a number of physicists. Ramsey has shown[1] that although many of these experiments were remarkably advanced for their time, they did little to bring technological application nearer. Most could be, in fact, classified under a heading that did not come into its own until half a century later: microwave optics. Typical of this group of investigations were those of Édouard Sarasin (1843–1917) and Lucien de La Rive (1834–1924) at Geneva,[2] Antonio Giorgio Garbasso (1871–1933) and Emil Aschkinass (1873–1909) at Berlin,[3] Jagadis Chunder Bose (1858–1937) at Calcutta,[4] and Augusto Righi (1850–1920) at Bologna.[5] The last had probably the most direct influence on the technological development of the subject, not only because it led to a second textbook[6] (devoted entirely to radiotelegraphy), but also because it was at Righi's lectures that young Guglielmo Marconi (1874–1937; Fig. 2) received the first proper grounding in the field that was to make him famous. But the single event that proved to be the most important technologically occurred in London on June 1, 1894, five months after the tragically premature death of Hertz at the age of 36. In commemoration, Oliver Joseph Lodge (1851–1940) gave a demonstration lecture at the Royal Institution that was to have far-reaching consequences.

* Received August 7, 1962.

† Department of Electrical Engineering, University of California, Berkeley, Calif.

[1] J. F. Ramsay, PROC. IRE, vol. 46, pp. 405–415; 1958.

[2] É. Sarasin and L. de La Rive, *Arch. Sci. Phys. Nat.*, vol. 29, pp. 358–393 and 441–470; 1893.

[3] A. G. Garbasso and E. Aschkinass, *Ann. Phys.* (3rd series), vol. 53, pp. 534–541; 1894.

[4] J. C. Bose, *Proc. Roy. Soc.* (*London*), vol. 59(A), pp. 160–167, 1895; vol. 60(A), pp. 167–168, 1896; vol. 63(A), pp. 146–155, 1898.

[5] A. Righi, "L'ottica delle oscillazioni elettriche," Nicola Zanichelli, Bologna, Italy; 1897.

[6] A. Righi and B. Dessau, "Telegrafia senza filo," Nicola Zanichelli, Bologna, Italy; 1903. (Published simultaneously in German, "Die Telegraphie ohne Draht," Friedrich Vieweg und Sohn, Braunschweig, Germany; 1903.)



Fig. 2—Guglielmo Marconi (1874–1937) in 1897.

Lodge's Lecture

Lodge, who had been appointed professor of physics at the newly established University of Liverpool in 1881, had also looked for the waves predicted by Maxwell, but along wires rather than in free space. In 1888 he had discovered[7] the standing waves produced by reflections —but the identity of free and guided waves was not properly appreciated until after Hertz's experiments, as a result of the work of Ernst Lecher (1856–1926)[8] and of Sarasin and de La Rive.[2] Moreover, Lodge had noticed that the two metal spheres of a shunting spark-gap transmission-line lightning protector of his own design tended to fuse together after the passage of a spark, but could be separated again by a gentle tap;[9] when Édouard Branly (1846–1940) discovered that the resistance of metal powders diminished substantially under the action of a nearby spark but could be likewise restored by tapping,[10] Lodge pointed out the similarity of the two phenomena,[11] which he put down to what he called "cohesion." (The name stuck, and the resulting device was called the *coherer*, though Branly would have preferred another name, *radioconductor*.[12])

Lodge's lecture, "The Work of Hertz and Some of His Successors," was widely reported,[13] reprinted verbatim in *The Electrician*,[14] and issued as part of a book that went into several editions.[15] When the British Association met at Oxford, Lodge gave another lecture, "Experiments Illustrating Clerk Maxwell's Theory of Light," in which he included essentially the same demonstration experiments.[16] He described various methods of detecting electromagnetic waves and demonstrated several, including an electroscope method first proposed by Ludwig Boltzmann (1844–1906),[17] coherers of his own design, and a Branly coherer.

Among his own coherers was the spark-gap coherer consisting of a pair of spheres, whose action was demonstrated in a remarkably well-designed experiment. The coherer was mounted in series with a battery and an electric bell. When a spark occurred in the vicinity, the spheres cohered together, the circuit was closed, and the bell began to ring and continued to ring until the spheres were separated by a gentle tap. Lodge found that if the bell was mounted on the same stand as the coherer, the mechanical vibration of the first stroke of the bell was usually sufficient to tap the spheres apart, so that each spark was signalled by a single stroke of the bell. Variants of this method of automatically "decohering" were to be widely employed in early apparatus. In addition, Lodge had been experimenting with resonance for several years[18] and had devised a sharply tuned resonant circuit that was a substantial advance over Hertz's relatively primitive arrangements.

But Lodge evidently made no attempt to transmit intelligence. He did not even attempt to measure the distance to which his equipment would remain effective. Referring to the two lectures in a paper on the history of the coherer written a few years later, he wrote: "In both cases signalling was easily carried on from a distance through walls and other obstacles, an emitter being outside and a galvanometer detector inside the room. Distance without obstacle was no difficulty in these experiments, only free distance is not very easy to get in a town, and stupidly enough no attempt was made to apply any but the feeblest power so as to test how far the disturbance could really be detected."[19] Still later, he admitted, "I was too busy with teaching work to take up telegraphic or any other development; nor had I the foresight to perceive, what has turned out to be, its

[7] O. J. Lodge, *Rept. Brit. Assoc. for the Advancement of Science*, p. 567; 1888.
[8] E. Lecher, *Sitzungsber. Akad. Wiss. Wien*, pt. 2a, vol. 99, pp. 340–361; 1890.
[9] O. J. Lodge, *J.I.E.E.*, vol. 19, pp. 346–379 and 382–410; 1890.
[10] É. Branly, *Comptes Rendus*, vol. 111, pp. 785–787, 1890; vol. 112, pp. 90–93, 1891.
[11] O. J. Lodge, *Phil. Mag.* (5th series), vol. 37, pp. 94–95; 1894.
[12] É. Branly, *Comptes Rendus*, vol. 125, pp. 939–942; 1897.
[13] See, for instance, *Nature*, vol. 50, pp. 133–139, 1894; *Engineering*, vol. 57, p. 751, 1894.
[14] O. J. Lodge, *Electrician*, vol. 33, pp. 153–155, 186–190, 204–205 ff; 1894.
[15] O. J. Lodge, "The Work of Hertz and His Successors," 'The Electrician' Printing and Publishing Co., London, England; 1894. A second edition appeared in 1898; the third and fourth, retitled "Signalling Through Space Without Wires," came out in 1900 and 1908, respectively. In its final form the book also contained reprints of several other papers and various correspondence.
[16] O. J. Lodge, *Engineering*, vol. 58, pp. 382–383; 1894.
[17] L. Boltzmann, *Ann. Phys.* (3rd series), vol. 40, pp. 399–400; 1890.
[18] O. J. Lodge, *Nature*, vol. 41, p. 368, 1890; *Proc. Roy. Soc. (London)*, vol. 50(A), pp. 2–39, 1891.
[19] O. J. Lodge, *Electrician*, vol. 40, pp. 87–91; 1897.

extraordinary importance to the navy, the merchant service, and indeed land and war service, too."[20]

A number of investigators were inspirited by Lodge's demonstrations. Alexander Muirhead (1878–1920) was among the first to foresee applications to telegraphy; he was to become Lodge's collaborator and his partner in a syndicate formed to exploit the resonant-tuning and other patents to which their work led. Henry Bradwardine Jackson (1855–1929), a British naval officer who subsequently rose to high rank (he was First Sea Lord of the Admiralty in 1915–1916 and became the first chairman of the Radio Research Board of Britain's Department of Scientific and Industrial Research in 1920), carried out secret trials of radiotelegraphic communications on and between ships in 1895–1896. Few details of these trials are generally available, but it may be safely assumed that they were on a high level of technical competence: Jackson was presently elected a Fellow of the Royal Society (in 1901) and knighted in 1906. His subsequent observations of variations in transmission with type of terrain were among the first quantitative propagation measurements ever published.[21]

In Germany, Adolf Carl Heinrich Slaby (1849–1913), professor of electrical engineering at the Institute of Technology in Charlottenburg, likewise started experimenting, although he did not (by his own admission[22]) begin to make much progress until after he had witnessed one of Marconi's demonstrations in 1897. And in Italy, Lodge's lecture gave a fresh impetus to the work of Righi, which led to the first experiments of Marconi.

Marconi's Earliest Work

Marconi had chanced upon an account of Hertz's experiments while vacationing in the summer of 1894. He began experimenting in the spring of 1895 on his father's estate at the Villa Grifone near Pontecchio, astutely taking advantage of the advances that had been made since Hertz's day. Marconi could hardly believe that no one had grasped the commercial possibilities of Hertzian waves, especially since the principal components of a workable system—the oscillator, the antenna, the resonator, the coherer, and the automatic decoherer—had already been invented and were presumably available to engineers much better qualified than a privately educated Italian youth barely out of his 'teens. "When I started my first experiments with Hertzian waves," he is cited as saying, "I could scarcely conceive it possible that their application to useful purposes could have escaped the notice of eminent scientists."[23]

In these first experiments, Marconi used an ordinary spark induction coil and home-made coherers of the Branly type. The spark gap was a four-ball arrangement of a type similar to one first proposed by Lodge[24] and afterwards improved by Righi, who had found that tarnishing of the balls owing to continuous sparking could be reduced if they were surrounded by a parchment bag containing a mixture of vaseline and vaseline oil. To turn the discharge on and off Marconi inserted a telegraph key in the primary circuit of the induction coil and was thus able to cause short or long trains of sparks to jump across the gap according to whether he held the key down for a short or a long time. He could soon detect these dots and dashes all the way across the room. By the summer of 1895 he had moved his equipment to the garden and had begun experimenting outdoors. In searching for methods of increasing the capacitance in the transmitter, he quickly hit upon the phenomenon that various observers had noticed before him: connecting one terminal of his transmitter to an elevated metallic object and the other to a grounded metal plate increased not only the capacitance, but also the distance over which the radiation could be detected.[25] This "grounded antenna," the elevated portion of which consisted of a piece of metal atop a pole, enabled him to signal across the entire length of the garden, and he found that there was a direct relationship between the height of the pole and the distance attainable. For a receiver, he used a similar arrangement with a Branly coherer inserted in the circuit.

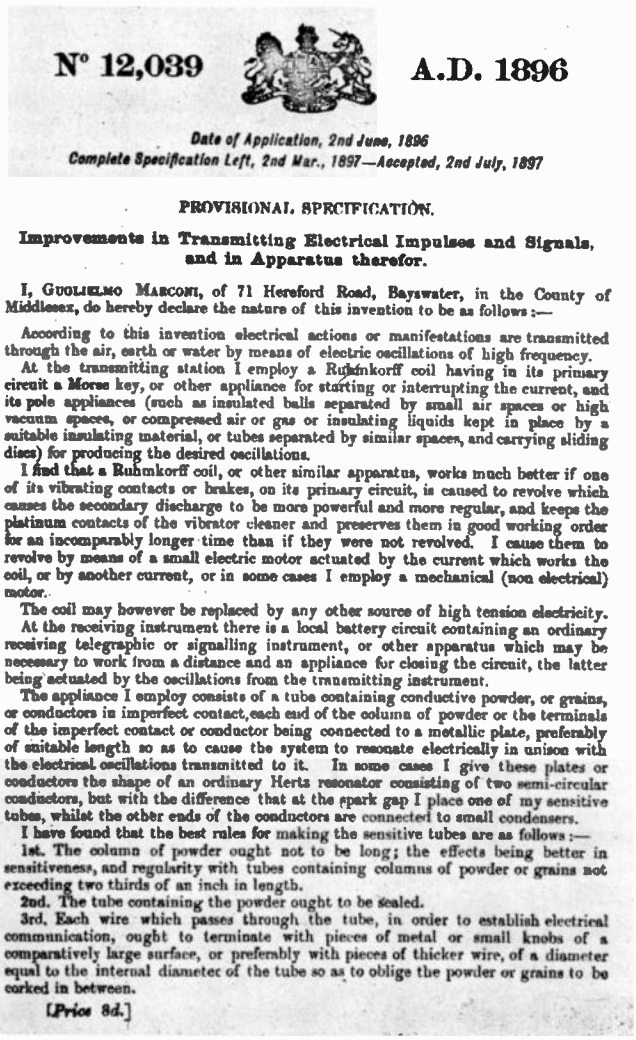
Marconi made important technical improvements in all the components of the system, notably in the design of the coherer. In series with the coherer he put a relay that fulfilled two functions; it actuated a tapper (which was simply an electric bell with the coherer itself taking the place of the bell gong) and worked a telegraphic printing instrument to record the received signals. In some experiments, he placed the spark gap of the transmitter along the focal line of a cylindrical parabolic metal reflector, much as Hertz had done. But even without the reflector, Marconi managed to receive signals at a distance of 2400 m, or about a mile and a half.

At that point, Marconi packed up his apparatus and

[20] O. J. Lodge, *Wireless Weekly*, vol. 2, pp. 305–307, 310, 337–339, and 412–413; 1923.
[21] H. B. Jackson, *Proc. Roy. Soc. (London)*, vol. 70(A), pp. 254–272; 1902.
[22] A. Slaby, "Die Funkentelegraphie," Leonhard Simion, Berlin, p. 18; 1897. See also *Century Magazine*, vol. 33, pp. 867–874; 1898.
[23] E. Hawks, "Pioneers of Wireless," Methuen & Co., London, England, pp. 221–222; 1927.
[24] O. J. Lodge, *Nature*, vol. 41, pp. 462–463; 1890.
[25] In the biography by Degna Paresce-Marconi, "My Father, Marconi" (McGraw-Hill Book Co., Inc., New York, N. Y.; 1962), it is stated that Marconi needed longer waves and in an effort to get them, replaced the outside balls of the Righi oscillator with two slabs of sheet iron. The author quotes what Marconi told his lifelong collaborator, Luigi Solari (1875–1957), years later: "By chance I held one of the metal slabs at a considerable height above the ground and set the other on the earth. With this arrangement the signals became so strong that they permitted me to increase the sending distance to a kilometer." Some commentators have singled out Marconi's use of the circuit in which the transmitting spark gap was connected between an elevated antenna and ground as his most important contribution. Although he doubtless arrived at this circuit independently, the efficacy of this arrangement (albeit not in connection with radiotelegraphy) had been observed even before Hertz's experiments by others, including Elihu Thomson (1853–1936) and Edwin James Houston (1847–1914), Thomas Alva Edison (1847–1931), Amos Emerson Dolbear (1837–1910), and David Edward Hughes (1830–1900). The contributions made by these precursors of Marconi (and of Hertz) are the subject of a forthcoming paper by the present author.

traveled to Britain, where he arrived in mid-February, 1896. He had chosen to exploit his invention there because Britain was a maritime nation whose dependence on marine communications seemed to offer the greatest scope for developing the commercial possibilities of his system. Another important reason for his decision to seek his fortune abroad was the fact that his mother, who was of British origin, had devoted herself to securing financial and other support for him. As a result of her efforts a meeting was arranged with William Henry Preece (1834–1913), the engineer-in-chief of the Government Telegraph Service, who very quickly decided to place the powerful support of the British Post Office at the young inventor's disposal. On June 2, 1896, Marconi disclosed the details of his system for the first time in print (Fig. 3): on that date he filed the specification for the first patent ever granted in radiotelegraphy.[26]

The subsequent events have been often recounted and need not concern us here. We need only add that the system was sufficiently advanced so that when a commercial enterprise based on it, the Wireless Telegraph and Signal Co. Ltd., was founded a year later (on July 20, 1897), Marconi and his British cousins had no difficulty in raising the initial capitalization of £100,000 among the hard-headed British businessmen. Not only that—there was, as one contemporary chronicle put it, "a great flutter in the dove-cotes of telegraphy, and holders of many millions of telegraph securities, and those interested in the allied industries, began to be alarmed for the safety of their property."[27]



N° 12,039 **A.D. 1896**

Date of Application, 2nd June, 1896
Complete Specification Left, 2nd Mar., 1897—Accepted, 2nd July, 1897

PROVISIONAL SPECIFICATION.

Improvements in Transmitting Electrical Impulses and Signals, and in Apparatus therefor.

I, GUGLIELMO MARCONI, of 71 Hereford Road, Bayswater, in the County of Middlesex, do hereby declare the nature of this invention to be as follows:—

According to this invention electrical actions or manifestations are transmitted through the air, earth or water by means of electric oscillations of high frequency.

At the transmitting station I employ a Ruhmkorff coil having in its primary circuit a Morse key, or other appliance for starting or interrupting the current, and its pole appliances (such as insulated balls separated by small air spaces or high vacuum spaces, or compressed air or gas or insulating liquids kept in place by a suitable insulating material, or tubes separated by similar spaces, and carrying sliding discs) for producing the desired oscillations.

I find that a Ruhmkorff coil, or other similar apparatus, works much better if one of its vibrating contacts or brakes, on its primary circuit, is caused to revolve which causes the secondary discharge to be more powerful and more regular, and keeps the platinum contacts of the vibrator cleaner and preserves them in good working order for an incomparably longer time than if they were not revolved. I cause them to revolve by means of a small electric motor actuated by the current which works the coil, or by another current, or in some cases I employ a mechanical (non electrical) motor.

The coil may however be replaced by any other source of high tension electricity.

At the receiving instrument there is a local battery circuit containing an ordinary receiving telegraphic or signalling instrument, or other apparatus which may be necessary to work from a distance and an appliance for closing the circuit, the latter being actuated by the oscillations from the transmitting instrument.

The appliance I employ consists of a tube containing conductive powder, or grains, or conductors in imperfect contact, each end of the column of powder or the terminals of the imperfect contact or conductor being connected to a metallic plate, preferably of suitable length so as to cause the system to resonate electrically in unison with the electrical oscillations transmitted to it. In some cases I give these plates or conductors the shape of an ordinary Herts resonator consisting of two semi-circular conductors, but with the difference that at the spark gap I place one of my sensitive tubes, whilst the other ends of the conductors are connected to small condensers.

I have found that the best rules for making the sensitive tubes are as follows:—

1st. The column of powder ought not to be long; the effects being better in sensitiveness, and regularity with tubes containing columns of powder or grains not exceeding two thirds of an inch in length.

2nd. The tube containing the powder ought to be sealed.

3rd. Each wire which passes through the tube, in order to establish electrical communication, ought to terminate with pieces of metal or small knobs of a comparatively large surface, or preferably with pieces of thicker wire, of a diameter equal to the internal diametec of the tube so as to oblige the powder or grains to be corked in between.

[*Price* 8*d.*]

Fig. 3—Titlepage of Marconi's first patent specification[26] for "transmitting electrical impulses and signals." Transmitting station is described first, then receiving instrument.

Popov's Contribution

Popov, a physics instructor at the Russian Navy's Torpedo School in Kronstadt near St. Petersburg, was still another investigator who was inspired by Lodge's 1894 lecture. He experimented with Branly coherers, set up a receiver with a protruding wire in 1895, and read a paper about it, "On the Relation of Metallic Powders to Electric Oscillations," at a meeting of the Russian Physico-Chemical Society on April 25 (May 7, New Style), 1895.[28] Popov had seen an account of Lodge's lecture and set up a circuit very similar to that of Lodge, with the coherer tube placed above the hammer of an electric bell. By means of this apparatus, Popov could register electric disturbances, including atmospheric ones, and in July, 1895, a similar instrument with an ink recorder was in fact installed at the meteorological observatory of the Institute of Forestry in St. Petersburg. The first publication[28] was merely a one-paragraph summary of the minutes of the meeting, but a more complete and extensively up-dated account of his "Apparatus for the Detection and Recording of Electrical Oscillations" appeared in January, 1896 (Fig. 4), which ended with the words: "In conclusion I may express the hope that my apparatus, when further perfected, may be used for the transmission of signals over a distance with the help of rapid electric oscillations, as soon as a source of such oscillations possessing sufficient energy will be discovered."[29]

On March 12 (24, N.S.), 1896, Popov gave a demonstration before the Physico-Chemical Society of which no verbatim record survives. The demonstration must have been quite short (it was one of nine items on the agenda) and evidently utilized largely the same apparatus as Popov's previous experiments. The minutes say, in their entirety, "A. S. Popov shows instruments for the lecture demonstration of the experiments of Hertz. A description of their design is already in the Zh. R. F.-Kh. Obshchestva."[30]

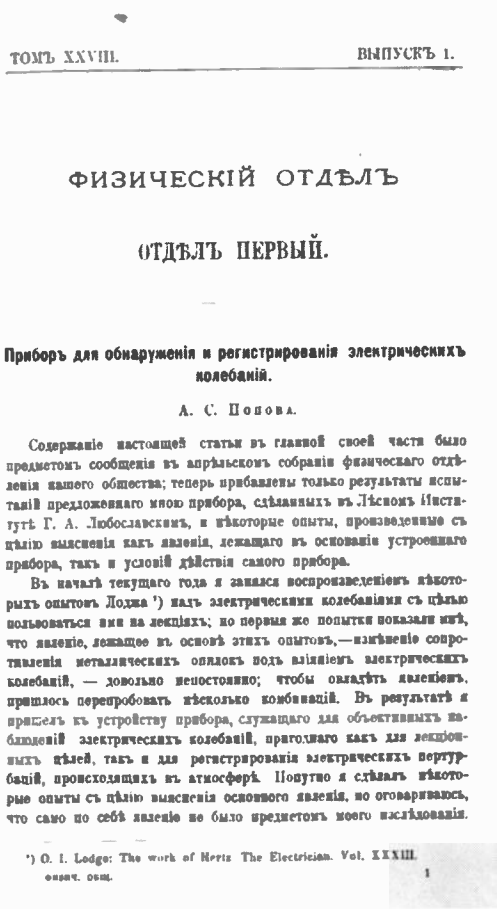
[26] British Patent No. 12,039; 1896. Also U. S. Patent No. 586,193; July 13, 1897.

[27] J. J. Fahie, "A History of Wireless Telegraphy," William Blackwood and Sons, Edinburgh and London; 1899.

[28] *Zh. Russ. Fiz.-Khim. Obshchestva* (Physics, pt. 1), vol. 27, pp. 259–260; 1895.

[29] A. S. Popov, *Zh. Russ. Fiz.-Khim. Obshchestva* (Physics, pt. 1), vol. 28, pp. 1–14; 1896.

[30] *Zh. Russ. Fiz.-Khim. Obshchestva* (Physics, pt. 1), vol. 28, pp. 121–124; 1896.



ТОМЪ XXVIII. ВЫПУСКЪ 1.

ФИЗИЧЕСКІЙ ОТДѢЛЪ

ОТДѢЛЪ ПЕРВЫЙ.

Приборъ для обнаруженія и регистрированія электрическихъ колебаній.

А. С. Попова.

Содержаніе настоящей статьи въ главной своей части было предметомъ сообщенія въ апрѣльскомъ собраніи физическаго отдѣленія нашего общества; теперь прибавлены только результаты испытаній предложеннаго мною прибора, сдѣланныхъ въ Лѣсномъ Институтѣ Г. А. Любославскимъ, и нѣкоторые опыты, произведенные съ цѣлію выясненія какъ явленія, лежащаго въ основаніи устроеннаго прибора, такъ и условій дѣйствія самого прибора.

Въ началѣ текущаго года я занялся воспроизведеніемъ нѣкоторыхъ опытовъ Лоджа [1]) надъ электрическими колебаніями съ цѣлью пользоваться ими на лекціяхъ; но первыя же попытки показали мнѣ, что явленіе, лежащее въ основѣ этихъ опытовъ,—измѣненіе сопротивленія металлическихъ опилокъ подъ вліяніемъ электрическихъ колебаній, — довольно непостоянно; чтобы овладѣть явленіемъ, пришлось перепробовать нѣсколько комбинацій. Въ результатѣ я пришелъ къ устройству прибора, служащаго для объективныхъ наблюденій электрическихъ колебаній, пригоднаго какъ для лекціонныхъ цѣлей, такъ и для регистрированія электрическихъ возмущеній, происходящихъ въ атмосферѣ. Попутно я сдѣлалъ нѣкоторые опыты съ цѣлію выясненія основного явленія, но оговариваюсь, что само по себѣ явленіе не было предметомъ моего изслѣдованія.

[1]) O. J. Lodge: The work of Hertz. The Electrician. Vol. XXXIII. опечат. обш.

Fig. 4—First page of Popov's paper on his "apparatus for the detection and recording of electrical oscillations."[29] First reference is to Lodge's lecture.[14]

After that, other matters apparently intervened. "During the academic year 1895–1896," wrote one of his colleagues afterwards, "A. S. Popov, without ceasing to make further improvements on his storm indicator, tested its operation in collaboration with G. A. Luboslavsky at the meteorological station of the Institute of Forestry; at the same time, in the Officers' Course at the Torpedo School, he becomes interested in setting up the experiments on Röntgen rays, which had just been discovered, and he lays aside for a time the realization of his ideas on the further development of the practical application of his storm detector and postpones for the moment his planned investigation of the optimum operating conditions of this new receiver."[31] He was indeeed working on X rays: at a meeting on February 13 (25, N.S.), 1896, he is recorded in the minutes as participating in a discussion of this subject and mentioning the results he had obtained in some X-ray experiments.[32]

As he had done every summer since 1889, Popov spent the summer vacation of 1896 working as the electrical engineer in charge of the power plant at the annual fair in Nizhni Novgorod. He had installed one of his storm detectors at that plant and had also shown it at the fair, where it earned a prize. It was there that he read the first accounts of Marconi's demonstrations in the newspapers in September, 1896; the news, one of his colleagues recalled later, "suddenly startled him."[33] (The Russian phrase, *srazu vstrepenuli ego*, is rendered in the French version of the journal as *le tirèrent brusquement de sa torpeur*, or "shook him brusquely out of his torpor.") He returned to Kronstadt, resumed his experiments, and presently wrote to a local newspaper pointing out that Marconi's receiver, details of which had not yet appeared in the press, was very likely quite similar to his.[34]

In the same year, Popov wrote a letter to the Paris instrument maker, Eugène Ducretet (1844–1908), a draft of which survived,[35] evidently in reply to a request for information that would establish whether Popov had anticipated Marconi: "I have no other printed papers at my disposal that can prove my participation in the practical solution of the problem of wireless telegraphy, other than the article already known to you. Nevertheless, I regard that article as sufficient to prove the identity of the constituent parts and their disposition in my apparatus with those in the receiving station of Mr. Marconi." It should be noted that Popov referred only to Marconi's *receiving* station. He went on to describe a demonstration (made in January, 1896, before the Kronstadt section of the Russian Technical Society) of which there is no published account, and in which he apparently used an improved version of his original receiver; and he added: "In March I demonstrated an apparatus for optical experiments with electromagnetic rays: reflection, refraction, effect of a grating and rotation of the plane of polarization by grained wood. In September 1896 the first notice of Mr. Marconi's experiments appeared in the daily press, but the nature of the apparatus remained secret and the technical journals lost themselves in conjectures about the new discovery. I then published a letter in a local newspaper in which, having referred to my apparatus, I showed that among the recordings of storms made by my apparatus there are some caused by discharges that occurred at least 30 kilometers away, that signaling by artificial discharges up to a mile is possible, and that, in all probability, Mr. Marconi's apparatus is identical to mine. This letter was published in the newspaper 'Kotlin' in October 1896." (That is not the same date as cited above, January 8, 1897;[34] his memory has evidently played him false. The latter date is taken from a biobibliography compiled much later.[36])

The date of Popov's letter to Ducretet is not known with certainty; the above draft is described as dated

[31] N. N. Georgievsky, *Elektrichestvo*, No. 4, pp. 211–215; 1925.
[32] *Zh. Russ. Fiz.-Khim. Obshchestva* (Physics, pt. 1), vol. 28, pp. 88–90; 1896.
[33] Ref. 31, p. 214.
[34] A. S. Popov, *Kotlin*, No. 5, p. 2; January 8, 1897. (Kotlin is the island on which the Kronstadt naval base is located.)
[35] Ref. 31, pp. 214–215.
[36] A. M. Lukomskaia, "A. S. Popov," Izdatelstvo Akademii Nauk SSSR, Moscow, U.S.S.R., p. 40; 1951.

"toward the end of 1897." In any event, Ducretet evidently made use of the information. At a scientific meeting in Paris on November 19, 1897, he exhibited a group of instruments made by his firm in Paris on the basis of the experimental apparatus employed by Hertz, Righi, Sarasin and de La Rive, Bose, Branly, and Popov. The minutes state that "M. Ducretet described the apparatus that M. Popoff had constructed in 1895 and employed in St. Petersburg for receiving and registering electric waves in proportion to their presence," and that Ducretet had modified the circuits to obtain greater sensitivity.[37] There is an oblique reference to Marconi: "M. Marconi was doubtless unaware of these endeavors when he employed the known components characterizing the wireless transmission that he realizes with them."

On April 16, 1898, Ducretet addressed the annual meeting of the French Physical Society "On Hertzian Wireless Telegraphy with the Radioconductor of M. Branly and the Circuits of MM. A. Popoff and E. Ducretet."[38] He had made further improvements in the several components of his experimental setup, all of which were exhibited with considerable pride. He described Popov's original receiver and paid a handsome tribute to Popov, who had not yet taken out any patent:[39] "The apparatus described and realized in 1895, by Professor A. Popoff, was constituted as above: *he utilized it to record electric waves produced by atmospheric disturbances and to transmit telegraphic signals recorded on a receiver over a great distance.* From 1895, M. Popoff, in his publications and his presentations before Russian scientific societies, has shown that his apparatus could be practically employed by the Navy for the *transmission of signals* over a great distance." (The last sentence had also appeared in connection with a radiotelegraphic receiver that Ducretet had described in a communication sent to the Academy of Sciences two months earlier.[40] He followed it with a couple of further notes about a recorder of atmospheric disturbances[41] and on some transmissions between the Eiffel Tower and the Panthéon.[42]) Ducretet contrived not to mention Marconi (who had patented a similar system nearly two years earlier[26]) even once, but ended his lecture on a patriotic note instead: "I hope, gentlemen, to have interested you and to have demonstrated to you that our French scientific industry, as far as I am concerned, *is not tributary* to foreign industry."

The manner in which Marconi's work had been pointedly ignored by Ducretet proved to be too much even for his French colleagues, one of whom, André Eugène Blondel (1863–1938), wrote indignantly on December 2, 1898: "The Physical Society, by virtue of its purely scientific and impartial character, ought not, I believe, let it be believed by its complete silence that it accepts as equitable the account of incomplete facts regarding the history of wireless telegraphy, and according to which it would appear that there had been no development between the works of M. Branly (1890–1891) and of M. Popoff (1895–1896) . . . . The facts bring out the fundamental rôle that Lodge and Marconi had played in the elaboration of wireless telegraphy . . . .

"Popoff, in 1895–1896, had replaced Lodge's receiver by a more sensitive tube, connected to an antenna in place of a circular exciter; he had achieved the recording of atmospheric discharges with an automatic tapper, and indicated that he would be able to transmit and receive signals if he had a sufficiently powerful oscillator at his disposal. But what he really lacked was a sufficiently sensitive tube of filings.

"Marconi is the first to have produced a tube of this kind . . . . He was able, without inventing any other new device, to enter the promised land glimpsed by his predecessors and to attain prodigious transmission distances that are reckoned by dozens of kilometers. . . . "[43]

Blondel's letter, which concluded with a ringing tribute to Marconi, elicited a reply from Branly and another from Ducretet. Branly thought that wireless telegraphy "resulted from" the experiments of Popov because he had developed Branly's idea of making the spark current pass along a metal rod, but he concluded that he "does not contest in any way the great interest of Marconi's experiments."[43] Popov was not mentioned in the note from Ducretet;[43] but when, at the next meeting of the Society (on December 16, 1898), he demonstrated a portable receiver that had been used at the Panthéon to receive transmissions from the Eiffel Tower, 4 km away, he reiterated that his receiver "derives from that of M. Popoff."[44]

Popov himself had in the meantime written to a British periodical in connection with an article about the coherer that had appeared in its pages, drawing the editor's attention to his own article[29] of January, 1896, and "citing" some extracts in his own translation.[45] He rendered the final paragraph of his paper as follows: "'In conclusion, I can express my hope that my apparatus will be applied for signalling on great distances by electric vibrations of high frequency, as soon as there will be invented a more powerful generator of such vibrations. From July, 1895, until now my apparatus has worked very well as a lightning recorder, which can be seen on the photograph (Fig. 3) of various records of thunder-storms made by this apparatus during the last summer . . . By using in the coherer tube a steel bead

[37] *Séances Soc. Fran. Phys.* (Résumés), pp. 65–67; 1897.
[38] E. Ducretet, *Séances Soc. Fran. Phys.*, pp. 51–61; 1898.
[39] Popov applied for a Russian patent on an improved coherer on July 14, 1899; the patent (No. 6066) was granted on November 30, 1901. The device was also patented in France (No. 296,534, January 22, 1900), Britain (No. 2797, February 12, 1900), and Sweden (No. 14,189, April 4, 1900). Popov obtained no patent for his receiver.
[40] E. Ducretet, *Comptes Rendus*, vol. 126, pp. 1266–1268; 1898.
[41] *Ibid.*, p. 1743.
[42] E. Ducretet, *Comptes Rendus*, vol. 127, pp. 713–716; 1898.
[43] *Séances Soc. Fran. Phys.* (Résumés), pp. 77–79; 1898.
[44] *Ibid.*, pp. 80–81.
[45] A. S. Popov, *Electrician*, vol. 40, p. 235; 1897

instead of iron filings I receive a good coherer, by which I can detect electromagnetic waves at a distance of 1 kilometer, if I work with Hertz's vibrator with 30cm. spheres and with the ordinary Siemens-Halske relay. With Bjerknes's vibrator of 90cm. diameter, and a more sensitive relay, I reach 5 kilometers of good working without exhausting the tube, and without any other resonance than by using the arrangement in my registering apparatus.' " Popov then concluded his letter to the editor with the following statement: "From the foregoing remarks may be inferred that the arrangement of Marconi's receiver is a reproduction of my lightning recorder."

Popov obviously did not mean that Marconi had copied his circuit, only that the two circuits were nearly identical. But the letter is remarkable for other reasons. First, the qualifying phrase that appeared in the original version in the sentence describing the hopes he had for his apparatus, "when further perfected" (*pri dalneishem usovershenstvovanii ego*), is missing. Second, only the first sentence of the cited extract appeared in the original paper, which contained no Fig. 3. It is thus quite evident that Popov was again adding some current information to his original paper, namely a description of some more recent results that was included either inadvertently or deliberately (and complete with ellipsis marks) as part of the extract from the earlier paper. (The additional information did not appear in an independent translation published subsequently in another British periodical.[46])

The protests helped to establish the priority of Popov's publication relating to the design of a practical receiver and recorder of electromagnetic disturbances. He had up to then published nothing about any work with a radio transmitter.

Popov continued to be active in the early development of radiotelegraphy, but it was evidently uphill work in the Russia of the Tsars. According to a Soviet source, "a shop established by Popov in 1900 at the Kronstadt port for the repair and manufacture of wireless-telegraphy apparatus had neither adequate equipment nor sufficient personnel to supply the Russian Navy with radio stations. On the eve of the Russo-Japanese War of 1904–1905, the Naval command was forced hurriedly to supply its ships with German-made radio stations."[47] This apparatus was derived from the designs of Marconi, whose merits were well appreciated in Russian official circles: in 1902, he had been made a Knight of the Order of St. Anne of Russia.

The Popov Legend

The contention that Popov originated radiotelegraphy got off to a good start during his own lifetime, as exemplified by the entries on "Popov" and on "telegraphy" in a Russian encyclopedia of the period.[48] After his death, the Russian Physical Society appointed a special committee to establish the circumstances of his priority over Marconi. The committee made a thorough search of the literature and even solicited opinions from abroad, but the two responses that were received (and published in the original languages) must have been disappointing. Branly sent a series of extracts from his publications showing that he had used vestigial antennas both at his transmitters and his receivers, and Lodge wrote a noncommittal letter that was in part reproduced as follows: "How far Popoff applied his device to actual telegraphic signalling of [an] intelligible message You will know better than I do . . . I conjecture that Popoff may have been one of the pioneers who applied the method to ship signaling of some rough kind at an early date."[49]

On the thirtieth anniversary of Popov's 1895 lecture, there were commemorative meetings in Leningrad and in Moscow; the Soviet journal *Elektrichestvo* dedicated an entire issue (published in Russian and French) to Popov, from which several of the above quotations have been taken.[31,33,35] At the Leningrad meeting A. S. Holtzmann, chief of the central authority of the electrical industry, presented a report on the relation between radioelectric science and industrial technology, in which he declared that Marconi's victory was a victory of industrialized Europe over ignorant Russia.[50]

But the curious campaign to create a full-blown Popov legend by government ukase did not really get under way until 1945, the fiftieth anniversary of Popov's first lecture, when the Soviet authorities formally declared that he was the "inventor of radio." An elaborate celebration was held in Moscow, characterized by what one foreign observer, the Australian botanist Sir Eric Ashby, has described as "shrill festivities."[51] Several books were published about Popov, all in Russian. The most elaborate is a two-volume history of radio comprising 1) an historical account of the work of Popov's predecessors and 2) a collection of papers, letters, and documents pertaining to the question of his priority.[52] A similar (but less complete) collection was published simultaneously.[53] There was a fictionalized popular biography[54] and an account written from the

[46] A. S. Popov, *Elec. Rev.*, vol. 47, pp. 845–846 and 882–883; 1900.
[47] B. A. Vvedensky, Ed., "Great Soviet Encyclopedia," Gosudarstvennoe Nauchnoe Izdatelstvo 'BSE,' Moscow, U.S.S.R., vol. 34, p. 159; 1955.
[48] I. E. Andreevsky, Ed., "Encyclopedic Dictionary," F. A. Brockhaus, Leipzig, Germany, and I. A. Efron, St. Petersburg, Russia, vol. 24, p. 558; 1898; vol. 32, pp. 777–790; 1901.
[49] *Zh. Russ. Fiz.-Khim. Obshchestva* (Physics, pt. 1), vol. 41, pp. 63–72; 1909.
[50] *Pravda;* May 8, 1925.
[51] E. Ashby, "Scientist in Russia," Penguin Books, Harmondsworth, England (Penguin Book No. A-186), pp. 198–200; 1947.
[52] C. M. Rytov and L. I. Mandelshtam, Eds., "Fifty Years of Radio," Izdatelstvo Akademii Nauk SSSR, Moscow and Leningrad, U.S.S.R., vol. 1 ("From the Prehistory of Radio"), 1948; vol. 2 ("The Invention of Radio by A. S. Popov"), 1945.
[53] G. I. Golovin and R. I. Karlina, Eds., "A. S. Popov: Collection of Documents," Leningradskoe Gazeto-Knizhnoe Izdatelstvo, Leningrad, U.S.S.R.; 1945.
[54] G. I. Golovin, "A. S. Popov, Inventor of Radio: Life and Work," Gosudarstvennoe Izdatelstvo Literatury, Moscow, U.S.S.R.; 1945.

Navy's viewpoint.[55] A. I. Berg and M. I. Radovsky co-authored two books[56,57] on Popov in 1945 and 1948, and the Soviet Academy of Sciences published a Popov lecture by the editor-in-chief of the Great Soviet Encyclopedia in 1948.[58] There is also a booklet of reminiscences by Popov's assistant,[59] a later book of recollections by several contemporaries,[60] and the previously mentioned biobibliography.[36] May 7, the date of Popov's 1895 lecture according to the new-style calendar, was declared "Radio Day," to be celebrated annually. Most of the world had been content to consider Marconi as the father of radiotelegraphy, since he was the first to employ (and to publish a description of) a transmitter and receiver specifically designed for the transmission of intelligence. Soviet official circles now began to press the claim of Popov's priority with considerable conviction, arising in the first instance from the indubitable fact that Popov's first receiver publications antedated those of Marconi.

One cannot very well consider the controversy from the viewpoint of patent law, which varies from country to country. In Britain, for instance, the date of application for a patent is the paramount consideration; in the U.S.A., the date of conception. In the Soviet Union, the date of the first publication (*i.e.*, public disclosure) is the critical date. An international convention governs patent-law relations among various countries, but the Soviet Union is not a signatory; neither was prerevolutionary Russia. Nor is it very relevant to note that the mainstream of technological development in radio communications proceeded largely from the efforts of Marconi and his followers and scarcely at all from those of Popov, not even (as we have seen) in Russia. The argument turns principally on whether Popov used his instrument merely to register lightning flashes and their man-made equivalents, or whether he actually did publish, before mid-1896, a description of his use of his instrument for the transmission of intelligence.

The record shows that he did not. There is no mention of radiotelegraphic experiments in the minutes of the 1895 meeting,[28] nor in the terse minutes of the 1896 demonstration.[30] There is no mention of radiotelegraphic experiments in the 15-page paper[29] that appeared in January, 1896, nor in the artfully updated excerpts[45] that appeared in *The Electrician* in 1897, nor yet in the translation[46] that the *Electrical Review* printed in 1900. What, then, was the basis for the official claim that was made on Popov's behalf?

Signaling Demonstrated by Popov?

The claim that Popov had demonstrated the use of electromagnetic waves for signaling was first made in 1925. Its appearance was attended by some confusion. Victor S. Gabel, an official in the Soviet Bureau of Weights and Measures, supplied a short note in connection with the forthcoming commemoration of the 30th anniversary of Popov's lecture of May 7, 1895, to a British periodical which duly printed the note[61] on May 6, 1925. It stated that during his lecture Popov performed "the first successful transmission . . . of a communication by means of electromagnetic waves . . . As each signal in Morse code was received, the President of the Society wrote the corresponding letter on the blackboard, and the enthusiasm of the audience was very great when the words 'Heinrich Hertz' were eventually spelt out."

The editor wrote to Gabel for details. Gabel in turn wrote to three of Popov's erstwhile colleagues. Excerpts from their replies were published in a Soviet journal.[62] The first, Orest Danilovich Khvolson (1852–1934), was not sure of the date, but could remember the circumstances clearly: "The transmission took place in such a manner that the letters were transmitted in the Morse alphabet and the signs were moreover clearly audible. The president of the Phys. Society, Prof. F. F. Petrushevsky, stood at the blackboard, holding a paper with the key to the Morse alphabet and a piece of chalk. After each sign was transmitted he glanced at the paper and then wrote the corresponding letter on the blackboard. Gradually there appeared on the blackboard the words *Heinrich Hertz*, moreover in Latin characters. It is difficult to describe the joy of the many members of the audience, and the ovation given to A.S. Popov, when these two words were written down. The meeting took place at the beginning of 1896, but I cannot establish the exact date."

The second correspondent, Vladimir Konstantinovich Lebedinsky (1868–1937), identified the occasion on which this transmission took place as the March 12 (24, N.S.), 1896, demonstration. He noted that he had kept the tape on which the historic message had been recorded and had saved it until his library perished in Riga in 1918–1919; and he explained why the minutes of the meeting described the event in two sentences without mentioning the transmission: "Such a terseness in the wording of the minutes, which gives an altogether inadequate description of the significance and high importance of the paper, may be explained by the fact that in 1896 the work of A. S. [Popov] was under the

[55] S. Kudriavcev-Skaif, "The Russian Navy—Cradle of Radio," Voenno-Morskoe Izdatelstvo NKVMF SSSR, Moscow and Leningrad, U.S.S.R.; 1945.

[56] A. I. Berg and M. I. Radovsky, "Aleksandr Stepanovich Popov: To the 50th Anniversary of Radio," Gosudarstvennoe Energeticheskoe Izdatelstvo, Moscow and Leningrad; U.S.S.R.; 1945.

[57] A. I. Berg and M. I. Radovsky, "The Inventor of Radio—A. S. Popov," Gosudarstvennoe Energeticheskoe Izdatelstvo, Moscow and Leningrad, U.S.S.R.; 1948.

[58] B. A. Vvedensky, "A. S. Popov: Inventor of Radio," Izdatelstvo Akademii Nauk SSSR, Moscow and Leningrad; U.S.S.R.; 1948.

[59] P. N. Rybkin, "Ten Years with the Inventor of Radio," Sbiazizdat, Moscow, U.S.S.R.; 1945.

[60] M. I. Radovsky and K. K. Baumgart, Eds., "Aleksandr Stepanovich Popov, in Vignettes and Reminiscences of Contemporaries," Izdatelstvo Akademii Nauk SSSR, Moscow and Leningrade, U.S.S.R.; 1958.

[61] *Wireless World*, vol. 16, p. 410; 1925.

[62] V. S. Gabel, *Telegrafia i Telefonia bez Provodov*, vol. 7, pp. 247–249; 1926. (Cited in Ref. 52, vol. 2, pp. 270–272.)

control of the Ministry of the Navy and could not be made public."

This theme was developed by the third correspondent, V. V. Skobelcyn, who recalled questioning the Physical Society's secretary, Aleksandr Lvovich Gershun (1868–1915), why the demonstration had been so inadequately described in the Society's journal. His reply was that the wording was an exact reproduction of a note written by Popov himself and submitted with the request that "nothing should be changed and nothing should be added." Neither Gershun nor Skobelcyn understood the reason for Popov's request.

The information contained in the three letters now led Gabel to send a correction to London, which appeared on February 24, 1926: "I regret that on account of some unfortunate misunderstanding my information regarding the transmission by means of Hertzian waves of the words 'Heinrich Hertz' in the Morse alphabet at the meeting of May 7th, 1895, was incorrect. On that occasion Prof. A. S. Popoff demonstrated only with a receiving set consisting of a Branly coherer, a relay, and an electric bell which automatically tapped the coherer. All the apparatus was enclosed in a metallic case, out of which protruded a small antenna. By taking away the antenna Prof. Popoff demonstrated the decrease of reception. A receiving set was operated from a Hertz vibrator.

"The transmission of words in the Morse alphabet was demonstrated by him on March 24th, 1896; also at a meeting of the Russian Physico-Chemical Society. Transmission was conducted at a distance of about 250 metres (with two antennas, one for receiving and one for transmitting)."[63]

There the matter rested, except for an occasional reminder in the correspondence columns of technical journals,[64] until the 1945 anniversary, when the full apparatus of Soviet government propaganda was brought to bear on the problem of establishing the simple and easily grasped idea that Popov "invented radio" on May 7, 1895. In view of the admission that all he had demonstrated on that occasion was a receiver of waves originating from electromagnetic disturbances, the selection of that particular date seems to be at first quite unjustified. It would have been more logical if the date of the "Heinrich Hertz" demonstration, March 24, 1896, had been selected—except that there was no printed record of it. Faced with the choice of backing a well-documented weak claim or a later, undocumented, but stronger one, the authorities evidently decided to press the earlier one. They may have been influenced in that decision by the fact that the 1925 anniversary had likewise harked back to the earlier date. Moreover, Marconi was known to have experimented with transmissions of signals at least as early as June, 1895, and although he published nothing until his patent application of June 2, 1896, he had incontestably brought his apparatus to a degree of perfection that justified his coming to London in mid-February, 1896 (*i.e.*, over a month before the "Heinrich Hertz" demonstration), to apply for a patent and to demonstrate his system. But just for good measure, the Soviet authorities made a determined effort to document the "Heinrich Hertz" demonstration better.

For this purpose, Popov's aged collaborator, Piotr Nikolaevich Rybkin (1864–1948), an octogenarian and an Order of Lenin holder, was recruited to publish a booklet of reminiscences,[59] which was quite possibly ghost written. (Aksel Ivanovich Berg, the coauthor of two other books on Popov,[56,57] is listed as "editor.") The contention that the naval authorities had stopped Popov from publicizing his experiments was given a firmer backing. "Before the impending lecture, the chief of the Torpedo School, Commander V. F. Vasilev, came to see us in the Physics Department," Rybkin now remembered. "It is true that he used to come to the Physics Department frequently even before that. But this time Vasilev was particularly courteous and thoroughly interviewed Popov and myself about the experiments on wireless telegraphy. In conclusion the Commander cautioned Aleksandr Stepanovich [Popov] that it behooved him to regard his work very cautiously, since it had military significance."

Another eyewitness, Academician Vladimir Fiodorovich Mitkevich (1872–1951), recounted the circumstances of the demonstration and recalled that "among those attending this meeting were representatives of the Admiralty and the most distinguished Russian physicist-electricians of that time: O. D. Khvolson, I. I. Borgman, A. I. Sadovsky, V. K. Lebedinsky, M. A. Shatelen, A. A. [*sic*] Gershun, G. A. Luboslavsky, N. N. Georgievsky, N. A. Smirnov, V. V. Skobelcyn, N. A. Bulgakov, N. G. Egorov, and F. F. Petrushevsky."[65]

Foreign Reactions

What were the reactions of non-Soviet experts? If they had any previous opinions of the importance of Popov's contribution, they were probably based on a statement made in a very popular early radiotelecommunications textbook by John Ambrose Fleming (1849–1940) who says, in citing excerpts from Popov's 1896 paper, "It is beyond question, however, that the use he made of his apparatus was not the communication of intelligence to a distance, but for studying atmospheric electricity."[66] Fleming also cited (in his own translation) the final paragraph of Popov's paper.

On learning of the Soviet efforts to publicize Popov from discussions with Eric Ashby upon his return to

[63] V. S. Gabel, *Wireless World*, vol. 18, p. 319; 1926. See also editorial, p. 282.

[64] See, for instance, *Wireless World*, vol. 29, pp. 155, 186–187, and 240–241; 1943.

[65] Ref. 57, 3rd ed., p. 70.

[66] J. A. Fleming, "The Principles of Electric Wave Telegraphy," Longmans, Green and Co., London, England; 1906.

Australia, two of his colleagues, V. A. Bailey and K. Landecker, re-examined what evidence was available to them and concluded that Fleming had been grossly unfair to Popov by omitting a vital portion of his paper. "To explain this curious situation," they wrote, "some cynical persons might point out that Fleming was for some years closely associated with Marconi in the early stages of the development of Marconi's work in radiotelegraphy."[67] Fleming was a warm partisan of Marconi and had, in fact, been his paid consultant for many years; but he may be safely absolved from any charges of tendentiousness in this instance. The paragraph that the Australian scientists had found missing from Fleming's citation was merely the additional information that Popov had appended to update his paper in writing to the editor of *The Electrician*[15] in 1897—information that was not, as we have seen, in the original paper; since they had been unable to find the original publication in Australia, they had simply used Popov's letter to *The Electrician* instead and were thus misled. Had they bothered to look up the other indexed reference to Popov in Fleming's book,[68] they would have found that the missing information, correctly identified as originating from Popov's letter to *The Electrician*, was given on another page. This unfortunate misunderstanding was presently enhanced by the circumstance that no lesser personage than the president of Section A of the British Association referred to their "careful study" in his presidential address the next summer.[69] Moreover, a distinguished American engineer next suggested in a curious essay on the nature of invention (prompted by the Popov claims) that the task of establishing priority was altogether too difficult.[70]

The controversy attracted the attention of Professor George William Osborne Howe (1875–1960), respected doyen of British radio engineers and since 1926 the editor of the *Wireless Engineer*. In a series of editorials,[71] Howe reviewed the materials available to him, casting doubts on the plausibility of the "Heinrich Hertz" demonstration and correctly surmising that the "omission" on which the two Australians had based their case was no omission at all. He did not go so far as to check the original, although it was available in London, causing a reader (another Australian) to ask why Howe did not "settle the question unequivocally."[72] Howe replied in another editorial[73] but he did not touch upon this point, even though another correspondent had meanwhile compared the two versions and advised Howe.[74]

But Howe did raise certain questions, which may be summarized as follows.

If a well-attended lecture featured a demonstration during which a radiotelegraphic message consisting of the words "Heinrich Hertz" was transmitted, why was there no mention of it in the minutes? Even if we are prepared to accept that a naval officer had cautioned Popov and Rybkin to be careful, would that have prevented Popov from mentioning the transmission in his letter to Ducretet (a friendly correspondent who was manifestly trying to establish Popov's priority) written 18 months later, *after* Marconi's advances had become well known? Even if we are prepared to accept that it was this warning which prevented Popov from mentioning the transmission, why is it that Georgievsky, Lebedinsky, and Rybkin, who were all present at the demonstration, did not mention the transmission in the 1925 memorial number of *Elektrichestvo*, to which each of them contributed an article? Furthermore, if Popov slacked off work on his receiver and turned to X-ray experiments during the winter of 1895–1896, how was he able to make so much progress with the receiver that he attracted the attention of the naval authorities? Even if we are prepared to accept that he performed radiotelegraphic experiments early in 1896, why did a colleague recall that when the news of Marconi's experiments reached Popov in September, 1896, it shook him brusquely out of his torpor?

Soviet Claim Evaluated

In evaluating the Soviet claim that Popov "invented radio," we should recall that, patent law apart, priority of invention may be established on one of two bases: 1) priority of printed publication and 2) subsequently unearthed historical evidence.

The first, public disclosure in print, is by far the more generally accepted *sine qua non* of priority of invention. For instance, it is well known and adequately documented that Cavendish discovered in 1771 that electrostatic charges produced a force proportional to the inverse square of the distance between them, but he did not publish his findings.[75] The first to do so, 13 years later, was Coulomb, by whose name the law is known; no one would dream of recommending that it should be renamed "Cavendish's law." On this basis, Marconi's patent application of June 2, 1896, indisputably establishes his priority in the invention of radiotelegraphy (and hence of "radio," although, as has been pointed out,[76] that term did not come into use until much later). Alleged secret military tests cannot be taken into account in Popov's case any more than in the case of Jackson, who was perhaps experimenting with radio signals before Popov *or* Marconi. None of Popov's three

[67] V. A. Bailey and K. Landecker, *Australian J. Sci.*, vol. 9, pp. 126–129; 1947.
[68] Ref. 66, pp. 361–363 (pp. 369–371 in the 1919 edition cited by Bailey and Landecker in Ref. 67).
[69] E. V. Appleton, *Advancement of Sci.*, vol. 4, pp. 157–165; 1947.
[70] I. A. Mouromtseff, Proc. IRE, vol. 38, pp. 609–611; 1950.
[71] G. W. O. Howe, *Wireless Eng.*, vol. 25, pp. 1–5, 135–137, and 300; 1948.
[72] J. B. Thornton, *Wireless Eng.*, vol. 26, pp. 141–142; 1949.
[73] G. W. O. Howe, *Wireless Eng.*, vol. 26, pp. 249–250; 1949.
[74] Private letters to Howe from V. L. Rastorgoueff dated May 12 and 21, 1949.
[75] See, for instance, C. Süsskind, *J. Franklin Inst.*, vol. 249, pp. 181–187; 1950.
[76] F. Hamburger, Jr., Proc. IRE (*Editorial*), vol. 49, p. 1373, 1961; C. Süsskind, Proc. IRE (*Correspondence*), vol. 50, pp. 326–327; L. Espenschied, *ibid.*, pp. 327–328.

appearances in print before that date mentions his use of his equipment for the *transmission of intelligence by unguided invisible electromagnetic waves*, a definition of "radio" so generous in its compass that even the most severe critic could surely not find fault with it. Popov did publish the suggestion that his receiver might be used in connection with radiotelegraphy, and one may even speculate whether that offhand remark could have been conceivably utilized in a possible court suit to disallow some of the receiver claims of Marconi's patent. According to the British patent law then in force, the success of such an action would have depended on the court's view of the availability of Popov's paper to the British public; moreover, he would have had to show that he not only had the idea, but also knew how to implement it. One may reasonably doubt that the court would have been satisfied on that account. His remark, as Blondel presently pointed out,[44] showed that Popov was on the wrong tack: the primary obstacle to be overcome was not the lack of a more powerful transmitter, but of a more sensitive receiver. But however far afield such flights of speculation may take us, it is a fact that Marconi's patent was never successfully challenged; and Popov's only British patent related to an improved coherer.[39]

Turning to the second basis (historical research), if we take into account Popov's practical engineering background and the fact that in his early research he clearly distinguished between atmospheric and man-made discharges, it seems entirely plausible that he may have been aware of the practical possibilities of his apparatus from the first. (A newspaper account of his first experiments even states that "the *theoretical* possibility [my italics] of signaling over distances without conductors, after the manner of the optical telegraph but with the help of electric rays, serves as the motive for all these experiments."[77]) It is a little more difficult to cast aside all of the misgivings first raised by Howe in connection with the "Heinrich Hertz" demonstration. Nothing appeared in print about it until 1925, with the exception of an indirect reference in the report of the 1908 investigating committee. (The report states: "During the winter of 1896 Popov constructed the apparatus that is now preserved at the Torpedo School at Kronstadt. The purpose of the apparatus was the transmission of signals over distances . . . This apparatus was demonstrated by him in the application to the demonstration of the Hertz experiments at the meeting of the R. F.-Kh. O. on March 12, 1896; the application of the apparatus to the transmission of signals over distances was then still kept secret."[49]) Nevertheless, let us cast aside any residual skepticism for the moment and accept that on March 24, 1896, Popov demonstrated the transmission of a message to others. But if we admit that claim, made on the basis of historical research rather than priority of printed publication, we must also re-examine Marconi's achievements on the same basis. Every account of Marconi's early experiments shows that he was transmitting messages well before that date. The principal biographer of Marconi states that he started experimenting in the autumn of 1894 and demonstrated sending the Morse letter *s* "after having worked for a month or more."[78] Fleming says that Marconi "began experimenting in June, 1895."[66] R. N. Vyvyan says of Marconi (whom he knew intimately) that "his first efforts were made in 1894."[79] Marconi himself stated under oath, in a subsequent court action,[80] that he had conceived the idea of his first patent[26] "during the fall of 1894 or early in 1895" and added, "I commenced my experiments on this subject in the early summer of 1895," which would seem to bear out Fleming's "June, 1895." In his classic paper on wireless telegraphy, Marconi spoke of "experiments in Italy in 1895" that were by no means his first.[81]

There is not the slightest doubt that Marconi likewise demonstrated transmission of messages to others—certainly to friends and to his family, whose financial support was necessary to enable him to continue his experiments. A month before Popov's demonstration, Marconi arrived in London, having perfected his radiotelegraphic equipment to a point at which the considerable expense entailed in such a trip was justified.

Conclusions

In conclusion, then, it may be stated that:

1) On the basis of printed publication, Popov cannot be said to have "invented radio," since he did not describe in print his use of his equipment for the transmission of intelligence before Marconi's patent application of June 2, 1896. According to this generally accepted criterion, if any single individual may be designated as the "inventor of radio" (*i.e.*, radiotelegraphy), it is Marconi.
2) On the basis of historical research, there is indirect evidence that Popov demonstrated the transmission of intelligence to others on March 24, 1896. There is comparable evidence that Marconi demonstrated the transmission of intelligence at an even earlier date (though admittedly not before a scientific audience). According to this criterion, Popov's signaling work was *at the most* contemporary with Marconi's, and probably later.

[77] *Kronstadtsky Vestnik*, No. 51(4156); April 30 (May 12, N.S.), 1895; cited on p. 13 of Ref. 80 (this book was not available to me during the present study; the quotation was supplied indirectly by Popov's daughter E. A. Popova-Kianskaia and by G. K. Serapin of Leningrad, via a mutual correspondent, E. Aisberg of Paris).

[78] O. E. Dunlap, "Marconi: The Man and His Wireless," The Macmillan Company, New York, N. Y., p. 13; 1937.

[79] R. N. Vyvyan, "Wireless Over Thirty Years," George Routledge & Sons, London, England, p. 12; 1933.

[80] Testimony in "Marconi Wireless Telegraph Co. of America vs De Forest Wireless Telegraph Co.," U. S. Circuit Court, Southern District of New York, vol. 2, p. 532; September 22, 1904.

[81] G. Marconi, *J.I.E.E.*, vol. 28, pp. 273–297 and 300–316; 1899.

In Howe's words, "nobody questions the fact that Popov transmitted and received Hertzian waves and published an account of it six months before Marconi applied for his patent."[71] Many investigators transmitted and received Hertzian waves and published accounts of it before 1895, beginning with Hertz himself. Many others suggested their use for communications, including Huber (1889), Threlfall (1890), Trotter (1891), and Crookes (1892).[82] But Marconi was the first to publish a description of an existing apparatus specifically adapted for transmitting a message and used for that purpose.

All credit should go to Popov for independently evolving the same practical receiver design from Lodge's first suggestion as Marconi did, and for carrying out further experiments in the face of substantial obstacles. From a tour of radiotelegraphic installations in Germany and France, Popov could write to his assistant Rybkin with justifiable pride in 1899: "I have seen and learned everything possible, I have spoken to Slaby and seen his apparatus, I have visited Blondel at the station in Boulogne. In a word, I have learned everything possible and I see that we have not fallen much behind the others."[83] There is every indication that Popov's subsequent work was likewise of the highest caliber; had he had the opportunity (he died in 1905), he would have doubtless continued to make important contributions. The Russians have good reason to be proud to have produced a pioneer of Popov's rank; but the officious Soviet campaign to designate him the "inventor of radio" and to enlarge his reputation out of proportion with his achievements amounts to a deviation from objectivity that must be deplored by all historians of technology who remain untouched by chauvinistic considerations.

The entire world has been impressed by the recent technical achievements of the Soviet peoples; as they near the front rank in the march of modern technology, they can well afford to abandon the unbecoming and recurrent protestation of priority characteristic of an earlier era, when they were farther behind and their leaders considered it necessary to carry over the personality cult from the political sphere to the field of modern invention.

[82] Heinrich Huber asked Hertz in a letter dated December 1, 1889, whether his waves could be used for telephony; Hertz replied on December 3, 1889, that the proposed scheme would not work at the long wavelengths corresponding to audio frequencies. (The correspondence is preserved at the Deutsches Museum in Munich.) Richard Threlfall (1861–1932) suggested radiotelegraphy in his presidential address to Section A of the Australasian Association for the Advancement of Science (*Rep. Australasian Assoc.*, vol. 2, pp. 27–54; 1890). Alexander Pelham Trotter (1857–1947) was editor of *The Electrician* when an unsigned editorial suggesting radiotelegraphic communications with lighthouses and lightships appeared (*Electrician*, vol. 26, p. 685; 1891). William Crookes (1832–1919) predicted radiotelegraphy and correctly enumerated the three principal obstacles that would have to be overcome—the design of reliable transmitters, of sensitive tuned receivers, and of directional antennas—in an article written for a literary journal (*Fortnightly Rev.*, vol. 51, pp. 173–181; 1892).

[83] P. N. Rybkin, *Radiotekhnik*, No. 8, p. 269; 1919. (Cited by V. K. Lebedinsky, *Elektrichestvo*, No. 4, pp. 207–211; 1925 and in Ref. 59, pp. 38–40.)

CORRECTION

R. H. Pantell, co-author with J. R. Fontana of "Theoretical Considerations on Millimeter Wave Generation by Optical Frequency Mixing," which appeared on pages 1796–1800 of the August, 1962, issue of PROCEEDINGS, has called the following to the attention of the *Editor*.

On page 1798, (13) should read:

$$W_{1,0} + W_{0,1} + 2W_{1,-1} = P_{\text{in}} - 2P_{\text{out}}$$

$$= \frac{1}{4\pi^2}\int_0^{2\pi}\int_0^{2\pi} dxdy\,\frac{di}{dv}\left[\left(\frac{\partial v}{\partial x}\right)^2 + \left(\frac{\partial v}{\partial y}\right)^2\right].$$

On page 1799, column one, line 16 should read:

. . . For this case, we have that $di/dv = 0$ . . . .

Neuristor—A Novel Device and System Concept*

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Summary—**Certain basic problems in integration and extreme miniaturization of electronic systems may require new forms of device structure. A device structure that shows promise is hypothesized in this paper. The device is not discrete; it is rather a distributed "line," with active and passive processes so distributed that a signal propagates along the line without attenuation, much as a burning zone moves along a fuse or an ionic discharge along a nerve axon. It is shown that lines of this type can be interconnected in certain direct ways so that complete logic capability can be achieved with networks of such lines alone. Techniques of logical synthesis of these networks are introduced. Certain special properties of systems built in this fashion are delineated, especially properties that appear to be significant for the synthesis of highly-miniaturized systems. There is some discussion of physical realizability of distributed electronic structures.**

I. Introduction

General

MASS FABRICATION and miniaturization of electronic systems introduce serious problems of wiring and interconnections. As device dimensions decrease, there is no corresponding reduction in the number of connections per device; eventually it will be necessary to make a dense set of connections at a "point." Again, people talk of depositing "wires" having dimensions in the thousand-Angstrom to one-micron range, but even a one-micron copper wire has a resistance level on the order of 10^3 ohms/in. This may not be troublesome when the connected elements are but microns apart, but where several inches of wire are involved, signal attenuation may become critical.

The neuristor concept evolved from a consideration of these problems [1]. To overcome potentially high attenuation, it was of interest to consider the use of "active wire." The kind of "active wire" sought is exemplified by the chemical fuse that permits attenuationless propagation of a burning zone, or the axon of a nerve fiber that permits the attenuationless propagation of the ionic discharge. The fuse does not recover after such a propagation, and can be used only once. The axon, however, recovers in a few milliseconds and can be used an indefinite number of times. A forest fire burns in much the same manner, attenuationlessly with recovery. *Neuristor*, derived from neuron, is the name suggested for the whole class of lines that exhibit attenuationless propagation with recovery. Of special interest for our purposes is the synthesis of neuristor lines that operate on electronic principles.

The general structure and use of neuristor lines is the primary consideration of this paper. It is pointed out that there are ways in which such lines can be interconnected so that complete digital logic capability can be achieved with networks of neuristor lines alone. Systems synthesized in this manner offer the potential for a high degree of physical miniaturization.

Sections I and II offer an over-all picture of line structure and the interconnection of lines. Section III introduces the logical design of neuristor networks, while physical realization of lines based on electronic processes is discussed in Section IV. Section V is focused on some of the unique properties of neuristor systems.

Study of the structure and properties of such lines, and of networks of such lines, involves interesting problems and considerations in device physics, nonlinear differential equations, circuit analysis and synthesis, and switching theory. This paper is intended as a general introduction to neuristors; individual topics will be discussed in detail in subsequent papers. Eight reports on the subject are listed below [1]–[8].

Some Neuristor Realizations

A neuristor may be visualized as a channel having energy available to it everywhere along its length. The line or channel includes certain triggerable (active) processes arranged so that when any section of line is triggered, the locally available energy is itself converted into "trigger form." Thus, successive neighboring sections of line are activated. A signal propagating in this manner is generally referred to as a *discharge*, since it is continually regenerated by discharging available energy into the line as it propagates.

A fire burning along a narrow, uniform channel has the characteristics of such a propagating discharge. A less familiar but equally valid example is the nerve axon.

Nerve Axon: A nerve axon is a narrow channel on the order of one micron in diameter and up to several feet in length [2], [9]. An ionic discharge triggered at one end propagates toward the other end without attenuation and at a constant velocity (from about 1 meter/sec to about 100 meters/sec, depending on the size of the fiber). The energy available along the line is in the form of energetically separated ion concentrations (chemical potential energy) achieved by a suitable metabolic ion pump. The mechanics of the discharge are governed primarily by the properties of the 50- to 100-Angstrom-thick semipermeable nerve membrane. The typical recovery period is one millisecond, so that pulse rates along nerve fibers are generally limited to 1000 per sec-

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ond, each pulse propagating in the attenuationless mode. The time during which a portion of line is recovering, and can therefore not respond to further triggering, is referred to by the neurophysiologists as the *refractory period*. This term will be used here to designate the recovery period for any neuristor line.

Simple Relay Line: In Section IV, techniques for synthesizing distributed electronic neuristor lines are discussed. Here, however, let us limit our consideration to the discrete-element relay neuristor line of Fig. 1.

The line operates as follows: in the resting state, all relays are de-energized and the capacitors, C, are charged to line voltage. There exists a resting energy, $1/2C(kV_T)^2$ per stage, where k is a constant and V_T represents the energizing voltage of the relay. If any node voltage reaches the value V_T, the corresponding relay closes, applying voltage kV_T to the node. For $k>1$, this represents a signal gain when the contacts close. With k sufficiently large, voltage kV_T will cause the voltage of the following node to be $\geq V_T$, so that the contacts of that stage close, and so on. Note that once the contacts in any stage close, the relay remains energized until the corresponding capacitor is essentially depleted of energy (each section of line "burns" to completion) after which the contacts open and the capacitor recharges through resistor R_c. The length of the propagating active zone, *i.e.*, the number of adjacent relays energized at any instant, depends on the exact properties of the line. (In experiments with lines of this type, the active zone was typically six or seven stages wide [3].)

The line in Fig. 1 is longitudinally symmetric. Also, if the recharge paths were removed ($R_c \rightarrow \infty$), then after a single discharge the line could not be used again, providing a good electric analogy for an ordinary chemical fuse.

Comment on Line Structure and Properties

The operation of a general neuristor line seems to be describable in terms of the interaction of a trigger wave and an energy flow. This interaction is governed by a transducer that reacts to the trigger; but there must be gain. The trigger must cause a release of energy which forces the trigger parameter further than the initiating cause itself would. Thus, an electrical transducer element could not be simply a nonlinear resistance; instead, its dynamic resistance must actually be negative over part of its range. (Design of lines with certain common types of negative resistance elements is discussed in Section IV.)

It may be that more than one process is sensitive to the trigger parameter. For example, in a second version of the relay line two antagonistic processes (relays) could be arranged, each responding at a different rate [3]. The fast process (first relay) initiates the energy flow, but the effect of the fast process is nullified as soon as the slower process (second relay) is initiated. Recovery follows the cessation of energy flow. In the relay

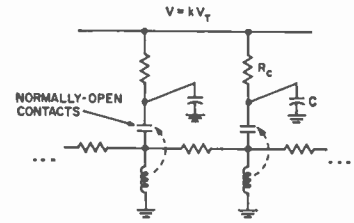


Fig. 1—Discrete-element relay neuristor line.

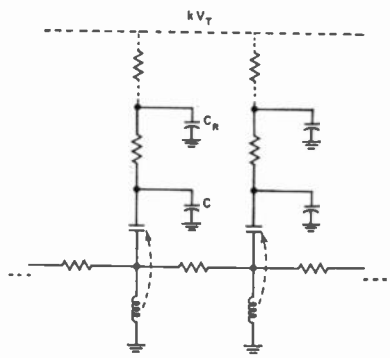


Fig. 2—Relay line that exhibits fatigue.

line in Fig. 1, on the other hand, only one transductive process controls the flow of energy, and energy flow ceases when available energy is depleted. Another process provides for recharging the local energy sources.

For convenience in this paper, we will henceforth refer to the one-process line as an F-type line, since a "recoverable" chemical fuse would operate in much this manner. The two-process line will be referred to as an N-type line, since a nerve fiber operates with a number of antagonistic processes [2]. But these two types are neither well defined nor all-inclusive.

The number of possible line structures is large. Besides those discussed, lines could be designed in which some processes depended on the time derivative, for example, rather than directly on the trigger parameter itself. Although all of the various possible lines might exhibit the desired attenuationless, recoverable propagation, they might differ in many important properties. As one example the property of fatigue will be exhibited by some lines, but not by others.

Fatigue can be shown in the following line that has a "multiple-level" recovery system. First, let the recovery circuits of the relay line in Fig. 1 be rebuilt as indicated in Fig. 2. Then, assume that all capacitors are fully charged, and that $C_R \gg C$. After each propagated discharge, line capacitor C in each stage is recharged from reservoir capacitor C_R. The line can thus sustain only a finite number of propagations before the line shows fatigue, each propagation being slightly weaker than the preceding one. If the "second layer" of storage were itself connected to a recharging source (dashed circuit), then still more complex behavior would result. (In Section II some interconnection possibilities between "fatigue lines" are pointed out.)

The term *fatigue* is adopted from the physiologists who use this term to describe a similar property in nerve axons.

Summary of Important Points

Symmetry: All of the lines discussed are longitudinally symmetric. In Section III it is demonstrated that complete logic capability can be achieved using symmetric lines only. (No advantage is obtained in these systems by postulating asymmetric lines that could sustain propagation in only one direction.)

Form of the Propagating Signal:[1] The velocity and form of the propagating signal depend only on the properties of the line itself. "Sufficiently far" from the trigger source, the velocity and form of the burning zone along a fuse are independent of the strength of the trigger source. An input signal can thus determine "when" a discharge should propagate, but not what form it will take.

Refractoriness: Each portion of line becomes refractory for some period after the passage of a discharge. A primary effect of the refractory period is that some minimum spacing must exist between pulses propagating in the same direction along the line. The existence of a refractory period also has great significance for pulses propagating towards each other on a collision course.

Destructive Collision: A basic property of neuristor lines, of utmost importance to the logical development to follow, is that two discharges propagating towards each other are annihilated upon collision. At the instant of collision, the line is refractory on both sides of the collision point; thus, neither signal can continue to propagate. For example, two fires burning towards each other are annihilated when they meet.

Reflectionless Collisions: It is well known that signal reflections occur at discontinuities in linear transmission lines. A neuristor line is inherently free of such reflections of the discharge. When a burning zone reaches the (open) end of a fuse line, the burning terminates; it cannot propagate back through its own refractory zone. This property of reflectionless collisions (whether with another pulse or at discontinuities in a line) is very important in the logical design of neuristor networks.

II. Basic Interconnections

The two main attributes of a signal propagating on a neuristor line are the ability to trigger further energy discharge and the presence of a refractory period. These attributes provide the basis for a number of different kinds of couplings between lines, but not all of these are required simultaneously for logical completeness. The logical design techniques of Section III use only the first two coupling schemes noted below, the T junction and the R junction.

[1] To reduce the repetitious use of the term *discharge*, we will also use the term *pulse* to describe the propagating signal, as well as the term *signal*.

T Junction

In a T (trigger) junction, symbolized in Fig. 3(a), lines are interconnected so that each line can trigger the others. A signal propagating toward the T junction on any one line will thus initiate a signal on all connected lines. Such a connection could be made by joining together the ends of separate fuse lines. A T connection between relay lines is indicated in Fig. 3(b).[2]

R Junction

In an R (refractory) connection, two or more lines are coupled so that they are mutually refractory, but a signal on one will not trigger a signal on the others. The symbol used for this junction is indicated in Fig. 4(a), where the vertical lines represent the region of coupling. With lines having only one transductive process (the F-type line), the effect is achieved by sharing the available energy between the coupled lines over the length of the junction. Such a junction between relay lines is indicated in Fig. 4(b). Note that a pulse passing along either line will not initiate a pulse on the coupled line. But two pulses, one on each line, attempting to pass on the junction, are destroyed just as though they collided on a single line. (At the instant of collision, the energy to the right of the collision point is depleted by passage of the pulse from the right. The energy to the left is depleted by the passage of the pulse from the left.) This junction provides a logical inhibition function, in the sense that a pulse on one line can inhibit, or control, the passage of a pulse on the coupled line without initiating any pulse on that line.[3]

A convenient structure in logic design is the T-R connection of Fig. 5, in which two R-coupled lines are T coupled at one edge of the R junction. In this logical fan-in structure, a pulse arriving on input line A (or on line B) can continue to propagate only along output line C. There is no crosstalk between lines A and B.

Other Forms

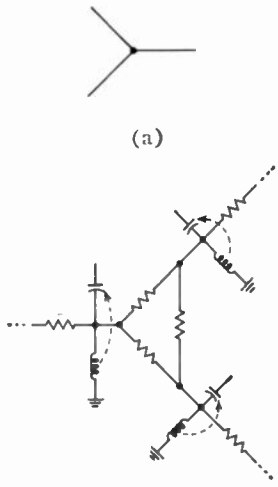
Although the logical design techniques of Section III depend only on the T and R junctions noted above, several other potentially useful interconnection forms can be hypothesized.

Weak Coupling: The T and R junctions may be thought of as "strongly coupled" in the sense that a

[2] Wherever one-dimensional lines come together to form junctions, or wherever there are nonuniformities of any type, the analysis of the propagation in those regions requires a two-dimensional treatment.

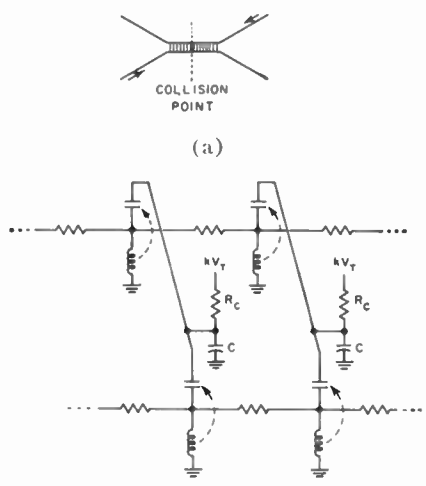
The generalization to two or three dimensions is intuitively clear however. For example, a fire ignited at some point of a plane propagates radially, retaining the basic properties of energy depletion with subsequent recovery. Mushrooms similarly grow in expanding rings, called Fairy Rings [10], with "velocities" on the order of several feet per year. In any case, colliding radial signals are destroyed just as they are on a narrow channel, so that the collision of two mushroom rings results in the destruction of their common sector, as does the collision of two fire rings. Such two-dimensional propagation may be very interesting for certain scanning and pattern recognition schemes [11], but these are not treated here.

[3] A refractory function can be achieved with N-type lines by sharing the slow process, but not the fast process [3].



(a) (b)

Fig. 3—(a) Symbol for a T function; (b) T junction of relay lines.



(a) (b)

Fig. 4—(a) Symbol for an R junction; (b) R junction between relay lines.



Fig. 5—T-R junction.

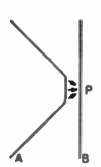


Fig. 6—Controlled-coupling junction. (A pulse on line A causes region P of line B to become refractory.)

single pulse arriving at a T junction actually causes a pulse to be initiated on every other connected line. But we could also consider weaker coupling, where a pulse is generated on a T-connected branch only if several pulses arrive simultaneously at the junction and their effects are summed. (This, in fact, is generally the situation at nerve synapses, where a single pulse generally has little effect [9].)

Fatigue Connection: An entirely different type of R junction results when the energy coupling between lines of the type in Fig. 2 occurs at the higher level. In other words, over the length of the junction, corresponding stages share a common reservoir capacitor C_R. In this arrangement, single pulses (one on each line) can pass each other on the junction without interference. Only if at least one line carries a relatively high-frequency pulse train will the common energy be sufficiently depleted so that pulse interaction can occur between the two lines.

Nonuniform Lines: Wherever the properties of a line vary, we can expect corresponding variation in the form of the propagating signal. For example, a burning zone propagating into a narrowing region of fuse line can be expected to reach temperatures higher than normal because proportionately less energy flows into the narrowing region ahead. A burning zone propagating into a widening region of line can be expected to develop a correspondingly lower discharge temperature. But barring extreme variation, the propagation can successfully proceed in both cases, implying that some nonuniformity is permissible in these lines.

For reasons of logic, however, more extreme nonuniformities might purposely be introduced. If a line expands very rapidly in size, the discharge temperature may become so low that forward triggering (ignition) actually fails. This arrangement offers a potentially useful unilateral structure, since a pulse propagating in the opposite direction, into a narrowing region, will surely survive [5].

Controlled Coupling: Consider the two lines of Fig. 6, connected so that a pulse on line A delivers energy to region P of line B, from over an extended distance and time. The strength of coupling is controlled so that in some first unit of time the energy transferred is not strong enough to initiate the propagating state. In the N-type line, such a weak trigger signal might still initiate some of the competitive slow process. Once the slow process had achieved its effective level, we could again couple an even stronger trigger signal without initiating propagation. After a number of repetitions, the inhibitory function at region P will have been fully activated, so that the line is refractory, without initiating any regenerating signal.

Similarly, in an F-type line, the proper rate of triggering can "drain off" the available energy without actually initiating a regenerative sequence. (This will be discussed later, in greater detail, in connection with Fig. 21.)

Suitable control of a trigger signal, therefore, might be used to cause a line to become inhibited. This suggests the possibility of achieving a *single basic type of coupling* between lines [5]. With strong coupling, a pulse on one line can cause triggering of the other line; with weak coupling, one line can inhibit or make refractory the

other line. For physical system realization, there may be considerable advantage to using a single basic connection scheme in which only the strength and form of the coupling is controlled. The potential physical simplification afforded by this approach is indicated in Section IV, in connection with the development of a distributed version of electronic line.

III. Logical Design of Neuristor Networks

This logical design study is based on the fundamental property of destructive head-on collision of two pulses, and on the use of strong coupled T and R junctions. The primary aim is to demonstrate complete sequential logic capability. It is well known that arbitrarily complex digital-logic systems can be synthesized with relays alone (early computer systems were built in just this way). We will therefore demonstrate complete logic capability by synthesizing with neuristor lines the direct analog of relay networks.

To emphasize the versatility of neuristor techniques, the relay analog demonstration will be followed by direct analogs for diode networks and transistor networks as well.

T Network

A *neuristor network* is defined as any configuration of neuristor lines and internal connections with any arbitrary designation of input and output terminals. A *T network* is a neuristor network in which the interconnections are all T junctions. Although T networks are not particularly useful by themselves, an understanding of their properties is important before generalizing to more complex networks.

The following two properties of T networks are stated without proof:

1) In a T network triggered at one point, every other point in the network is excited by one and only one pulse traveling in the direction of increasing distance from the point of triggering, and at a time determined by the shortest path from the point of triggering. (If one considers a fire burning along such a set of pathways, this property becomes quite clear.)
2) A T network cannot "oscillate." All network activity dies within a fixed time, all pulses being destroyed either by pairs in head-on collision, or by reaching the open end of a line. Recall that there are no reflections from collisions or discontinuities.

R Junction as a "Relay Contact"

Consider the R-junction arrangement of Fig. 7(a). As long as line C is unexcited, a pulse can be transmitted from A to B without interference. But if line C is excited with a sufficiently high-frequency pulse train, a pulse initiated at A must suffer a destructive collision somewhere along the junction. If the signal on line C represents the state of some logic-variable x, so that there is a pulse train for the state $x=1$, and no pulse train for $x=0$, then we can define a binary-valued transmission function, $T_{AB}=\bar{x}$ (logical negation of x), which is the same as the binary-valued transmission function for a normally closed contact on an x relay, as shown in Fig. 7(b). (With the relay energized ($x=1$) the contacts are open and transmission is zero. In the de-energized state, transmission is unity.)[4]

To make the transmission functions T_{AB} and T_{BA} equal, as they are for the relay contacts, the symmetric structure of Fig. 7(c) can be used. (Note that in the simple R-junction gate, a pulse from B is not on a collision course with the pulse train on C.) This results in a useful control junction, since the passage of a pulse on A or B is controlled by the presence of a pulse train on C. Let us now see how control junctions like these can be used in larger networks before we consider the source of the control trains themselves.

Controlled T Network

Consider the arbitrary relay network of Fig. 8(a) and the corresponding neuristor network in Fig. 8(b). For simplicity, contacts have not been labeled, although the same variable is assumed to control corresponding relay contacts and neuristor gates. For any particular state of the control variables, certain paths in the relay net are closed and others are open. For the same control state, the corresponding branches of the neuristor net can pass a pulse or not, so that for every open relay contact, the corresponding neuristor gate is excited by a pulse train. No branch of the neuristor network that contains even one excited control gate can pass a pulse from one end to the other. For an arbitrary control state, let us indicate by dashed lines those branches with one or more excited gates. The remaining network, indicated by solid lines, is defined as the *signal network*, and has the properties of a T network. The fact that control R junctions are connected to the branches of the signal network is not pertinent, since none of the control inputs to these R-junction gates carry signals.

Therefore, for every control state for which a closed path exists from input to output in the relay net (unity transmission), there is a corresponding path in the neuristor net over which a pulse can successfully propagate. Furthermore, the first-noted property of the T network assures that a single input pulse will produce only one pulse at the output, regardless of how many different paths are available between input and output terminals. The second-noted property of the T network assures that all pulses traveling on the network become

[4] If L is the length of the R junction, pulse spacing in the train on line C must be $\leq 2L$. With greater spacing, note that a randomly-timed pulse from A may or may not be destroyed by collision. Thus, if N random pulses are entered at A an average of pN of them will pass the junction without collision, where the probability value p depends on the frequency of the train and the specific dimensions. (Synthesis of probabilistic networks is treated in some detail by Crane [1]. In particular it is indicated that a generalized "switching algebra" can be developed for probabilistically controlled gate networks.)

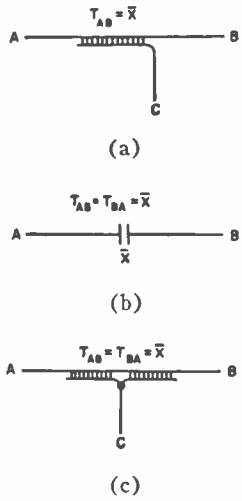


Fig. 7—R junction used to logically resemble a relay contact. (a) Asymmetrically excited R junction; (b) Normally closed contact on X relay; (c) Symmetrically excited R junction.

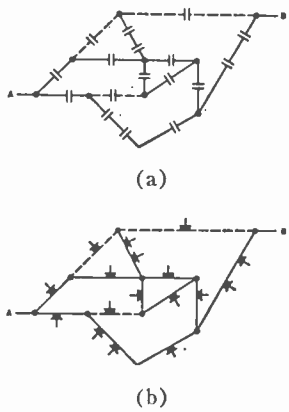


Fig. 8—(a) Arbitrary combinational logic relay network; (b) Corresponding controlled T network.

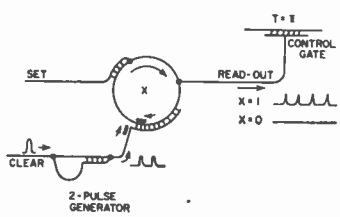


Fig. 9—Storage ring arrangement.

annihilated, so that no permanent circulations can be established. (Any pulses traveling along dashed branches are destroyed at some control gate along the path.)

Note that the T-network properties as stated do not involve any restriction on relative path lengths in the network. Hence, except for actual time delay between input and output, the transmission function for the controlled T network is totally independent of the length of the various branches, or of variations in velocity of propagation along the various paths.

The transmission functions for the networks of Fig. 8 are therefore identical, and any analysis and synthesis techniques used for relay networks can be used to design neuristor networks of this type.

To demonstrate the general usefulness of these combinational logic networks in sequential digital systems, we must investigate the manner in which control variables can be stored (source of the pulse trains). In addition, we must demonstrate that the signals obtained from the (controlled T) logic networks are compatible with those required for altering the state of the stored logic variables.

Storage Ring

One technique for storing binary variables uses the closed ring of Fig. 9. If the ring is longer than one refractory length, then a pulse once started in the ring (in either direction) will propagate continuously.[5] This does not violate the principle of conservation of energy because energy continually flows into the ring as the signal propagates. (Patterns of pulses can propagate over long rings, although we are interested here in circulating only one pulse.)

Let one pulse circulating in the ring represent the binary *one* state of the variable x. The binary *zero* state is represented by no circulating pulse. The question then is how the state of the storage element can be altered.

A pulse can be entered (set) into the ring via the input T-R junction. With this arrangement, the circulating pulse cannot cause a pulse to be transmitted back on the input line. To clear a circulating pulse, we can arrange to have it destroyed by collision. There is one timing condition, however, in which only the Clear pulse, but not the circulating pulse, will be destroyed in the "collision" at the junction edge. (Pulse positions are illustrated in Fig. 9 by heavy dashes at the left edge of the junction). But successful clearing can be assured even in this case if the single Clear pulse is replicated by the "two-pulse generator" and the ring arranged so that destruction of the first Clear pulse would leave the second Clear pulse to destroy the circulating pulse.

This provides an adequate, although by no means unique, method of alterable logical storage. Read-out from a T junction can be used to directly control an R-junction gate. If the stored variable were in the *one* state ($x=1$), a pulse train propagates on the output line, since a pulse is generated on each pass of the circulating pulse in the ring. For the binary *zero* state ($x=0$) the gate remains unexcited.

This demonstrates the sequential logic capability offered by neuristor lines limited to T and R interconnection. But to illustrate the variety of logic network properties, a few other approaches to logic synthesis will be briefly noted.

[5] Over a line in which signals propagate at constant velocity, the basic unit of length is equal to the unit of time multiplied by the velocity. In particular, then, the refractory length is equal to the refractory period multiplied by the value of the velocity.

Diode Analog

A pulse initiated at A in Fig. 10(a) cannot reach B, since the two secondary pulses A' and A'' are destroyed by collision in the R junction. A pulse from B, however, can successfully reach A since the pulse from the T junction propagating back toward the R junction is too late to affect the initial pulse. This structure thus exhibits a unilateral transmission property similar to that shown by a diode.

A diode network and its neuristor analog are illustrated in Fig. 10(b) and (c). Pulse-train frequency in the neuristor net is analogous to dc voltage level in the diode net. (This is a generally useful analogy, and will be used again in development of the transistor analog.) Given ideal diodes, the output of the diode net is equal to the level of the highest input voltage, and all other diodes are cut off. Similarly in the neuristor network, the frequency of the output train is precisely equal to the highest-frequency input. (To see this, consider the simpler case of two pulse trains of unequal rate colliding on a single neuristor line. The point at which pulses of the two trains collide is continually "driven back" by the higher-frequency train.) Also, as in the diode case, all other lines are cut off by the unilateral structures, and the higher-frequency output train does not penetrate back into them.

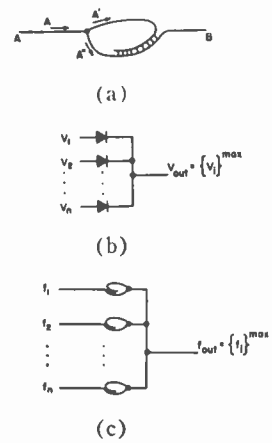


Fig. 10—(a) Unilateral structure; (b) and (c) Logically identical networks—voltage level in conventional diode network corresponds directly to frequency in neuristor network.

R Junction as a Transistor Analog

A transistor and an R-junction neuristor analog are shown in Fig. 11. The dc emitter-collector supply voltage V in Fig. 11(a) is analogous to excitation by a pulse train (PT) of the corresponding "emitter-collector circuit" in Fig. 11(b). In the transistor case, the higher the input voltage is, the lower the collector voltage will be. Analogously, the higher the input pulse-train rate is, the lower the output rate will be, assuming that each input pulse destroys exactly one emitter pulse. This corresponds to a unity voltage-gain transistor circuit. A gain, G, greater than unity can be achieved by arranging to have each input pulse destroy G pulses of the emitter train. This can be accomplished by inserting a "G-pulse generator," as in Fig. 11(c), into the "base" line in Fig. 11(d). (In the generator circuit, an input pulse makes G copies of itself which are then gathered together onto a single line. A $G=2$ pulse generator was used in the storage-ring clearing circuit, Fig. 9.) Note that the collector is saturated when the input rate becomes higher than the emitter rate divided by G.

To test the validity of this arrangement, examine the flip-flop arrangement of Fig. 12. If the input train emerges on output line A, then the output line B is completely cut off, and conversely. Thus, the emitter train passes out the left circuit or the right circuit, and this condition is stable until the state of the flip-flop is altered by a single pulse on the Set or Reset line (collector gating). "Base" or "emitter" gating could also be arranged.

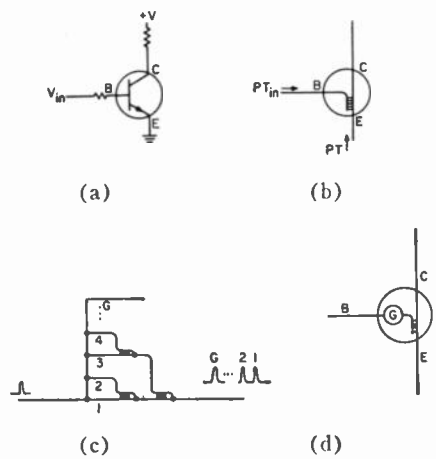


Fig. 11—(a) and (b) R junction excited in such fashion as to be analogous to grounded emitter transistor circuit. (c) and (d) G-pulse generator in "base" line provides gain of G.

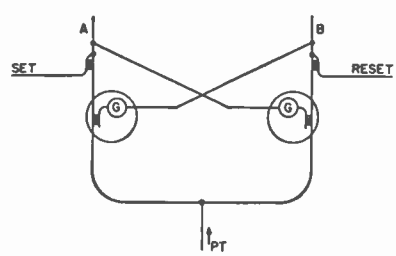


Fig. 12—Two neuristor "transistors" interconnected as a flip-flop.

Further Comment on Logical Design

A recurrent problem in the study of neuristor networks has been the application of familiar terms used in conventional logical design. The operation and structure of neuristor networks is sufficiently different from that of conventional networks (say, transistor networks) that conventional terms are not easily applied. For example, diode networks are "obviously" gating networks, while relay networks are just as "obviously" branch networks. Yet the same R junction can be used

to simulate diodes, transistors, or relays. It was necessary therefore to review and define more precise meanings for these terms.[6]

We have here considered logical design techniques based on data transmission in terms of frequency, where one state of a binary variable is represented by a uniform pulse train and the other state by absence of transmission. But in other systems that have been studied, the binary state of a variable might be represented by only a single pulse. In such systems every pulse is important, and a primary problem is that of keeping accurate track of the individual signals as they propagate through the network, so that different pulses actually interact at the proper time and place. This dynamic interaction condition has led to naming these systems dynamic-logic systems. The primary reason for considering such a design approach is the potential increase in processing speed when each variable uses a minimum space along a line [6].[7]

IV. Electronic Lines

Let us consider how lines based on familiar electronic processes might be physically realized.

Nonlinear Transmission-Line Viewpoint

Let us view a distributed neuristor line as we usually view a linear transmission line, identifying an equivalent circuit per unit length. We are here more interested in understanding the required basic structure of the line than in the analysis of any particular structure. Thus, let us start with a highly lossy (linear) transmission line. A section of such line is indicated in Fig. 13(a), where there is a longitudinal resistance R per unit length, and shunt capacitance, C, per unit length.

We have already pointed out the need for active processes in these lines if a signal is to continually reconstitute itself as it propagates. Three-terminal devices, such as transistors, could be used to achieve these objectives, but since our lines exhibit identical transmission properties in both directions, let us rather consider the generally simpler two-terminal elements. We will examine here the use of 1) negative-resistance elements that are single-valued in voltage, which we will refer to as $I(V)$ elements, 2) negative-resistance elements that are single-valued in current, which we will refer to as $V(I)$ elements. $I(V)$ elements will be indicated by a tunnel-diode symbol, since this is a typical number of that class, and $V(I)$ elements will be indicated by a rectangle.

The lumped circuits derived here can be viewed as the circuits of lumped-element lines, in the same manner as the earlier relay lines, or they can be viewed as the lumped-element equivalent of certain distributed lines.

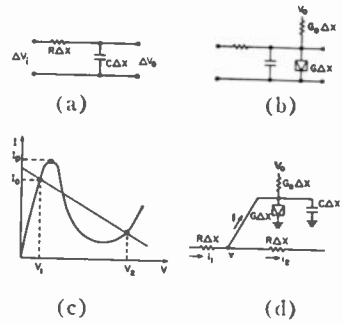


Fig. 13—(a) Single section of a lossy (linear) transmission line. (b), (c), and (d) Addition of an $I(V)$ device to each section in order to achieve attenuationless propagation.

(The worth of the equivalent circuit depends of course on how well it models the actual distributed line.)

Lines Based on $I(V)$ Negative-Resistance Elements

With the transmission line of Fig. 13(a), an input signal ΔV_i will lead to an output signal $\Delta V_o < \Delta V_i$. If an $I(V)$ element is connected into each stage, as in Fig. 13(b), and if the input signal is sufficiently large so that the "tunnel-diode" current becomes greater than the breakdown current I_p, then the "diode" would enter its active region and a significant increase in voltage would result. In order to increase the sensitivity of the "tunnel diodes," we add in each stage a load, G_0, connected to a supply, V_o, so that with no signal propagating on the line, all "diodes" are biased near to their breakdown point, with a current I_0 flowing [Fig. 13(c)].

The way the load-line is drawn, each stage is locally bistable; such a line, although it could propagate an attenuationless discharge, could not subsequently recover. In other words, with the line initially stable at voltage V_1 (every stage at voltage V_1), triggering will set off a signal-front that will propagate at a constant velocity; the section of line behind the signal-front will be left stable at voltage V_2. Although this is intuitively clear, proof can be obtained from the equation of propagation for this line. (This subject is treated in detail by Pease [4].)

If these local circuits had been made suitably monostable instead of bistable, then recoverable propagation could be achieved. Attenuationless propagation along a coupled chain of monostable circuits implies that each successive stage has the same history. Furthermore, if the basic monostable circuits are bilaterally symmetric, then propagation can proceed in either direction. In order to indicate the form that such circuits might take, three relatively simple lines are briefly described.

In the first case, Fig. 14, monostability is achieved by choosing the value of load conductance so that there is but a single crossing with the characteristic of the diode. A sufficiently large transverse inductance provides the necessary stored energy per unit length, $1/2LI_0^2$, where I_0 is the quiescent current through the inductance [12].

In the second case, Fig. 15, monostability is achieved by using a variable load conductance whose value is a

[6] Crane, Minnick, and Short [6] present a relatively comprehensive study of logical design procedures for neuristor networks.
[7] These dynamic-logic concepts appear to apply equally well to neuristor systems and to more conventional device systems. See [6].

function, for example, of its own power dissipation. In the resting condition with a current I_1 passing through the load, the conductance is of such a value that the load-line crosses (line A) in three places. But when the stage switches towards its "upper state," where a current I_2 tends to flow, the value of conductance changes and the load-line swings so that eventually no upper crossing exists, load-line B, and the stage recovers.

In the third method, Fig. 16, an active $V(I)$ element is added to each stage. The breakdown voltage, V_p, of the new element is made less than voltage V_2. As the $I(V)$ element breaks down, the node voltage heads towards V_2, but before it can reach V_2, the $V(I)$ element breaks down. If the characteristics of the $V(I)$ and $I(V)$ devices are properly matched, the circuit will be driven back towards its original state.

Lines Based on $V(I)$ Negative-Resistance Elements

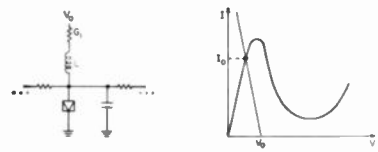
If a $V(I)$ element were added to each stage in the same way that an $I(V)$ element was added in Fig. 13(b), the line could not work. The responding voltage change for $V(I)$ elements is in the opposite direction from the triggering voltage. (When a gas-discharge element breaks down, for example, its voltage decreases, whereas triggering requires an increase in voltage.) With $I(V)$ elements, on the other hand, voltage continues to increase after breakdown. If such a line of $V(I)$ elements is to work, the polarity of every stage must be reversed, as suggested in Fig. 17 [13].

By using $V(I)$ elements in a "three-terminal" mode, as suggested in Fig. 18(a), a line of monostable circuits can be formed in which alternate polarity along the line is not required. In the line shown the $V(I)$ element is a four-layer diode. As each stage fires the three terminals of the device are effectively short-circuited, and capacitor voltage is impressed on the line (resistor R prevents the capacitor from being short-circuited). The resulting longitudinal current fires the next following stage. (Capacitor C per stage represents the total line capacitance, including the input capacitance of the device.)

The line of Fig. 18(b), based on common logitudinal layers, operates in the same manner except that interstage signaling is by direct diffusion of minority carriers. As each stage fires, there is generated a high density of minority carriers in the central layers. These minority carriers diffuse, causing breakdown of the forward adjacent element, and the propagation is maintained.[8]

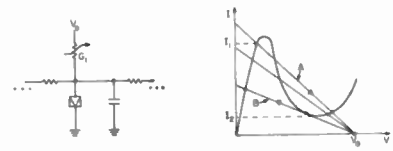
To complete the distribution of this "semidistributed" line would require fabricating a common fourth layer and a continuous layer of RC circuits. One potential

[8] At the time of writing, an experimental semidistributed structure of this type having three common layers had been constructed and operated. The first version of such a line operated with a propagation velocity on the order of 10^4cm/sec; and the propagating pulse had a refractory length of several millimeters. Thus, pulses propagating in the same direction had to be spaced several millimeters apart.



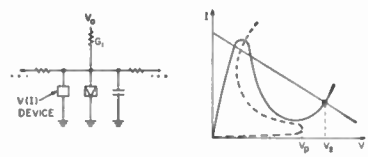
(a) (b)

Fig. 14—Monostability achieved by having load line cross $I(V)$ characteristic at just one point.



(a) (b)

Fig. 15—Monostability achieved by having a suitably variable load line.



(a) (b)

Fig. 16—Monostability achieved by having suitably matched $V(I)$ and $I(V)$ devices.

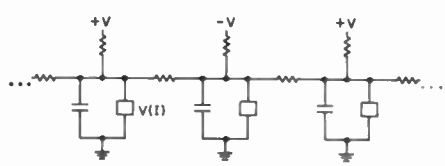
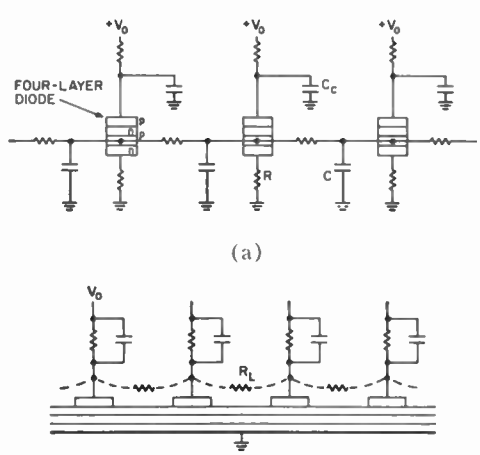


Fig. 17—Line based on $V(I)$ devices. (As shown in text, neighboring stages must be connected to opposite polarity supply voltages.



(a)

(b)

Fig. 18—$V(I)$ Elements used in a three-terminal mode. (a) Lumped-element line. (b) Semi-distributed version.

problem exists here. The equivalent circuit of any totally distributed structure of this type will always have some longitudinal component of resistance between stages, the resistance R_L in Fig. 18(b). But for any given line there is a minimum to the value of R_L that can be tolerated. With $R_L=0$, for example, there can be no propagation since all of the storage elements (capacitors) would be short-circuited.[9]

If the longitudinal-resistance minimum can be met, the structure of Fig. 19(a) is a feasible distributed model. The sawtooth line represents a lossy dielectric layer, which exhibits, in an equivalent-circuit sense, parallel RC components. (The equivalent circuit of a differential slice through such a line would have the approximate form shown in Fig. 18, where one side of the capacitance element is returned to supply voltage rather than ground; this does not affect the ac response.) Supply voltage is applied to the top and bottom of the structure. Viewed from the top, the line can be considered as having two major parts, a trigger pathway and a covering (dotted) layer of energy storage capacity, as in Fig. 19(b).

Junctions

With the distributed line of this type, a T junction is realized as indicated in Fig. 20(a). For an R junction, the energy storage is shared over the length of the junction by coupling the energy-storage layers. A potential difficulty here is that the conductive coupling film must have a high longitudinal resistance to give an acceptable value of R_L, while the transverse resistance must be low to provide a good energy coupling. What might be required to make this physically realizable is some form of anisotropic film.

The controlled-coupling scheme discussed in Section II might eliminate the need for the direct R interconnection (the coupling of the energy-storage layers). Thus, consider a single stage (equivalent circuit of a unit length) of line. Whenever the operating point is to the right of the load-line, the relatively high current must be supplied from the capacitor, which is thereby discharged. Hence, to the right of the load-line, node voltage v is decreasing. A typical path of the operating point when this stage is fired is path F of Fig. 21(a). If the triggering rate is low enough, however, it should be possible to discharge the capacitor without ever actually firing the stage. In this case the path labelled I would apply.

The controlled-coupling capability might be useful in structures such as in Fig. 21(b), where there is a continuous underlayer of trigger pathway, but the energy storage layer is deposited selectively, in the dotted region shown [5]. A discharge arriving along branch B moves along the junction region, being continually regenerated along the path having energy storage; at the same time, carriers diffuse through the continuous trigger layer towards branch A. Hence, any point P on line A sees a relatively low rate of carrier triggering. With proper design, the energy at point P could be depleted before the arrival of the pulse back along branch A, with resultant destruction of this propagating pulse. The pulse along branch C, however, would propagate unimpeded. This structure, although based on trigger coupling, therefore has the same forward characteristics as the T-R connection of Fig. 5.

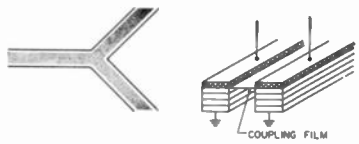
Possible simplifications emerge when we extend this construction. Consider for example the structure of Fig. 22(a). A pulse arriving on any branch propagates suc-

[9] Because of concern for a low value of R_L it appeared safer in the first experimental model to remove the relatively high conductance fourth layer entirely and replace it with individual dots.



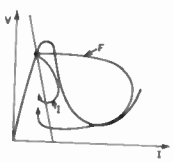
(a) (b)

Fig. 19—Distributed version of line of Fig. 18(b).

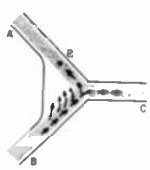


(a) (b)

Fig. 20—(a) T connection; (b) R connection of distributed lines shown in Fig. 19.

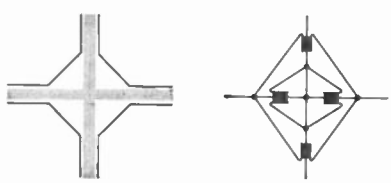


(a)



(b)

Fig. 21—(a) Path of operating point with strong coupling, F, and with weaker coupling, I; (b) Controlled-coupling junction having the same forward characteristics as the T-R junction of Fig. 5.



(a) (b)

Fig. 22—(a) Controlled-coupling crossing structure. (b) Logically similar structure based on T and R junctions. (In both cases an incoming pulse propagates only straight through the structure.)

cessfully only straight through the structure, since the side branches are inhibited by slow (controlled-rate) triggering. This type of crossing structure permits two lines to cross in two dimensions without interaction.[10] The comparative simplification offered by controlled-coupling structures becomes apparent when examining an equivalent structure realized with only T and R junctions, as in Fig. 22(b). Here too an incoming pulse will propagate only straight through the network, since side branches are directly inhibited by auxiliary pulses.

V. Special Systems Properties

From the material developed so far, several observations can be made about properties of neuristor systems. Some of these imply potential advantage for the realization of miniaturized systems.

Attenuation

Two reasons for beginning this study were 1) the potentially high-signal attenuation along tiny deposited "wires," and 2) the fact that reducing device dimensions does not decrease the number of connections to be made at each device. The attenuationless signal mode of neuristor lines eliminates the first problem.

Connection Problem

We can synthesize neuristor lines by interconnecting discrete stages, as in the relay lines of Section I, or the discrete, four-layer diode line of Fig. 18, but the large number of connections required could nullify the worth of such lines for miniaturization. The problem of making a large number of bonds between heterogeneous materials already promises to be one of the major technological problems in mass fabrication. In distributed neuristor lines, on the other hand, the material over which the signal travels never changes form. There are no longitudinal connections along such a line, as in Figs. 18(b) and 19(a).

From the discussion of logic synthesis in Section III, and the comments on distributed lines and their interconnection in Section IV, it should be clear that, for an arbitrary logic system, a physical realization can be demonstrated in which every signal would propagate along a continuous pathway of identical material. The form of the signal would never change until it was destroyed by collision with another pulse, or it reached the end of a line.

Transverse Connectivity

We have previously observed that lines can work satisfactorily in either a semidistributed form or a totally distributed form; the longitudinal path for signal flow is continuous in either case. Which technique is actually of more importance depends upon the specific device and material systems being considered. For example, in the semidistributed version of Fig. 18(b), the third layer is exposed between stages, and loss of minority carriers by surface recombination may be sufficiently serious to demand special treatment for these surface areas. Again, certain techniques discussed below for supplying energy to such systems might demand that the sections between stages of the semidistributed line be electrically insulated. Thus, it might actually be simpler to use continuously distributed lines. On the other hand, design of a continuously distributed line might involve a difficult longitudinal resistance problem, as discussed in connection with Fig. 18(b).

In principle, a totally distributed line could be made by multiple depositions through a single mask, whereas the semidistributed version would require multiple masks. On the other hand, if these lines were made by multiple tracings of an electron beam [14] it may not matter whether or not the beam is applied "staccato fashion" on the last pass.

It does not seem possible to make a blanket case for either the semidistributed or totally distributed line. The more basically important point seems to be the continuous longitudinal trigger pathway.[11]

Coupled Logic and Interconnection Properties

Let us now compare a tunnel-diode neuristor line (any of the type described in Section IV) with a more conventional tunnel-diode shift register. The particular form of tunnel-diode register is not especially important. The main point is to indicate that although both lines are physically symmetric, only the neuristor line is symmetrically energized as well. This results in certain interesting properties.

The shift register line of Fig. 23 is arranged so that each stage, when energized, is bistable and can hold one bit of data. In order to propagate the stored data along a physically symmetric line, it is well known that energy must be applied in an asymmetric fashion [15], [16]. If clock power is applied in the sequence V_1, V_2, V_3, V_1, V_2, $\cdots$ then information will be synchronously shifted to the right, and as in a three-phase motor, with reversed sequence V_1, V_3, V_2, V_1 $\cdots$ data will shift to the left. But once a clock sequence is selected, all data during that time move in one direction.[12]

[10] This structure is viewed as basically two-dimensional because both crossing lines could be realized simultaneously, say with deposition processes through masks. If the two lines crossed without touching, however, one under and one over, then we would view it as a three-dimensional crossing.

[11] Both semidistributed and distributed forms of nerve fibers appear to be found in the nervous system, these being referred to as myelinated and unmyelinated, respectively [11].

[12] The continuous magnetic-film shift register is an example of a distributed version of a line having similar properties [17]. Data are stored in terms of "reverse magnetic domains" along the film strip, and these can be stably moved to the right or left upon application of a proper time sequence of magnetic fields. The resulting operation is somewhat analogous to a magnetic drum surface on which the "bits" move instead of the drum.

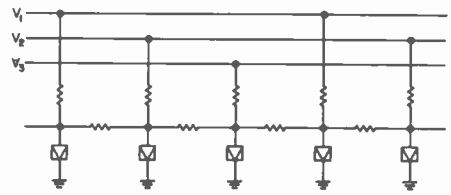


Fig. 23.—Tunnel-diode shift register excited from three-phase clock. (Binary data travels to right for sequence V_1, V_2, V_3, V_1, · · · ; to left for sequence V_1, V_3, V_2, V_1, · · · .)

Thus, the neuristor and shift register lines are both physically symmetric, and can potentially shift data in either direction. But the neuristor line is symmetrically energized as well, so that different signals can *simultaneously* propagate in opposite directions along the same line. For this reason, discharges can actually collide, with resulting annihilation. This destructive collision property, as we have seen, offers important logic possibilities.

In more conventional logic systems, the job of combining binary variables (logic) is separate from the job of transmitting the new results to other logic elements (wire interconnection). In neuristor systems, no such simple distinction can be made. In a controlled T network, for example, pulse collisions can occur at completely different points for different states of the control variables. Again, a pulse may propagate in one direction along any particular line segment for one state of the control variables, but in the opposite direction for a different control state. Thus, at least in some sense, the processing capability and the interconnection are tightly coupled in these systems.[13]

Planar Crossings

It was indicated in connection with Fig. 22 that two-dimensional crossing of lines can be achieved. In these structures there is no signal interaction between the lines, as long as pulses reaching the junction on the two lines have some minimum spacing (same requirements as for railroad trains crossing at a two-dimensional switch). In spite of the timing limitation it has been shown that for any digital logic system a realization can be demonstrated in which each such crossing structure is used in a safe manner [1].

Thus, not only can logic systems be realized in which there is not a single break in longitudinal connectivity, but the physical realization could be just "one-layer thick" on a two-dimensional substrate. Such a realization would not involve multiple masking and registration problems, in which certain wire patterns are deposited, followed by an insulating layer, then another crossing pattern of wires, and so on. Whether this planar crossing property will be of practical use or not cannot be determined at this point. Nonetheless the property does exist.

Energy Distribution

Let us next examine the problem of delivering energy to various circuits in a mass-fabricated, integrated network of considerable complexity. The amount of signal attenuation along tiny interconnecting wires makes it unappealing to consider bringing power into some single "terminal" on the wafer, and delivering this power over similar deposited wire lines in the usual fashion. If the power plant could be distributed throughout the network, each "substation" supplying only a local area, the problem of signal attenuation would be less critical. For example, dc voltage could be directly supplied by sandwiching the entire network between a pair of metal plates, one plate insulated by a mask having holes only at those places at which voltage supply is desired. Alternatively, appropriate forms of energy converters could be inserted where voltage supply is needed, and the entire network could then be embedded in a suitable energy environment, such as a chemical bath if chemical-electric (battery or fuel-cell) elements were used as converters.

In the case of neuristor networks, the distribution of power is taken to the limit. One would hardly consider delivering power over discrete wire paths to a continuously distributed line. Instead, energy would be made available everywhere. Consider, for example, an entire logic system realized in a single plane with lines of the type indicated in Fig. 19. Energy could be supplied to this network by sandwiching it directly between two conductive plates, across which a dc potential is maintained. Alternatively, suitable energy-converting processes could be distributed along with the line, in which case voltage would be generated everywhere simultaneously. The nerve axon in its chemical environment is an example of the latter, since a resting potential is generated everywhere simultaneously by suitable metabolic processes [2].

Dimensional Scale

The ability to achieve actual miniaturized neuristor systems of the type described in Section III will depend on 1) the ability to construct extremely small lines, and 2) the ability to achieve very short refractory length. Refractory length sets the physical scale of the networks; a storage ring, for example, must be at least one refractory length long. Although it does not seem possible to develop any meaningful estimates of ultimate limits, a particular form of superconductive neuristor

[13] Physiologists generally consider that neural "logic" is performed at the end of the nerve axons—at the synapses and cell bodies—and that axons mainly transmit the results. But treating the axon merely as "interconnecting wire," may mean overlooking some important processing that can be done directly in the axon transmission paths. It is intended to summarize these possibilities in a future paper, and to indicate how these properties may suggest processes worth searching for in the nervous system. (Neuron simulation studies that overlook these possibilities may therefore be similarly restrictive. Examples of such studies are indicated in Harmon [18].)

line can be used to illustrate the size range that may be possible. The line consists simply of a sandwich of two long, thin films, each about 50 microns wide, excited from a longitudinal current source.[14] (Although this line is small in terms of conventional circuits, nerve fibers considerably smaller than one micron in diameter are quite common.)

VI. Review and Conclusions

In studying neuristors, it was possible to go considerably beyond the notion of using such lines merely as means of interconnecting conventional devices and circuits. By observing that there are ways in which neuristor lines themselves may directly interconnect, it has been possible to demonstrate that digital systems can be synthesized with nothing more than suitably interconnected neuristor lines.

It is too early even to speculate about the practical significance of this approach. However, there are certain important implications for highly miniaturized systems that should not be overlooked. The key properties seem to be related to homogeneity. Successful synthesis of large-scale highly miniaturized systems will demand a significant degree of uniformity and homogeneity—the fewest possible different materials, in the fewest possible different configurations, and with a minimum of connections between different materials. It is here that neuristor systems appear to offer a great deal.

A decade or two ago distributing active processes throughout a network would probably have been considered impractical. At that time, there were few active elements apart from relays and tubes. The advent of solid-state electronics, however, has changed the picture vastly. Active processes are now so numerous that distribution of active and passive structures will in time probably offer about equal difficulty. Thus, just as passive elements are now becoming commonly distributed —*e.g.*, distributed RC lines—so we can expect active structures to be distributed in the future. In any case, analyzing the performance of integrated structures will require considering the active processes as distributed, whether distribution was the actual intent or not.

Finally, the study of neuristor logic systems has been stimulating and useful, quite apart from any question of physical realization. These studies have required a re-examination of terms, phrases, and concepts that have grown up around the more conventional systems of logic synthesis. Examination in the light of neuristor techniques has shown that a number of these notions require modification if they are to be truly general. These studies have also resulted in new insights in such areas as the analysis and synthesis of probabilistic networks,[4] in the design of high-speed systems,[7] and in certain aspects of physiology.[12]

The usefulness of the neuristor approach in the area of main concern—miniaturized, integrated systems—will depend primarily upon the ability to develop simple techniques of physical realization. Meaningful predictions cannot be made apart from such development.

[14] This type of line was initially conceived by Milton Green of Stanford Research Institute. The manner in which the line operates is discussed by Padovani [19] of Stanford University. Initial analog computer solutions indicate that a zone of "normal" conductivity could steadily propagate along this line at a velocity on the order of 10^6 cm/sec. The distance along the pulse over which there is significant temperature variation is several hundred microns. The actual refractory length depends partly on electrical time constants, however, and might be several times this amount, or a few millimeters.

Acknowledgment

The author is very indebted to J. Linvill, R. Anderson, M. Pease, A. Rosengreen, T. Janusz, F. Gair, R. Short, and R. Minnick for their many technical contributions, as well as for the many discussions regarding the significance, and equally as often the lack of significance, of the techniques described here.

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An Active Pulse Transmission Line Simulating Nerve Axon*

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Summary—**To electronically simulate an animal nerve axon, the authors made an active pulse transmission line using tunnel diodes. The equation of propagation for this line is the same as that for a simplified model of nerve membrane treated elsewhere. This line shapes the signal waveform during transmission, that is, there being a specific pulse-like waveform peculiar to this line, smaller signals are amplified, larger ones are attenuated, narrower ones are widened and those which are wider are shrunk, all approaching the above-mentioned specific waveform. In addition, this line has a certain threshold value in respect to the signal height, and signals smaller than the threshold or noise are eliminated in the course of transmission. Because of the above-mentioned shaping action and the existence of a threshold, this line makes possible highly reliable pulse transmission, and will be useful for various kinds of information-processing systems.**

I. Introduction

IN THE CASE of conventional pulse transmission lines, provided they are of great length, signals suffer attenuation and distortion as they travel down the line, whereas an electric pulse signal which is transmitted along an animal nerve axon suffers neither attenuation nor distortion, regardless of the distance covered. A pulse transmission line with such qualities will be useful for pulse transmission in communication systems, data processing systems, electronic computers, etc.

It is our purpose to realize such a pulse transmission line by simulating the animal nerve axon.

II. Excitation of Nerve Axon

A great deal of research on the electro-physiology of the nerve axon has already been made,[1] the most important being that of Hodgkin and Huxley.[2] They have made a quantitative study of the excitation of a nerve axon and the propagation of the excitation, and have derived the Hodgkin-Huxley equations (H-H equations) which describe the phenomena.

In the case of a "space clamp," that is, in a case where the excitation of a nerve axon is spacially uniform, the H-H equations describing the excitation of the nerve axon are as follows (see Appendix):

$$\begin{cases} I = C_0 \dfrac{dV}{dt} + \bar{g}_{Na} m^3 h (V - V_{Na}) + \bar{g}_K n^4 (V - V_K) \\ \qquad + \bar{g}_L (V - V_L), \\ \dfrac{dm}{dt} + \{\alpha_m(V) + \beta_m(V)\} m = \alpha_m(V), \\ \dfrac{dh}{dt} + \{\alpha_h(V) + \beta_h(V)\} h = \alpha_h(V), \\ \dfrac{dn}{dt} + \{\alpha_n(V) + \beta_n(V)\} n = \alpha_n(V), \end{cases} \tag{1}$$

$$\alpha_m(V) = 0.1(V + 25)\left[\exp\left(\frac{V + 25}{10}\right) - 1\right]^{-1},$$

$$\beta_m(V) = 4 \exp\left(\frac{V}{18}\right),$$

$$\alpha_h(V) = 0.07 \exp\left(\frac{V}{20}\right),$$

$$\beta_h(V) = \left[\exp\left(\frac{V + 30}{10}\right) + 1\right]^{-1},$$

$$\alpha_n(V) = 0.01(V + 10)\left[\exp\left(\frac{V + 10}{10}\right) - 1\right]^{-1},$$

$$\beta_n(V) = 0.125 \exp\left(\frac{V}{80}\right),$$

where

I = the membrane current density (μa/cm²) [inward current positive],
V = the membrane voltage (mV) [difference from resting potential, depolarization negative],
m = the sodium activation (dimensionless) [varying between 0 and 1],
h = the sodium inactivation (dimensionless) [varying between 0 and 1],
n = the potassium activation (dimensionless) [varying between 0 and 1],
t = time (msec),
C_0 = the membrane capacitance (μf/cm²),
$\bar{g}_{Na} = 120$, $\bar{g}_K = 36$, $\bar{g}_L = 0.3$ (m℧/cm²),
$V_{Na} = -115$, $V_K = 12$, $V_L = -10.5989$ (mV).

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[1] Readers unfamiliar with the electro-physiology of the nerve axon may wish to refer to J. W. Moore, "Electronic control of some active bioelectric membranes," PROC. IRE, vol. 47, pp. 1869–1880; November, 1959.

[2] A. L. Hodgkin and A. F. Huxley, "A quantitative description of membrane current and its application to conduction and excitation in nerve," *J. Physiol.*, vol. 117, pp. 500–544; August, 1952.

Several investigations of the H-H equations by the use of analog computers[3,4] and digital computers[5–8] have appeared in recent literature. Nevertheless, it is still difficult to make an electronic simulator of the H-H equations on account of their complexities.

III. The BVP Model

Recently, FitzHugh ingeniously simplified the H-H equations in case of a "space clamp," making use of an analog computer, and proposed the following BVP model (Bonhoeffer-van der Pol model).[9]

$$\begin{cases} J = \dfrac{1}{c}\dfrac{du}{dt} - w - \left(u - \dfrac{u^3}{3}\right), \\ c\dfrac{dw}{dt} + bw = a - u, \end{cases} \tag{2}$$

where a, b and c are constants satisfying the relations

$$1 > b > 0, \quad c^2 > b, \quad 1 > a > 1 - \tfrac{2}{3}b. \tag{3}$$

The variables u, w and J in the BVP model (2) correspond to the pair of variables (V, m), the pair of variables (h, n) and I in (1), respectively.

Fig. 1(a) shows the trajectories on the (u, w)-plane of the BVP model (2) in the case where $J=0$, while Fig. 1(b) shows the trajectories on the (u^*, w^*)-plane of (1) in the case where $I=0$. Here,

$$u^* = V - 36m, \qquad w^* = \tfrac{1}{2}(n - h). \tag{4}$$

Comparing Fig. 1(a) to Fig. 1(b), it is observed that there is fairly good correspondence between the H-H equations and the BVP model. We shall therefore try to simulate (2) instead of (1).

IV. An Electronic Simulation of the BVP Model

Let us consider the two-terminal circuit of Fig. 2, where TD is a tunnel diode.[10] From Kirchhoff's law

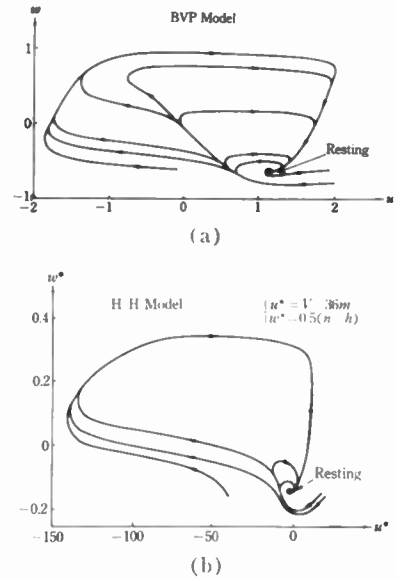


Fig. 1—The qualitative similarity of (a) and (b) suggests that the BVP model can be considered as belonging to the same class of excitable systems as the Hodgkin-Huxley model. (Reproduced from FitzHugh,[9] courtesy of the author.)

$$\begin{cases} j = C\dfrac{dv}{d\tau} - i - f(e), \\ L\dfrac{di}{d\tau} + Ri = -v = e - E_0, \end{cases} \tag{5}$$

where $f(e)$ is a function, as shown in Fig. 3, which represents the voltage vs current characteristic of the tunnel diode. For simplicity, we assume

$$f(e) = i_0 - \frac{1}{\rho}\left\{(e - e_0) - \frac{(e - e_0)^3}{3K^2}\right\} \quad (\rho > 0,\ K > 0), \tag{6}$$

where

$$i_0 = f(e_0).$$

By the introduction of new variables

$$t = \frac{\tau}{\sqrt{LC}}, \quad u = \frac{v + (e_0 - E_0)}{K}, \quad w = \frac{\rho}{K}(i + i_0),$$

$$J = \frac{\rho}{K}j, \quad a = \frac{Ri_0 + (e_0 - E_0)}{K}, \quad b = \frac{R}{\rho},$$

$$c = \frac{1}{\rho}\sqrt{\frac{L}{C}},$$

it is seen that (5) is reduced to (2).

Next, we shall examine the conditions in (3) for the constants a, b and c.

The first condition: $1 > b > 0$ is equivalent to

$$\rho > R. \tag{7}$$

The second condition: $c^2 > b$ is equivalent to

$$\frac{L}{R} > \rho C. \tag{8}$$

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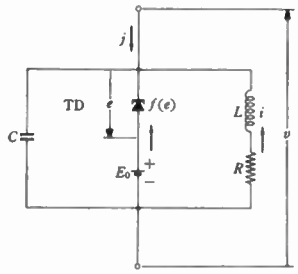


Fig. 2—An electronic simulator of the BVP model.

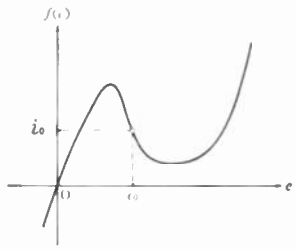


Fig. 3—The voltage-current characteristic of the tunnel diode.

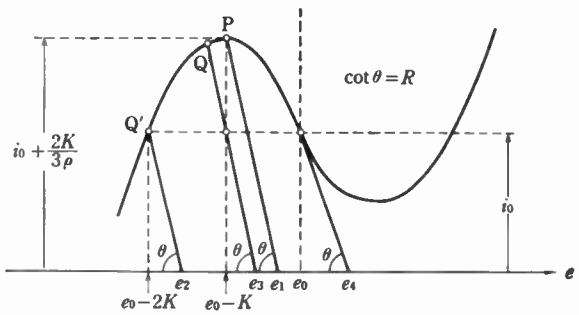


Fig. 4—When bias voltage E_0 is set between e_1 and e_2, the circuit shown in Fig. 2 is monostable. Whereas, if $E_0=e_4$, the circuit oscillates spontaneously.

The last condition: $1>a>1-\frac{2}{3}b$ is equivalent to

$$Ri_0+(e_0-K)<E_0<R\left(i_0+\frac{2K}{3\rho}\right)+(e_0-K), \quad (9)$$

or

$$e_3<E_0<e_1, \quad (10)$$

in Fig. 4.

The conditions in (7), (8) and (10) imply that in the case of the "current clamp," $j=0$, the circuit shown in Fig. 2 is nothing but the well-known monostable circuit of the tunnel diode. In particular, the condition in (8) means that the circuit oscillates spontaneously if the bias voltage is set at e_4 in Fig. 4. In fact, if $E_0=e_4$, then $a=0$, and in the case of $j=0$, (2) becomes as follows:

$$\frac{d^2u}{dt^2}-c\left\{\left(1-\frac{b}{c^2}\right)-u^2\right\}\frac{du}{dt}+\left\{(1-b)u+\frac{b}{3}u^3\right\}=0. \quad (11)$$

Since $c^2>b$, $1>b>0$, $c>0$, this differential equation belongs to the Liénard type[11] which represents self-oscillations.

Therefore it is concluded that the BVP model can be simulated by the monostable circuit shown in Fig. 2. However, condition (10) which restricts the bias voltage E_0 is too severe; since even if the bias voltage is set as

$$e_2<E_0<e_1, \quad (12)$$

in Fig. 4, the circuit in Fig. 2 remains monostable. The condition in (12) is written as

$$2>a>1-\tfrac{2}{3}b,$$

and hence we shall employ the conditions

$$1>b>0, \quad c^2>b, \quad 2>a>1-\tfrac{2}{3}b, \quad (13)$$

instead of (3), for the constants in the BVP model.

V. Propagation of Excitation

The equations which describe the propagation of the excitation along the nerve axon are easily obtained from the equations of the "space clamp."[12]

Denoting the distance along the nerve axon by S (cm), the radius of the nerve axon by R_0 (cm) and the specific resistance of the axoplasm by r_0 (KΩcm), we find that

$$I=\frac{R_0}{2r_0}\frac{\partial^2V}{\partial S^2}. \quad (14)$$

Consequently, taking (1) into account, the H-H partial differential equations describing the propagation of the excitation become as follows:

$$\begin{cases}\dfrac{\partial^2V}{\partial S^2}=\dfrac{2r_0}{R_0}\left\{C_0\dfrac{\partial V}{\partial t}+\bar{g}_{Na}m^3h(V-V_{Na})+\bar{g}_Kn^4(V-V_K)+\bar{g}_L(V-V_L)\right\},\\ \dfrac{\partial m}{\partial t}+\{\alpha_m(V)+\beta_m(V)\}m=\alpha_m(V),\\ \dfrac{\partial h}{\partial t}+\{\alpha_h(V)+\beta_h(V)\}h=\alpha_h(V),\\ \dfrac{\partial n}{\partial t}+\{\alpha_n(V)+\beta_n(V)\}n=\alpha_n(V).\end{cases} \quad (15)$$

Similarly, let us consider the circuit shown in Fig. 5(a) or Fig. 5(b), which is constructed by cascading the many two-terminal circuits of Fig. 2 through interstage coupling resistances.

[11] S. Lefschetz, "Differential Equations: Geometric Theory," Interscience Publishers, New York, N. Y., pp. 249–254; 1957.
[12] Hodgkin and Huxley, *op. cit.*, p. 522.

Regarding the above circuit as a distributed line, we have

$$j = \frac{1}{r}\frac{\partial^2 v}{\partial s^2}, \tag{16}$$

corresponding to (14), provided the interstage coupling resistance per unit length of the line is r. From (2) and (16), it follows that

$$\begin{cases} h\dfrac{\partial^2 u}{\partial s^2} = \dfrac{1}{c}\dfrac{\partial u}{\partial t} - w - \left(u - \dfrac{u^3}{3}\right), \\ c\dfrac{\partial w}{\partial t} + bw = a - u, \end{cases} \tag{17}$$

where $h = \rho/r$.

The system of partial differential equations (17) is the distributed BVP model which may be considered as a simplified system of the H-H partial differential equations (15).

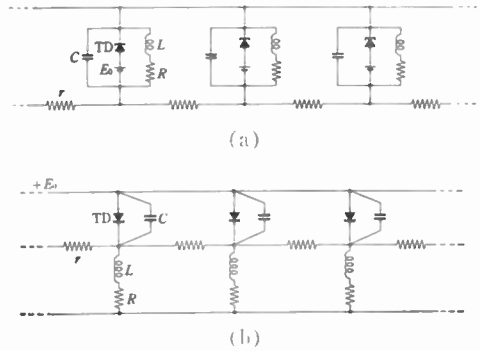


Fig. 5—A distributed active line is obtained by cascading the two-terminal circuits in Fig. 2 through interstage coupling resistances. The circuit shown in (b) is equivalent to that in (a).

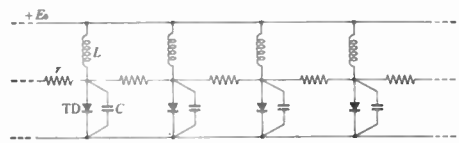


Fig. 6—Electronic model corresponding to (22).

By eliminating w from (17), a single partial differential equation for u is obtained as follows:

$$ch\frac{\partial^3 u}{\partial t \partial s^2} + bh\frac{\partial^2 u}{\partial s^2} = \frac{\partial^2 u}{\partial t^2} - c\left\{\left(1 - \frac{b}{c^2}\right) - u^2\right\}\frac{\partial u}{\partial t} + \left\{(1-b)u + \frac{b}{3}u^3\right\} - a. \tag{18}$$

For simplicity, let us consider a case where $R=0$ $(b=0)$. In such a case, (18) takes the form

$$ch\frac{\partial^3 u}{\partial t \partial s^2} = \frac{\partial^2 u}{\partial t^2} - c(1-u^2)\frac{\partial u}{\partial t} + u - a, \tag{19}$$

where

$$c > 0, \qquad 2 > a > 1, \qquad h > 0. \tag{20}$$

By setting

$$x = \frac{s}{\sqrt{ch}}, \quad z = \frac{2a}{a^2-1}(a-u), \quad \mu = c(a^2-1), \quad \epsilon = \frac{a^2-1}{4a^2}, \tag{21}$$

(19) is transformed as follows:

$$\boxed{\frac{\partial^3 z}{\partial t \partial x^2} = \frac{\partial^2 z}{\partial t^2} + \mu(1 - z + \epsilon z^2)\frac{\partial z}{\partial t} + z, \quad \mu > 0, \quad \frac{3}{16} > \epsilon > 0.} \tag{22}$$

It seems likely that the partial differential equation (22) is one of the simplest mathematical models of the nerve axon. The structure of the electronic model which corresponds to (22) is shown in Fig. 6.

VI. A Boundary-Value Problem

We now proceed to investigate the following boundary-value problem for the partial differential equation (22) (Fig. 7).

$$\begin{cases} z = z(x, t), \ x \geqq 0, \ t \geqq 0, \\ \dfrac{\partial^3 z}{\partial t \partial x^2} = \dfrac{\partial^2 z}{\partial t^2} + \mu(1 - z + \epsilon z^2)\dfrac{\partial z}{\partial t} + z, \\ \mu > 0, \ \dfrac{3}{16} > \epsilon > 0, \\ \text{on the line } t = 0, \ z = 0, \ \dfrac{\partial z}{\partial t} = 0, \\ \text{on the line } x = 0, \ z = F(t): \text{given}. \end{cases} \tag{23}$$

Some results of the numerical calculation using a digital computer are given below. We chose $\mu = 10$, $\epsilon = 0.1$ and

$$F(t) = \frac{z_0}{2}\left(1 - \cos\frac{2\pi t}{t_0}\right), \qquad t_0 \geqq t \geqq 0,$$
$$= 0, \qquad t \geqq t_0. \tag{24}$$

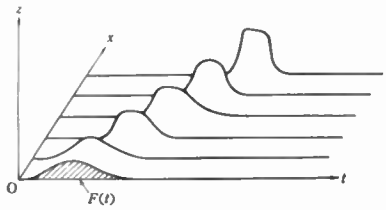


Fig. 7—A schematic display of the boundary-value problem (23).

A. Signal Amplification

The case of $t_0=3.0$, $z_0=5.0$ is shown in Fig. 8. In this case the signal is amplified in the course of transmission.

B. Signal Attenuation

The case of $t_0=1.0$, $z_0=20.0$ is shown in Fig. 9. In this case the signal is attenuated in the course of transmission.

C. Signal Elimination

The case of $t_0=3.0$, $z_0=3.0$ is shown in Fig. 10. In this case the signal is eliminated in the course of transmission.

From these results, it is expected that with respect to the signal height (there being an asymptotic value and a threshold value),

1) a signal with a height between the asymptotic value and the threshold value is amplified during transmission,
2) a signal higher than the asymptotic value is attenuated to the asymptotic value during transmission,
3) a signal lower than the threshold value is eliminated during transmission (see Fig. 11).

These actions may be called "shaping" with respect to signal height. In order to clarify the situation, we shall try to seek for the asymptotic waveform in Section VII.

VII. Asymptotic Waveform

If the partial differential equation (22) has a waveform, as its solution, which is transmitted along a line without suffering distortion and with a constant velocity (say θ), then the solution must be a function of $\tau=t-x/\theta$. In such a case the substitution

$$\xi(\tau) \equiv z(x, t), \qquad \tau = t - \frac{x}{\theta}$$

reduces the partial differential equation (22) to an ordinary differential equation for ξ:

$$\beta\xi''' - \xi'' - \mu(1 - \xi + \epsilon\xi^2)\xi' - \xi = 0, \qquad (25)$$

where $\xi'=d\xi/d\tau$ and

$$\beta = \theta^{-2} > 0. \qquad (26)$$

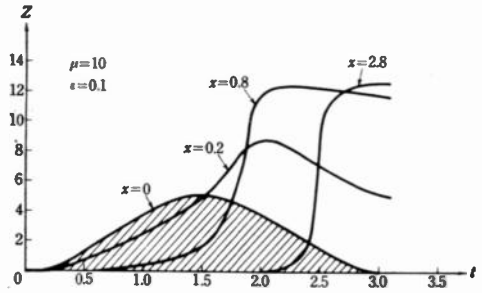


Fig. 8—A signal above the threshold value and below the asymptotic value is amplified during transmission.

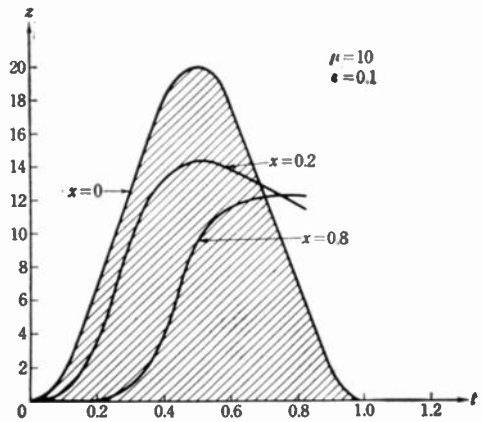


Fig. 9—A signal above the asymptotic value is attenuated during transmission.

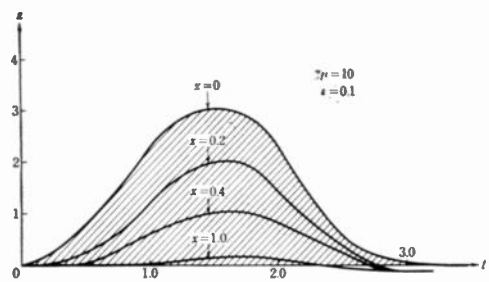


Fig. 10—A signal below the threshold value is eliminated during transmission.

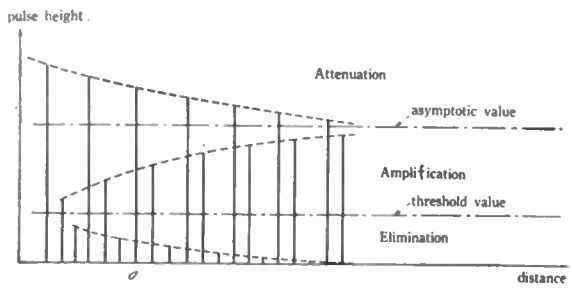


Fig. 11—Signals are amplified or attenuated or eliminated according to their heights.

For the time being, β remains an unknown constant. In fact, its determination constitutes a part of the object of the following procedure. Now, it is obvious that (25) has a resting solution $\xi \equiv 0$. If (25) has a solution, except for $\xi \equiv 0$, such as $\xi(\tau) \to 0$ when $\tau \to \pm \infty$ for some β, then the solution corresponds to the asymptotic waveform which we are seeking, and the transmission velocity of the asymptotic waveform is determined from β.

The behavior of $\xi(\tau)$ in the neighborhood of the resting solution is determined approximately by a linear differential equation

$$\beta\xi''' - \xi'' - \mu\xi' - \xi = 0. \tag{27}$$

The characteristic equation of (27)

$$H(\lambda) \equiv \beta\lambda^3 - \lambda^2 - \mu\lambda - 1 = 0 \tag{28}$$

has one real positive root, since $H(0) = -1 < 0$, $H'(0) = -\mu < 0$ and $H(+\infty) > 0$. By denoting this root as $\lambda_0 (\lambda_0 > 0)$, $H(\lambda)$ is factorized as

$$H(\lambda) = (\lambda - \lambda_0)(\beta\lambda^2 + \gamma\lambda + \lambda_0^{-1}),$$

where

$$\gamma = \beta\lambda_0 - 1 = \frac{\mu}{\lambda_0} + \left(\frac{1}{\lambda_0}\right)^2.$$

Since $\gamma > 0$, it is apparent that both of the other two roots are either real negative or complex conjugate with negative real parts.

From a more precise examination it is easily seen that when $\mu > 2$ and $\beta < \beta_0(\mu)$, they are real negative; otherwise they are complex conjugate with negative real parts, where

$$\beta_0(\mu) = \frac{1}{27}(2\mu^2 - 9\mu + 2\sqrt{\mu^4 - 9\mu^3 + 27\mu^2 - 27}).$$

Now, if $\xi(\tau) \to 0$ when $\tau \to -\infty$, as shown in Fig. 12, then $\xi(\tau) \sim Ae^{\lambda_0\tau}$ (A is an arbitrary constant), when τ has a negative large value. Therefore we have

$$\xi'(\tau) \sim A\lambda_0 e^{\lambda_0\tau}, \qquad \xi''(\tau) \sim A\lambda_0^2 e^{\lambda_0\tau}.$$

We shall perform numerical calculations for the third-order ordinary differential equation (25) with the following initial conditions:

$$\begin{cases} \xi(0) = \Delta, \\ \xi'(0) = \lambda_0\Delta, \\ \xi''(0) = \lambda_0^2\Delta, \end{cases}$$

where $\Delta (\Delta > 0)$ is the smallest step of ξ. It is noted that λ_0 depends on β.

In general, the solution of (25) corresponding to the initial conditions in (29) depends on the value of β; and we are seeking for exactly the β for $\xi(\tau) \to 0$ when $\tau \to +\infty$ as shown in Fig. 12.

When $\mu = 3.0$, $\epsilon = 0.1$, two such values of β were expected to exist from the results of digital computations. One of them is $\beta \doteqdot 0.44488$, which corresponds to a stable asymptotic waveform (Fig. 13), and this is what we are aiming at. The other is $\beta \doteqdot 0.938$ and corresponds to an unstable asymptotic waveform (Fig. 14), which is a critical signal traveling down the line along the threshold and being physically unrealizable.

Because the stable asymptotic waveform is peculiar to this line, it may be expected that the line will show a waveform shaping action, in respect not only to the signal height but also to the signal width.

Fig. 15(a) shows the relation between μ and signal propagation velocity θ in the case of $\epsilon = 0.1$. The velocity of the stable asymptotic waveform seems to be a monotonically increasing function of μ.

Since

$$\frac{\delta s}{\delta t} = \sqrt{ch}\,\theta = \frac{1}{\sqrt{r}}\sqrt[4]{\frac{L}{C}}\,\theta\left(\frac{(e_0 - E_0)^2 - K^2}{\rho K^2}\cdot\sqrt{\frac{L}{C}}\right), \tag{30}$$

the propagation velocity is inversely proportional to $r^{1/2}$. Furthermore, it may be concluded that the velocity increases with increasing L and/or decreasing C, provided $\theta(\mu)$ is monotonically increasing.

Fig. 15(b) shows the relation between μ and pulse height in case of $\epsilon = 0.1$. The shape of the stable asymptotic waveform approaches to a rectangular pulse as μ increases, although its height hardly depends on μ.

It may be expected that the stable asymptotic waveform coincides with the unstable one at $\mu = 2$.

VIII. Experiments

We fabricated a lumped constant cascaded circuit with nine stages as shown in Fig. 16 and the expected results were obtained. In particular, it was observed that the shaping action in respect to the signal height was completed at the first stage, while the shaping action in respect to the signal width was very weak in comparison with the former (Fig. 17).

In order to make the shaping action in respect to pulse height more moderate, it is effective to add some inductance to each interstage coupling resistance in series. The circuit (Fig. 18, page 2068), in the case of $C = 0.05$ μf, shapes traveling signals with respect to the height (Fig. 19). On the other hand, Fig. 20 illustrates the shaping action with respect to the width of the circuit shown in Fig. 18, in the case of $C \doteqdot 20$ pf (junction capacitance of TD).

Since this line has a symmetrical structure, the signal transmission is bidirectional, that is, a signal applied to the right (left) end is transmitted to the left (right) and a signal applied to the middle of the line is transmitted in both directions. An interesting phenomenon is that two signals traveling in opposite direction from both ends vanish at the collision point.

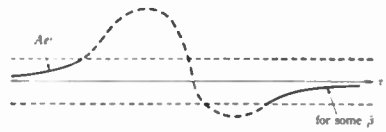


Fig. 12—If (25) has a solution, except for $\xi(\tau)\equiv 0$, such as $\xi(\tau)\to 0$ when $\tau\to+\infty$ for some β, then the solution corresponds to the sought-for asymptotic waveform, and the transmission velocity of the asymptotic waveform is determined from the β.

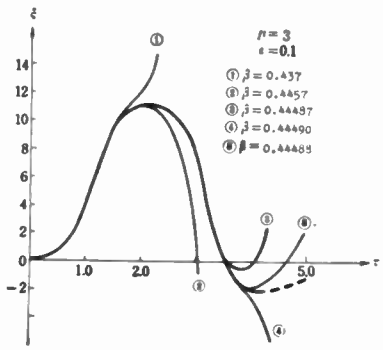


Fig. 13—For $\beta\doteqdot 0.44488$, it is expected that a stable asymptotic waveform exists.

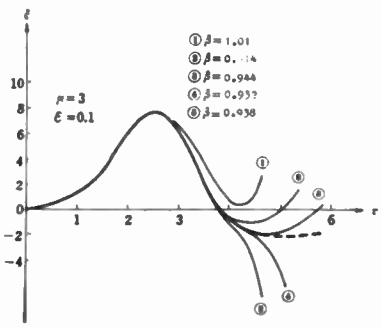
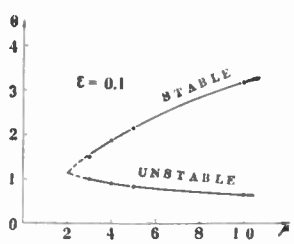
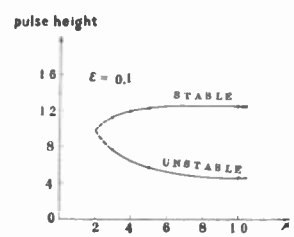


Fig. 14—For $\beta\doteqdot 0.938$, it is expected that an unstable asymptotic waveform exists, which travels down the line along the threshold and is physically unrealizable.

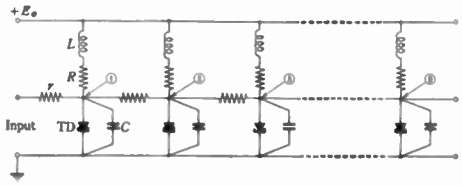


(a)

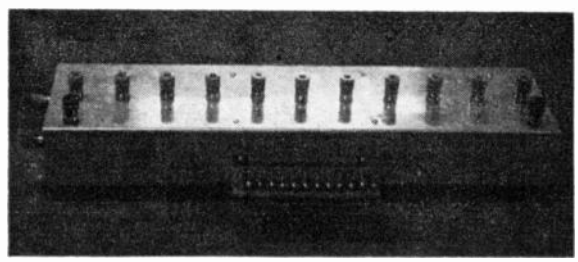


(b)

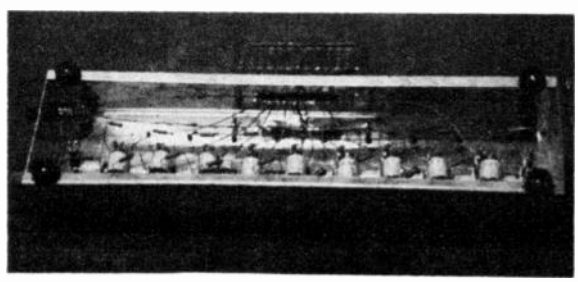
Fig. 15—The relation between μ and the signal propagation velocity and the relation between μ and pulse height in case of $\epsilon=0.1$. It seems that the stable asymptotic waveform coincides with the unstable one at $\mu=2$. (a) The velocity of the stable asymptotic waveform seems to be a monotonically increasing function of μ. (b) The pulse height of the stable asymptotic waveform hardly depends on μ.



(a)

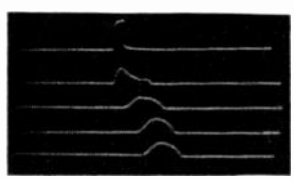


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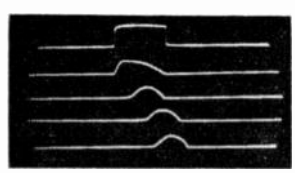


(c)

Fig. 16—The circuit used in our experiments. $L=4$ mh, $R=70\ \Omega$, $r=500\ \Omega$, $E_0=100$ mv, $C=0.01\ \mu$f.



(a)



(b)

Fig. 17—The circuit in Fig. 16 shapes a signal waveform during transmission. In each figure, the first waveform is that of the input signal; the second, third, fourth and the fifth waveforms are those observed at the first, fourth, seventh and the ninth stages in the circuit in Fig. 16(a), respectively. In both cases, the signals achieve the same waveform. (a) A narrow signal (6 μsec) is widened in the course of transmission. (b) A wide signal (32 μsec) is narrowed in the course of transmission.

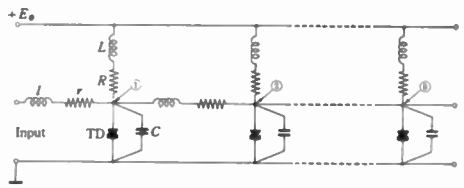
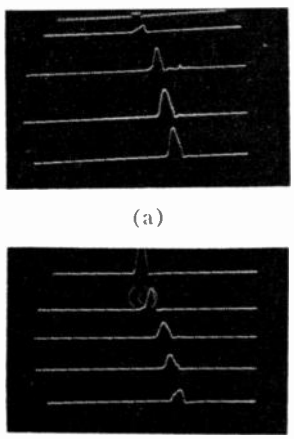


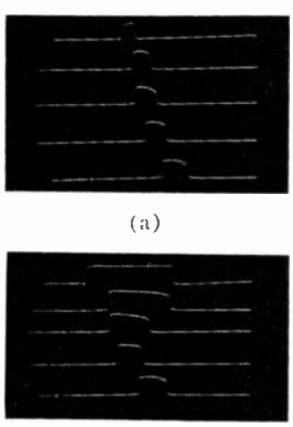
Fig. 18—By adding some inductance to each coupling resistance in series, the shaping action can be made more moderate.



(a)

(b)

Fig. 19—Waveform shaping action with respect to the signal height of the circuit in Fig. 18 displayed on a CRO screen. $L=l=4$ mh, $R=115\ \Omega$, $r=200\ \Omega$, $E_0=150$ mv, $C=0.05\ \mu$f. In each figure, the waveform in the first, second, third, fourth and fifth lines are those observed at the first, third, fifth, seventh and ninth stages in the circuit shown in Fig. 18, respectively. Pulse heights in the bottom lines in (a) and (b) are equal and 150 mv. [Note that vertical scales are different in (a) and (b)]. (a) A small signal with 30-mv height is amplified in course of transmission. (b) A large signal with 380-mv height is attenuated in course of transmission.



(a)

(b)

Fig. 20—Waveform shaping action with respect to the signal width of the circuit in Fig. 18 displayed on a CRO screen. In this case, $C\doteqdot 20$ pf, other constants are the same as those in Fig. 19. The correspondence of lines and stages is the same as that in Fig. 19. Pulse width in the bottom lines of (a) and (b) is equal—10 μsec. (a) A narrow signal with 5-μsec width is widened in the course of transmission. (b) A wide signal with 30-μsec width is narrowed in the course of transmission.

IX. Related Problems

A. The Neuristor

Crane proposed a novel device termed neuristor.[13] A neuristor may be visualized as a one-dimensional channel along which signals may flow. The mode of signal propagation is somewhat analogous to that which occurs along a nerve axon. The neuristor may be used to synthesize all digital logic functions, so that any digital logic system can be realized using arrays of neuristors only. Neuristors are interconnected in two basically different modes—T junctions and S junctions.

Our active pulse transmission line can lead to an electronic realization of the neuristor. The T junction is readily realized by connecting three active pulse transmission lines at a point. The S junction is realized by the circuit in Fig. 21. Corresponding to (22), the differential equations which represent signal propagation along the S junction are as follows:

$$\begin{cases} \dfrac{\partial^3 z_1}{\partial t \partial x^2} = \dfrac{\partial^2 z_1}{\partial t^2} + \mu(1 - z_1 + \epsilon z_1^2)\dfrac{\partial z_1}{\partial t} \\ \qquad\qquad + \dfrac{1}{1-k^2} z_1 + \dfrac{k}{1-k^2} z_2, \\ \dfrac{\partial^3 z_2}{\partial t \partial x^2} = \dfrac{\partial^2 z_2}{\partial t^2} + \mu(1 - z_2 + \epsilon z_2^2)\dfrac{\partial z_2}{\partial t} \\ \qquad\qquad + \dfrac{1}{1-k^2} z_2 + \dfrac{k}{1-k^2} z_1, \end{cases} \tag{31}$$

where $k=M/L$, $0<k<1$.

The storage ring of the neuristor was realized by making a loop with forty stages of the monostable circuit and five stages of the S junction (Fig. 22).

Cote has published an active pulse transmission line using p-n-p-n diodes toward an electronic realization of the neuristor.[14] The use of current-controlled negative resistance elements, however, seems to make the circuit more complicated.

B. Active Surface

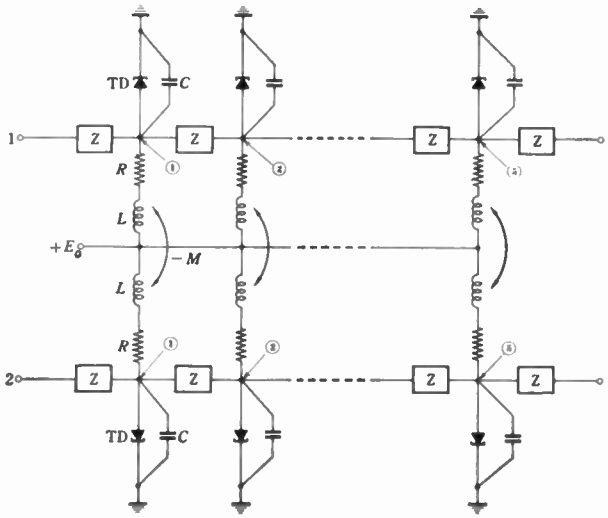
Our active pulse transmission line may easily be extended to an active surface with the structure shown in Fig. 23. The differential equation of this active surface, corresponding to (22), becomes

$$\frac{\partial^3 z}{\partial t \partial x^2} + \frac{\partial^3 z}{\partial t \partial y^2} = \frac{\partial^2 z}{\partial t^2} + \mu(1 - z + \epsilon z^2)\frac{\partial z}{\partial t} + z, \tag{32}$$

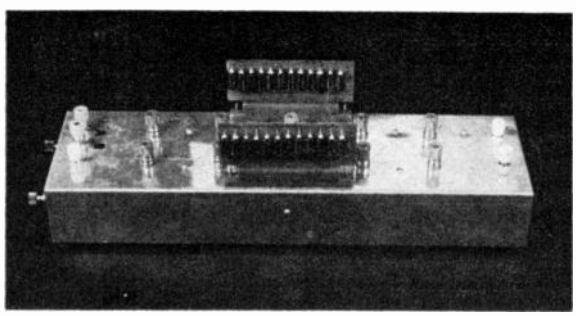
$$\mu > 0, \qquad \frac{3}{16} > \epsilon > 0.$$

[13] H. D. Crane, "The neuristor," IRE Trans. on Electronic Computers, vol. EC-9, pp. 370–371, September, 1960; also, "Neuristor Studies," Solid-State Electronics Lab., Stanford University, Stanford, Calif., Tech. Rept. No. 1506-2, July, 1960; also H. D. Crane, "Neuristor—a novel device and system concept," this issue, p. 2048.

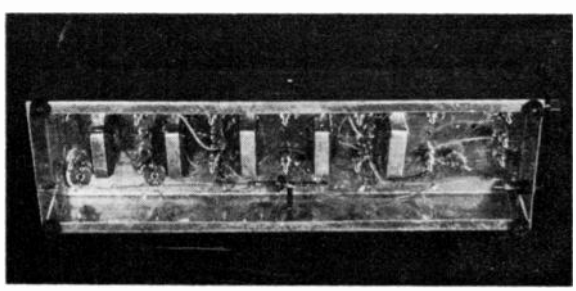
[14] A. J. Cote, Jr., "A neuristor prototype," Proc. IRE (*Correspondence*), vol. 49, pp. 1430–1431; September, 1961.



(a)



(b)



(c)

Fig. 21—An S junction of the neuristor can be realized by coupling two monostable circuits by a mutual inductance. $L=358$ mh, $M=355$ mh, $R=70\ \Omega$, $Z=r+pl$, $l=4$ mh, $r=70\ \Omega$, $E_0=70$ mv, $C=0.01\ \mu$f.

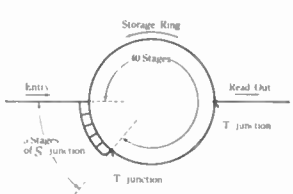


Fig. 22—Signals enter from the "entry" and continue to circulate along the "storage ring." The contents of the storage ring can be obtained from the "read out" line.

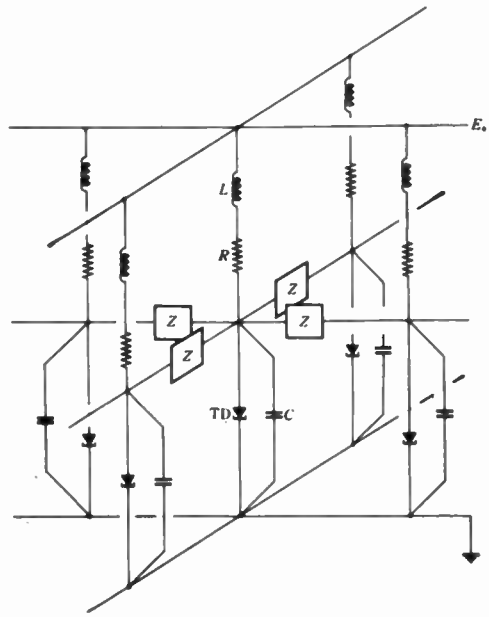


Fig. 23—The structure of the active surface represented by (32).

X. Conclusions

To electronically simulate the distributed BVP model, which may be considered as a simplified Hodgkin-Huxley model describing the propagation of the excitation of the nerve axon, we have made an approximately distributed active pulse transmission line using tunnel diodes. The line behaves similarly to the living axon, but the representation is gross.

The line is constructed by cascading many monostable circuits and consists essentially of the following four parts:

1) transverse power voltage supply,
2) transverse inductance,
3) transverse tunnel diode,
4) longitudinal coupling resistance.

The active line shows the following characteristics in the transmission of signals:

1) There is a certain threshold value in respect to the signal height, and signals below the threshold (or noise) are eliminated during transmission.
2) The line shapes signal waveforms. Namely, there being a specific pulse-like waveform peculiar to this line, signals above the threshold approach it during transmission.
3) Since this line has a symmetrical structure, the signal transmission is bidirectional. Two signals traveling in opposite directions from both ends vanish at the collision point.

On account of the existence of the threshold and the shaping action, this line makes possible highly reliable pulse transmission, and will be useful for various kinds of information-processing systems.

Appendix

We shall briefly sketch the electrical behavior of the surface membrane of a nerve axon following Hodgkin and Huxley.[2]

The electrical behavior of the membrane in the case of a "space clamp" may be represented by the network shown in Fig. 24. Current can be carried through the membrane either by charging the membrane capacity (C_0) or by movement of ions through the resistances in parallel with the capacity.

The capacity current I_C is given by

$$I_C = C_0 \frac{dE}{dt} \cdot \tag{33}$$

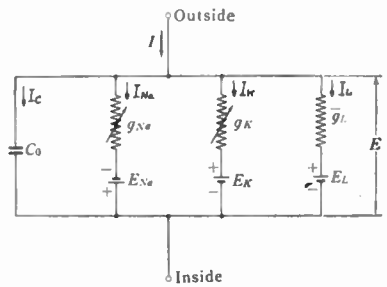


Fig. 24—Equivalent circuit representing the electric behavior of the membrane.

The ionic current is divided into components carried by sodium and potassium ions (I_{Na} and I_K) and a small "leakage current" (I_L) made up by chloride and other ions. Each component of the ionic current is determined by a driving force which may conveniently be measured as an electrical potential difference and a permeability coefficient which has the dimensions of a conductance. Thus the sodium current I_{Na} is equal to the sodium conductance (g_{Na}) multiplied by the difference between the membrane potential E and the equilibrium potential for the sodium ion E_{Na}:

$$I_{Na} = g_{Na}(E - E_{Na}). \tag{34}$$

Similarly we have

$$I_K = g_K(E - E_K), \tag{35}$$

$$I_L = \bar{g}_L(E - E_L). \tag{36}$$

Eqs. (34)–(36) may be rewritten

$$I_{Na} = g_{Na}(V - V_{Na}), \tag{37}$$

$$I_K = g_K(V - V_K), \tag{38}$$

$$I_L = \bar{g}_L(V - V_L), \tag{39}$$

where

$$V = E - E_R, \quad V_{Na} = E_{Na} - E_R, \quad V_K = E_K - E_R,$$

$$V_L = E_L - E_R,$$

and E_R is the absolute value of the resting potential. V, V_{Na}, V_K, and V_L can then be measured directly as displacements from the resting potential.

We then assume that the sodium conductance g_{Na} is represented by

$$g_{Na} = \bar{g}_{Na} m^3 h, \tag{40}$$

$$\frac{dm}{dt} = \alpha_m(1 - m) - \beta_m m, \tag{41}$$

$$\frac{dh}{dt} = \alpha_h(1 - h) - \beta_h h, \tag{42}$$

where $\bar{g}_{Na}$ is a constant with the dimensions of conductance/cm^2; α's and β's are rate constants which vary with voltage (V) but not with time (t) and have dimensions of [time]$^{-1}$; m and h are dimensionless variables which can vary between 0 and 1.

These equations may be given a physical basis if sodium conductance is assumed to be proportional to the number of sites on the inside of the membrane which are occupied simultaneously by three activating molecules but are not blocked by an inactivating molecule. m then represents the proportion of activating molecules on the inside and $1-m$ the proportion on the outside; h is the proportion of inactivating molecules on the outside and $1-h$ the proportion on the inside. α_m or β_h and β_m or α_h represent the transfer rate constants in the two directions.

The forms of α's and β's as functions of V are determined experimentally as shown in (1).

Similar equations for g_K are

$$g_K = \bar{g}_K n^4, \tag{43}$$

$$\frac{dn}{dt} = \alpha_n(1 - n) - \beta_n n. \tag{44}$$

Summing up these equations, we have

$$\begin{aligned} I &= I_C + I_{Na} + I_K + I_L \\ &= C_0 \frac{dV}{dt} + \bar{g}_{Na} m^3 h(V - V_{Na}) + \bar{g}_K n^4 (V - V_K) \\ &\quad + \bar{g}_L(V - V_L), \end{aligned} \tag{45}$$

as shown in (1).

Acknowledgment

The authors appreciate the aid of Dr. S. Nakajima, the aid of Prof. M. Takata and T. Shimizu in the digital computation and the aid of R. Suzuki and Messrs. Saruyama, Takeya and Nozawa in the experiments. The authors also thank Profs. H. Takahashi and S. Furuya for their valuable discussions. This research was supported by a research grant from the Asahi Press, for which we express our deep appreciation.

The Analytic Signal Representation of Modulated Waveforms*

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Summary—**This report offers a rather general and mathematically convenient formulation of analog-modulated signals which makes use of the analytic signal concept. Known types of modulation are readily identified as special cases. As a result of examining the various cases which the model embraces, a new type of modulation has been discovered—single sideband frequency modulation (SSB FM)—which can be derived from a conventional phase-modulated signal by an additional amplitude modulation, using the exponential function of the modulating signal's Hilbert transform. The resulting modulated signal will have a one-sided spectrum about the carrier frequency, will be compatible with existing FM receivers, and will cause a decrease in signal bandwidth.**

Introduction

RECENT STUDIES of compatible single sideband (SSB) modulation systems have stimulated interest in the theoretical aspects of simultaneous amplitude and phase modulation [1]–[4].[1] These studies are notable for at least two reasons which are not connected with the SSB application. First, they underscore the fact that although amplitude and phase modulation are distinct techniques, they have an intimate connection. Second, theoretical studies invariably introduce the Hilbert transform [5], [6] sooner or later and become, in fact, applications of the "analytic signal" [7], [8].

Other studies dealing with various types of signal and noise waveforms [9]–[12] further attest to the growing utility of the analytic signal representation as a theoretical device. This paper considers the application of the analytic signal to Baghdady's representation of analog modulation [13] in order to obtain a more general formulation which embraces known types of analog modulation and in which signals having one-sided spectra are implicit.

The Analytic Signal

Basically, the analytic signal is a complex function of a real variable whose real and imaginary parts are a Hilbert pair. In practice, the actual waveform under consideration is identified with either the real or the imaginary part of the analytic signal; the analytic signal is then simply substituted for the actual signal for the purpose of analysis.

Frequently the result is a more convenient and compact notation which can be used in convolution integrals or in conjunction with transfer functions in the same fashion as the real signal and which, in addition, has several interesting and useful properties:

1) The Fourier transform of the analytic signal vanishes for negative frequencies while that of its complex conjugate vanishes for positive frequencies.[2]
2) The real and the imaginary parts of the analytic signal have identical autocorrelation functions and power spectra.
3) The magnitude and phase of the analytic signal serve to define mathematically an envelope and phase for the actual waveform. In practice, the envelope and phase of an actual waveform have physical significance only for narrow-band signals and cannot be measured precisely except for pure sinusoids.

It is clear from the foregoing that the analytic signal is simply a formalized version of the "rotating vector" or "phasor" that is frequently used in circuit analysis and studies of communication systems.[3] The theoretical justification of this representation is gratifying because it brings to light some of the otherwise unsuspected properties listed above as well as eliminating the need for its intuitive introduction. Details of these properties are given in [5], [7]–[12], [15] and [16].

The Modulated Signal

While the actual processes employed in generating an analog-modulated signal may differ considerably for various types of modulation, the resulting signal appears representable as the product of two time functions, *viz.*,

$$s(t) = c(t)m(f(t)), \tag{1}$$

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† Electronics Department, RAND Corporation, Santa Monica, Calif.

[1] This list of references is by no means complete. The first three are merely the most recent which have come to the author's attention. Lyannoy [4] reviews the previous literature giving numerous references, and asserts that the notion of simultaneous amplitude and phase modulation originated with Tetel'baum in 1937; Taylor [1] credits Barkhausen, also in 1937. Additional references are given by Kahn [3].

[2] The question of the reality of positive vs negative frequencies is seen to be academic. The presence of only one or the other appears related to the fact that the analytic signal and its complex conjugate always rotate in opposite directions in the complex plane.

[3] The date at which the rotating vector was introduced is difficult to determine precisely. Many of the modern concepts associated with analysis of FM and SSB systems are given by Wheeler [14].

where c is a function representing the carrier and m is the modulation functional representing an operation on the modulating signal f. The form of (1) has the advantage of mathematically disassociating the operation of providing a carrier from that of characterizing the actual modulation technique.

The Carrier Function

The purpose of the carrier function in (1) is to transfer the intelligence spectrum to a frequency region which is more suitable for propagation. The effect on the resulting spectrum is seen from the Fourier transform of the product [17] in (1),

$$S(\omega) = \int_{-\infty}^{\infty} C(\mu) M(\omega - \mu) d\mu \qquad (2)$$

If the carrier is a narrow-band waveform as might be employed for noise modulation, then the convolution expressed in (2) results in a spreading of the signal band as well as its translation to the vicinity of the carrier frequency. If the carrier is a pure sinusoid, then a simple frequency translation is effected and (2) becomes

$$S(\omega) = \tfrac{1}{2}[M(\omega - \omega_0) + M(\omega + \omega_0)]. \qquad (3)$$

If the carrier is written as an analytic signal whose spectrum, therefore, exists only for positive frequencies, then the frequency transfer is only toward the positive frequencies as illustrated in Figs. 1(a) and 1(b) for $\omega_0 > 0$. Whether or not $s(t)$ is in turn analytic depends on the spectral extent of both $c(t)$ and $m(t)$, the test being whether or not $S(\omega)$ vanishes for ω negative.

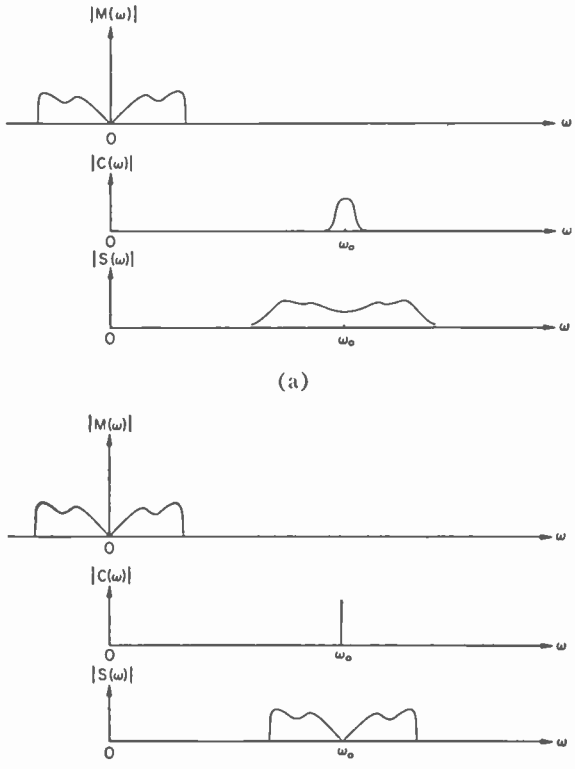


Fig. 1—Spectral translations due to introduction of analytic carrier. (a) General (band limited) analytic carrier. (b) Harmonic analytic carrier, $e^{i\omega_0 t}$.

The Modulation Function

Usable modulation functions must be limited, of course, to those which yield modulated signals amenable to subsequent demodulation. The techniques currently available are: *coherent* (amplitude) detection in which a replica of the carrier function is used to render the amplitude of the modulation function, *phase* or *frequency* detection in which the oscillatory character of a narrow-band modulated signal permits an approximate measurement of its phase or frequency, and *envelope* detection in which a suitable rectifying and filtering network can yield an approximation to the envelope of a narrow-band modulated signal. Thus, the modulation function must cause the modulating signal to appear (though not necessarily linearly) in either the amplitude, phase /frequency, or envelope of the modulated signal. The linear and exponential functionals are unique in their ability to produce just these effects and are, therefore, the only ones considered here.

As might be suspected, the linear functional corresponds to amplitude modulation. The exponential functional, however, corresponds not only to the familiar phase modulation[4] but also to the emergent technique of simultaneous amplitude and phase modulation which may be better characterized as envelope modulation. The latter will be discussed in greater detail in a subsequent section.

If the modulation function is written for the analytic as well as the purely real modulating signals, then both the SSB and double sideband (DSB) forms of the modulated signal appear. The SSB form results, of course, from the fact that an analytic function m of an analytic function f is itself analytic, and therefore has a spectrum which vanishes for negative frequencies. Frequency translation by the carrier function then results in a modulated signal which has only an upper sideband. A one-sided spectrum having only a lower sideband can be generated by using the complex conjugate of the analytic function f, since its spectrum vanishes for positive frequencies.

Harmonic-Carrier Systems

The modulated signals resulting from the foregoing modulation functions are best illustrated by using the harmonic analytic carrier $e^{i\omega_0 t}$ in (1). For each modulation function $m(t)$, a complex modulated signal $s(t)$ results; when analytic, both its real and imaginary parts are valid representations of the actual modulated signal, and its magnitude gives the envelope. The modulating

[4] No fundamental distinction need be made between phase and frequency modulation since they differ only by a linear operation (differentiation or integration) on the modulating signal. Frequency modulation can be obtained by phase modulating with the integral of the modulating signal and conversely using differentiation.

signal is denoted by $f(t)$ and its Hilbert transform by $\hat{f}(t)$.

Linear (Amplitude) Modulation

Double Sideband:

$$\begin{aligned} m(t) &= f(t) \\ s(t) &= f(t)\cos\omega_0 t + if(t)\sin\omega_0 t \\ |s(t)| &= f(t), \qquad \omega_0 \geq \omega_{\max}. \end{aligned} \tag{4}$$

The representation of a conventional amplitude-modulated signal[5] is immediately apparent in (4). It is analytic for $\omega_0 \geq \omega_{\max}$, where $\omega_{\max}$ is the highest angular frequency component in $f(t)$, as noted both by Oswald [9] and by Wozencraft [16]. Coherent detection can recover $f(t)$ exactly for all ω_0. Theoretically, the signal must be analytic to permit envelope detection; practically, it must also be narrow-band.

Single Sideband:

$$\begin{aligned} m(t) &= f(t) + i\hat{f}(t) \\ s(t) &= [f(t)\cos\omega_0 t - \hat{f}(t)\sin\omega_0 t] \\ &\quad + i[f(t)\sin\omega_0 t + \hat{f}(t)\cos\omega_0 t] \\ |s(t)| &= [f^2(t) + \hat{f}^2(t)]^{1/2}, \qquad \omega_0 > 0. \end{aligned} \tag{5}$$

That (5) represents a SSB amplitude-modulated signal is easily shown by noting that cos *at* and sin *at* are a Hilbert pair, so the Fourier series representations of $f(t)$ and $\hat{f}(t)$ are

$$f(t) = \sum_{n=0}^{\infty} c_n \cos(\omega_n t + \phi_n); \quad \hat{f}(t) = \sum_{n=0}^{\infty} c_n \sin(\omega_n t + \phi_n).$$

Thus, (5) can be written

$$\begin{aligned} s(t) &= \sum_{n=0}^{\infty} c_n \cos[(\omega_n + \omega_0)t + \phi_n] \\ &\quad + i\sum_{n=0}^{\infty} c_n \sin[(\omega_n + \omega_0)t + \phi_n] \end{aligned}$$

where the simple translation of all frequency components by an amount ω_0 is apparent. The representation is analytic for all values of $\omega_0 > 0$ and only coherent detection can recover $f(t)$ exactly.

Exponential (Phase) Modulation

Double Sideband:

$$\begin{aligned} m(t) &= e^{if(t)} \\ s(t) &= \cos[\omega_0 t + f(t)] + i\sin[\omega_0 t + f(t)] \\ |s(t)| &\cong 1, \qquad \omega_0 \gg \Omega_{\max}. \end{aligned} \tag{6}$$

The conventional representation of phase modulation is apparent in (6); it is designated as DSB to emphasize that the spectrum is two-sided about ω_0. Even if the modulating signal $f(t)$ is band-limited, the exponential operation used in generating $m(t)$ assures that it will have a spectrum of infinite extent in both positive and negative directions. Shifting the center frequency to ω_0 cannot, therefore, produce a spectrum containing only positive frequencies so, as noted also by Dugundji [11], the representation is not analytic. In a practical sense, of course, it is known that the spectrum falls off rapidly beyond some frequency. Thus, the representation can be taken approximately to be analytic if the center frequency is high compared with the highest significant frequency component $\Omega_{\max}$ in $m(t)$.

Single Sideband:

$$\begin{aligned} m(t) &= e^{i[f(t)+i\hat{f}(t)]} \\ s(t) &= e^{-\hat{f}(t)}\cos[\omega_0 t + f(t)] + ie^{-\hat{f}(t)}\sin[\omega_0 t + f(t)] \\ |s(t)| &= e^{-\hat{f}(t)}, \qquad \omega_0 > 0. \end{aligned} \tag{7}$$

The modulated signal given by (7) appears to be novel. Since the modulation function is now analytic, it contains no negative frequencies (though the spectrum still extends infinitely in the positive direction). Thus, $s(t)$ is analytic for all values of $\omega_0 > 0$.

By analogy with (5), the modulated signal given by (7) represents a SSB version of phase modulation since it contains no frequency components below ω_0. Though the instantaneous frequency

$$\omega_{\text{inst}} \equiv \frac{d\phi}{dt} = \omega_0 + \dot{f}(t)$$

has the same excursions for both (6) and (7), the multiplicative factor $e^{-\hat{f}(t)}$ causes the lower sideband to disappear in (7). To demonstrate that this is indeed the case and, also, to gain insight as to the effect on the upper sideband, consider frequency modulation by the single sinusoid

$$\dot{f}(t) = \Omega\cos\omega t$$

where Ω is the peak frequency deviation. The phase function and its Hilbert transform become

$$f(t) = \frac{\Omega}{\omega}\sin\omega t, \qquad \hat{f}(t) = -\frac{\Omega}{\omega}\cos\omega t$$

where Ω/ω is the deviation ratio or modulation index.

The modulation function in (7) becomes

$$m(t) = e^{i[f(t)+i\hat{f}(t)]} = \exp\left[\frac{\Omega}{\omega}e^{i\omega t}\right] = \sum_{k=0}^{\infty}\frac{1}{k!}\left(\frac{\Omega}{\omega}\right)^k e^{ik\omega t},$$

yielding the analytic modulated signal

$$m(t)e^{i\omega_0 t} = \sum_{k=0}^{\infty}\frac{1}{k!}\left(\frac{\Omega}{\omega}\right)^k e^{i(\omega_0 + k\omega)t}.$$

The actual modulated signal may be taken as the real part or

$$e^{-\hat{f}(t)}\cos[\omega_0 t + f(t)] = \sum_{k=0}^{\infty}\frac{1}{k!}\left(\frac{\Omega}{\omega}\right)^k \cos(\omega_0 + k\omega)t \tag{8}$$

[5] The carrier is suppressed in this representation, but the distinction is trivial since it is easily inserted by adding a constant to $m(t)$ in (4) and (5).

from which the one-sided nature of the spectrum is clear.[6] The mean-square value of the amplitude factor is $I_0[2(\Omega/\omega)]$, so the ratio of peak-to-average power over the modulation cycle becomes

$$e^{2(\Omega/\omega)} \Big/ I_0\left(2\,\frac{\Omega}{\omega}\right).$$

A conventional frequency-modulated (FM) signal using the same modulating signal has the familiar expansion

$$\cos\,[\omega_0 t + f(t)] = \sum_{k=-\infty}^{\infty} J_k\left(\frac{\Omega}{\omega}\right)\cos\,(\omega_0 + k\omega)t. \quad (9)$$

The line spectra of (8) and (9) are plotted in Fig. 2 for several deviation ratios. The component magnitudes have been adjusted so that the spectra represent signals of equal average power. Generally speaking, the cancellation of the lower sideband is accompanied by an extension of the upper sideband. Nevertheless, the "widths" of the one-sided spectra appear reduced by roughly one-third.

The one-sided signal can be generated from a conventional phase-modulated signal by forming the Hilbert transform (by means, say, of a $\pi/2$ phase shifter) and amplitude modulating with the exponential; limiting in the receiver will then restore the lower sideband and permit the use of a conventional discriminator. However, the new threshold behavior is not clear.

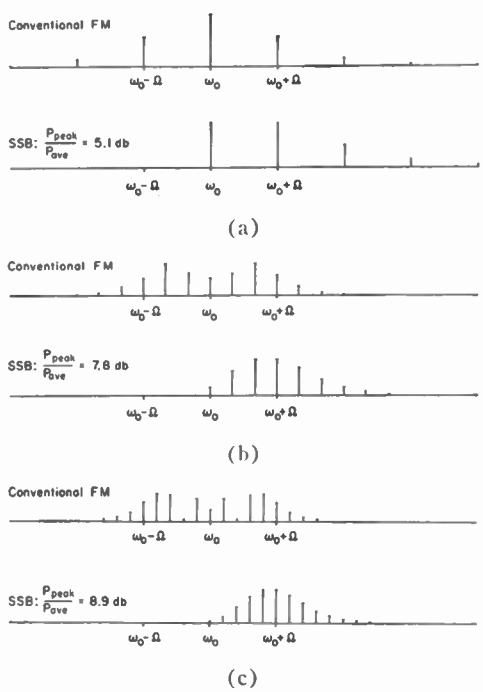


Fig. 2—Magnitudes of spectral components for conventional and SSB forms of FM with sinusoidal modulation. (a) Modulation index $\Omega/\omega = 1$. (b) Modulation index $\Omega/\omega = 3$. (c) Modulation index $\Omega/\omega = 5$.

[6] It would be difficult to find a more striking example of the mathematical simplicity afforded by the use of the analytic signal than that demonstrated by the development leading to the Fourier series expansion giving by (8). The conventional approach is to multiply the Fourier series expansion

$$e^{-j\hat{f}(t)} = \sum_{n=-\infty}^{\infty} I_n\left(\frac{\Omega}{\omega}\right)\cos n\omega t$$

by that for $\cos\,[\omega_0 t + f(t)]$ given by (9) and then reduce the double summation; needless to say, the procedure is tedious.

Exponential (Envelope) Modulation

Double Sideband:

$$\begin{aligned} m(t) &= e^{g(t)}, \quad g(t) = \alpha \log f(t), \quad f(t) > 0 \\ s(t) &= f^{\alpha}(t)\cos\omega_0 t + if^{\alpha}(t)\sin\omega_0 t \\ |\,s(t)\,| &\cong f^{\alpha}(t), \qquad \omega_0 \gg \Omega_{\max}. \end{aligned} \quad (10)$$

The intermediate function $g(t)$ is introduced to clarify the incorporation of the analytic signal in (11). The difference between (4) and (10) is obviously trivial and the latter has been included mainly for completeness; they are, of course, identical for $\alpha = 1$.

Single Sideband:

$$\begin{aligned} m(t) &= e^{[g(t)+i\hat{g}(t)]}, \quad g(t) = \alpha \log f(t), \quad f(t) > 0 \\ s(t) &= f^{\alpha}(t)\cos\left[\omega_0 t + \alpha\,\widehat{\log f(t)}\right] \\ &\quad + if^{\alpha}(t)\sin\left[\omega_0 t + \alpha\,\widehat{\log f(t)}\right] \\ |\,s(t)\,| &= f^{\alpha}(t), \qquad \omega_0 > 0. \end{aligned} \quad (11)$$

The SSB version given by (11) is related to (7), from which it can be derived by considering the logarithm of the modulating signal. The reason these two classes of exponential modulation have been made distinct here is that (6) and (7) contain the modulating signal in the phase of the modulated signal, while (10) and (11) contain the modulating signal in the envelope; hence, the designations.

The exponent α is important because of the relationship which the choice of its value as either 1 or $\frac{1}{2}$ bears on the question of compatible SSB modulation. Powers [2] considers the case $\alpha = \frac{1}{2}$, so that the modulating signal is contained in the square of the envelope, thus requiring a square-law envelope detector for distortionless reproduction. Since it is this case which produces a modulated signal having the same spectral width as a conventional SSB signal, he concludes that true compatibility (*i.e.*, a system usable with a linear envelope detector) is impossible.

Lyannoy [4] reaches the same conclusion and, incidentally, labels the case Optimal Amplitude and Phase Modulation (OAPM) for Square-Law Detection. He also considers $\alpha = 1$ or OAPM for Linear Detection and points out that this modulated signal occupies a bandwidth just twice the maximum modulating frequency and that it is compatible with a linear envelope detector.

Naturally, there would be no advantage to a compatible SSB system which has the same spectral width as a conventional AM system, but Lyannoy adds that the bandwidth can be halved in practice by suitable filtering because of the characteristics of speech. This derives from the fact that "a real signal has components

whose modulation depth approaches 100 per cent only in the range up to 1500 or 2000 cps, the modulation depth decreasing rapidly thereafter." Hence, removal of the upper half of the signal spectrum permits the large-modulation-index, low-frequency components to be received without distortion, while the small-modulation-index, high-frequency components are received with a small but tolerable distortion.

An experimental study of OAPM for Linear Detection is reported by Tetel'baum and Grinevitch [18]. Their method of deriving the proper phase function in (11) is to approximate the logarithm by a nonlinear vacuum-tube characteristic and to obtain the Hilbert transform from its defining integral by means of a delay-line summation.

A method of composite amplitude and phase modulation was demonstrated by Villard [19] as early as 1948. His technique is to form 90° phase-shifted versions of the modulating signal, then amplitude modulate with one and phase modulate with the other. Such a procedure amounts, in fact, to using the linear approximation to (11) for $\alpha = 1$.[7] As a result, his sideband elimination is imperfect, and third- and higher-order sidebands appear. Villard also observes that reception of the second-order sidebands is necessary only for the lower modulating frequencies, because of the characteristics of speech referred to above.

The Kahn system of Compatible SSB (CSSB) [3] is apparently closely related to that given by (11) for $\alpha = 1$. However, the Kahn system is unique in that it derives the phase modulation by limiting a SSB (plus carrier) version of the modulating signal. It can readily be deduced from (5) that this is equivalent to phase modulation with the function

$$\phi(t) = -\tan^{-1}\left(\frac{\hat{f}(t)}{1 + f(t)}\right),$$

which has the same leading term as $\widehat{\log f(t)}$ for $f(t)$ small. Kliagin [20] uses this method with sinusoidal modulation to compare the Kahn system with one of "quadrature" modulation which closely resembles Villard's.

In its latest configuration, the Kahn system includes an additive pair of unspecified nonlinear operations on the SSB signal before limiting. Thus, it is not possible, at the moment, to express the phase modulation analytically in order to make further comparisons with the optimal modulation of (11).

Conclusion

It has been shown that analog-modulated signals are representable in general as the product of a carrier function and a modulation function. When an analytic signal is chosen for the carrier, its effect spectrally becomes one of translating the signal spectrum to the vicinity of the carrier frequency. The process is one of pure translation when the carrier is harmonic, $\exp(i\omega_0 t)$, or one of translation plus spreading when the carrier has a finite spectral extent (*e.g.*, when noiselike).

The useful forms of the modulation function are the linear and exponential functionals from which three classes of modulated signals can be derived, *viz.*, amplitude, phase, and envelope modulation. Within these classes, the use of the purely real or analytic signal forms of the modulating signal leads naturally to the DSB and SSB (*i.e.*, two-sided and one-sided) spectral forms, respectively, of the modulated signal.

The existing familiar forms of modulation are readily identified within the above structure. In addition, a new form of modulation which may be called SSB FM is revealed. It can be derived from a conventional phase-modulated signal by an additional amplitude modulation using the exponential function of the modulating signal's Hilbert transform. The resulting modulated signal will have a one-sided spectrum about the carrier frequency and be compatible with existing FM receivers. The advantage is a decrease in signal bandwidth; the disadvantage, the loss of a constant transmitter output level.

From a practical viewpoint, this form of FM may find little acceptance in wide-band systems, not only because of the unfavorable peak-to-average power ratio required of the transmitter, but also because of the lack of any particular pressure to conserve the radio spectrum at present in services such as broadcast FM or specialized military and civilian communication systems.

On the other hand, some of the bands assigned to public use are quite crowded and users are therefore disposed toward modulation techniques which make more efficient use of the available spectrum. A compatible narrow-band SSB FM system of the type described here may prove useful in such applications by narrowing the signal bandwidth for a given modulation index without sacrificing the "improvement" or the immunity to impulse noise associated with FM. The peak-to-average power ratio of a narrow-band SSB FM signal may prove an acceptable burden in that case.

Observation of the comparable (*i.e.*, SSB) form of envelope modulation commends Powers' suggestion of the application of the noncompatible, square-law type of envelope modulation [$n = \frac{1}{2}$ in (9)] as a means of SSB data transmission [2]. Conventional SSB is adequate for speech or music because of the well-known tolerance of the human ear for phase or even frequency errors in reproduction. However, data waveforms must be reproduced without such errors, so exact carrier reinsertion is required (a procedure which is frequently difficult). The fact that the modulating signal is contained in the envelope of (9) indicates that square-law envelope detection should result in distortionless reception, thus obviating carrier reinsertion without sacrificing spectral economy.

[7] Villard's technique is actually a hybrid since it also corresponds to the linear approximation to (7) and can, therefore, be demodulated by a phase detector as well.

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Observations of Satellite-Related Ionization Effects Between 1958 and 1960*

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***Summary*—Three independent sets of observations over a two-year period (1958–1960) involving the satellites 57021 and 58042 are described which show a marked correlation between satellite passes and ionospheric-ionization effects. The effects tend to peak at the time of near approach of the satellite or somewhat before with most of the effects occurring in the interval of 20 minutes before to 10 minutes after the satellite's passage. Other observations of some short-range effects are described which indicate a dependence of the effects on the satellite's heading as a function of the local time and on its position with respect to the observing location.**

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Introduction

WHILE monitoring the 20-Mc transmissions of WWV at the Ohio State University Radio Observatory (OSURO) late in 1957 it was noted that large increases in the signal strength of WWV often occurred at the time of passage of the artificial earth satellites 57012 and 57021 (Sputniks I and II).[1] WWV is situated near Washington, D. C., at a distance of about 330 miles from the observing point at Columbus, Ohio. Much of the time, especially at night, the electron density in the ionosphere is too small to reflect signals from Washington to Columbus on 20 Mc. Thus,

[1] J. D. Kraus, "Detection of Sputniks I and II by CW reflection," PROC. IRE, vol. 46, pp. 611–612; March, 1958.

the WWV signals on 20 Mc are ordinarily very weak or entirely absent. When large bursts of signal from WWV lasting of the order of a minute were noted under these conditions at the time of passage of an artificial earth satellite, it suggested that temporary increases in the electron density in the ionosphere may have been induced by the satellite.

Sputnik II Observations

Continued studies of this satellite ionization phenomenon gave further evidence of the effect. For example, the bursts or enhancements of the 20-Mc WWV signal observed between about 0030 and 0620 local time (EST) on 26 consecutive days, February 3–28 (inclusive), 1958, are presented as a function of the time of the near approach of 57021 (Sputnik II) in Fig. 1(a). A total of 169 enhancements were observed during 78 passes of the satellite. At the time of the observations only two objects, 57021 and 58011 (Explorer I), were in orbit.

The histogram in Fig. 1(a) gives the number of enhancements observed per 10-minute interval during the 78 passes of 57021 for intervals up to 50 minutes before and after the near approach of the satellite to Columbus. The total interval (100 minutes) covers approximately one orbital period. The histogram shows marked peaks in the interval from 20 minutes before to 10 minutes after the near approach of the satellite. A chi-squared evaluation of the histogram indicates that the probability of such a distribution resulting from a random phenomenon (expected distribution uniform) is only one per cent. For purposes of comparison, the enhancements during the 0030 to 0620 period for all 26 days were added with respect to an arbitrary universal time (U.T.) with 100-minute separation as reference, corresponding to 0210, 0350, and 0530 local time. The enhancements per 10-minute interval obtained in this way are presented in Fig. 1(b). A chi-squared evaluation of this histogram indicates that the probability of such a distribution resulting from a random phenomenon is 84 per cent. Thus, when the data are added as in Fig. 1(b) the histogram is very likely representative of a random distribution, but with the pass times of 57021 as reference a distribution [Fig. 1(a)] is obtained that is nonrandom and which has a peak near the times of passage of 57021. Accordingly, one may conclude that there is a correlation between the WWV signal enhancements and the passes of 57021.

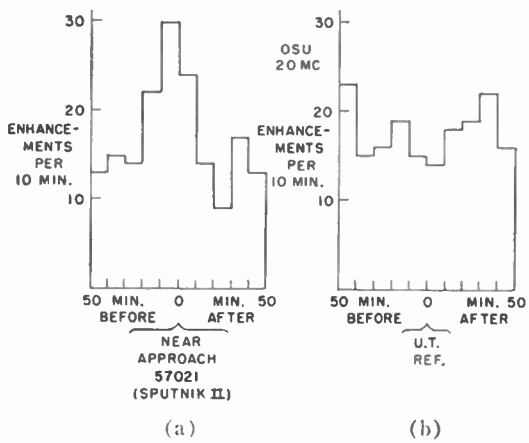
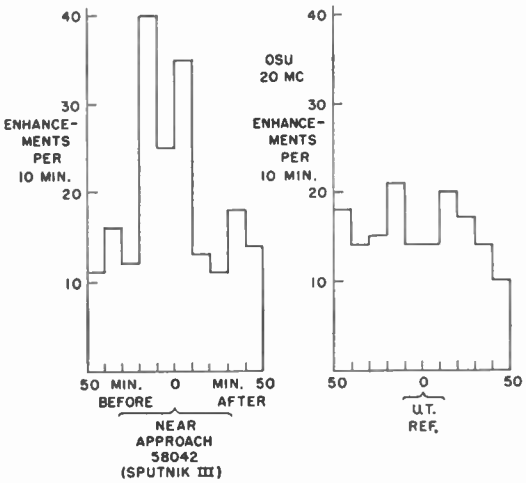


Fig. 1—(a) WWV signal enhancements observed at the Ohio State University Radio Observatory (OSURO) between 0030 and 0620 local time on 26 consecutive days in February, 1958, with respect to the time of near approach of 57021 (Sputnik II). (b) Enhancements during the same intervals, but with respect to arbitrary universal times (UT), as reference.

Sputnik III Observations

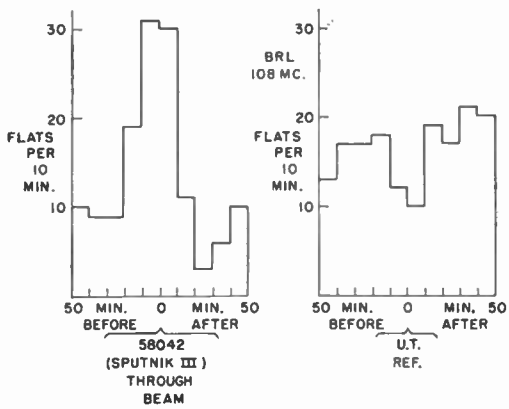
In a later experiment WWV signal enhancements were studied with respect to the time of near approach of 58042 (Sputnik III) under carefully controlled conditions.[2] These included the recording of the direct signals from the 58042 transmitter in addition to the WWV 20-Mc enhancements with separate receiving systems. Also only enhancements of magnitude greater than an arbitrary standard value and longer than 30 seconds in duration were counted. These requirements probably eliminated most of the enhancements which may have been caused by meteors. From the 58042 transmissions, using an interferometer antenna, an accurate determination of the time of near approach made by the satellite was possible. For reception of the 20-Mc WWV signals a 3-element directional antenna directed to the north was used. The orientation of this antenna was such that reception of WWV over a direct ionospheric path from Washington to Columbus was discriminated against. With this instrumentation observations were made between 0000 and 0600 local time on 14 nights in January and February, 1959. There were 50 passes of 58042 during this interval and 195 WWV signal enhancements were observed. The number of enhancements per 10-minute interval is shown in Fig. 2(a) as a function of time with respect to the near approach of the satellite. Large peaks in the histogram are evident in the interval of 20 minutes before to 10 minutes after the time of near approach of 58042. A chi-squared evaluation of the histogram yields a probability of less than 0.0001 per cent that it was due to a random phenomenon. By way of comparison, the enhancements observed during the 0000 to 0600 period for all 14 nights were added with respect to arbitrary universal times as reference corresponding to 0100, 0300 and 0500 local time. The enhancements per 10-minute interval obtained in this way are presented in Fig. 2(b). A chi-squared evaluation of this histogram gives a probability of 72 per cent that it was due to a random phenomenon. Thus, with an arbitrary time as reference the enhance-

[2] J. D. Kraus, R. C. Higgy, D. J. Scheer, and W. R. Crone, "Observations of ionization induced by artificial earth satellites," *Nature*, vol. 185, pp. 520–521; February 20, 1960.



(a) (b)

Fig. 2—(a) WWV signal enhancements observed at the OSURO between 0000 and 0600 local time on 14 nights in January and February, 1959, with respect to the time of near approach of 58042 (Sputnik III). (b) Enhancements during the same intervals, but with respect to arbitrary universal times as reference.



(a) (b)

Fig. 3—(a) Flats observed with the doploc system (BRL) on 25 days during 1959 and 1960 with respect to the time 58042 (Sputnik III) passed through the doploc beam. (b) Flats during the same intervals, but with respect to arbitrary universal times as reference.

TABLE I

SUMMARY OF DATA FOR FIGS. 1, 2, AND 3

Object	Observer	Date	Frequency	Number of Passes	Number of Events	Figure	$(Chi)^2$	Probability*	Figure	$(Chi)^2$	Probability*
57021 (Sputnik II)	OSU	February 3–28, 1958	20 Mc	78	169	1(a)	21	0.01	1(b)	4.7	0.84
58042 (Sputnik III)	OSU	January–February, 1959	20 Mc	50	195	2(a)	52	<0.000001	2(b)	6.2	0.72
58042 (Sputnik III)	BRL	July, 1959–April, 1960	108 Mc	27	138	3(a)	60	<0.0000001	3(b)	7.4	0.60

* $(Chi)^2$ and corresponding probabilities are for a random distribution (expected distribution uniform, *i.e.*, same number of events per each 10-minute interval).

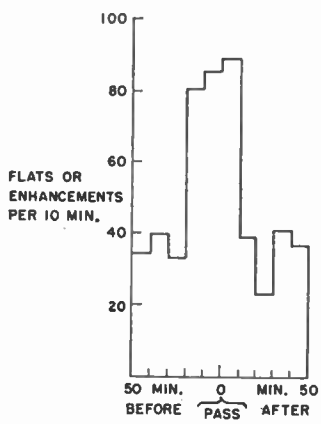


Fig. 4—Total events (WWV enhancements or doploc flats) of Figs. 1(a), 2(a), and 3(a) with respect to the time of the satellite passes. Over 500 events are involved in the histogram.

ments have a typically random distribution [Fig. 2(b)], but with the pass time of 58042 as reference a distribution [Fig. 2(a)] is obtained that almost certainly represents a nonrandom phenomenon with a pronounced correlation between the enhancements and the passes of 58042. Other studies at the OSURO by means of WWV signals[3,4] and by radar techniques[2] have also given evidence of satellite-related ionization effects.

Doploc Observations of Sputnik III

In a recent Ballistic Research Laboratories (BRL) report, Lootens[5] describes observations made with the doploc system which are of interest in connection with studies of satellite-induced ionization. The doploc observations were made on 108 Mc using a 50-kw CW transmitter at Fort Sill, Okla., and a narrow-band phase-locked receiver at Forrest City, Ark. Both output level and frequency were recorded for any signals detected above the system threshold. Sensitivity was sufficient to detect the largest satellites on their lowest passes by direct reflections from the satellite body, the received frequency following the normal-type Doppler curve. In addition, numerous short bursts of signal were observed which showed no perceptible Doppler shift. These constant frequency reflections, or "flats," frequently showed a tendency to cluster around the times of passage of artificial earth satellites. The flats had signal strengths which were generally equal to or greater than the direct Doppler-changing reflection. All flats were recorded using essentially vertically-directed high-gain fan-beam antennas at both transmitter and receiver, so oriented as to cover substantially the same space volume.

There are 5 figures in Lootens' report, each covering a 5-day interval, during which there were passes of 58042 (Sputnik III) through the antenna pattern overlap zone.[6] The time at which each flat was observed during these 5 periods of 5 days each (total 25 days) is indicated in these figures. During the 25 days covered, there were 27 passes of 58042 through the antenna beam, and 138 flats were observed. If all the flats observed during a continuous 100-minute interval centered on each pass time of 58042 are added (as a function of time before and after the pass), the result shown by the histogram of Fig. 3(a) is obtained. Here the ordinate gives the number of flats observed per 10-minute interval and the abscissa gives the time before and after the time of pass of 58042. The histogram indicates a strong tendency for the flats to cluster around the pass time of 58042 in the interval from about 20 minutes before to about 10 minutes after the pass. A chi-squared evaluation of this histogram gives a probability of less than 0.00001 per cent that it could be due to a random phenomenon. For purposes of comparison, the flats observed during the same periods were added with respect to arbitrary universal times, such as 0000, 0200 UT, etc., as reference. The number of flats per 10-minute interval obtained in this way are presented in Fig. 3(b). A chi-squared evaluation of this histogram yields a probability of 60 per cent that it was due to a random phenomenon. Thus, as in the previous two sets of data, one may conclude that there is a correlation between the events (in this case, flats) and the satellite passes. The statistical data for all three sets of observations (Figs. 1, 2, and 3) are summarized in Table I.

It is interesting to note that although two different satellites (57021 and 58042), two different frequencies (20 and 108 Mc) and two different groups (OSU and BRL) were involved, the histograms of Figs. 1(a), 2(a), and 3(a) have a marked similarity with a tendency for peaks in the interval from 20 minutes before to 10 minutes after the time of passage of the satellite. Actually there is reason to expect a similarity in the 3 histograms since both satellites were large and in very similar orbits (65° inclination). If these 3 histograms are added the histogram shown in Fig. 4 results. Here the minus 20 to plus 10 minute peak is especially prominent. Since over 500 flats or enhancements are involved, this histogram should have statistical significance.

Short-Range Effects

Of further interest in Lootens' report[5] are 12 figures (Figs. 3 to 14, inclusive) each of which presents one or two minutes of the original record showing Doppler frequency and signal strength at the time of the passage through the antenna beam of satellite 58042. The records are for 12 passes of 58042 (7 NW to SE and 5 SW to NE) between December, 1959, and April 4, 1960. On each of these passes a direct reflection from the satellite body was obtained when the satellite was in the antenna beam overlap zone. This reflection shows the expected Doppler shift in frequency. However, constant frequency reflections or flats were also recorded some seconds before or after the satellite passed through the antenna beam overlap zone. If flats occurred at random one would expect only 3 within ± 30 seconds of the pass time. However, 18 flats or 6 times as many were observed.

If the position of 58042 is indicated on a map for the passes on which flats were observed, the result is as shown in Fig. 5. A nearly E-W line passing through Fort Sill, Okla., and Forrest City, Ark., indicates the location of the overlap zone of the doploc fan-beam antenna patterns. The zone was wide in the E-W direc-

[3] J. D. Kraus, R. C. Higgy, and W. R. Crone, "The satellite ionization phenomenon," Proc. IRE, vol. 48, pp. 672–678; April, 1960.

[4] J. D. Kraus, "Evidence of satellite-induced ionization effects between hemispheres," Proc. IRE (*Correspondence*), vol. 48, pp. 1913–1914; November, 1960.

[5] H. T. Lootens, "Satellite-induced ionization observed with the doploc system," Ballistic Research Laboratory, Aberdeen Proving Ground, Md., Memo. Rept. No. 1362; 1961.

[6] The 5-day periods were between July 22–26, September 13–17, and December 15–21, 1959, and January 12–16, and March 31–April 6, inclusive, 1960.

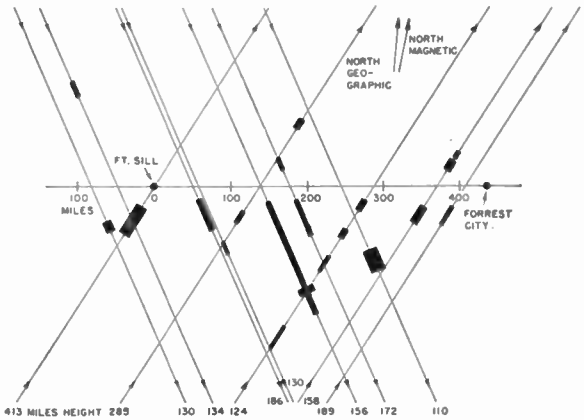


Fig. 5—Locations of 58042 at times when close flats were observed with doploc system. Twelve passes (7 NW-SE and 5 SW-NE) are involved with most of the flats occurring when 58042 was south of the antenna overlap zone.

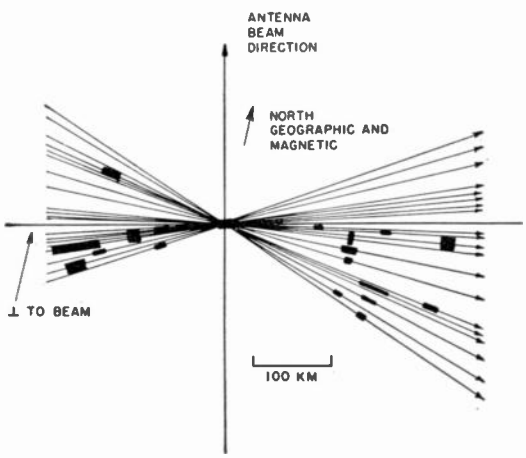


Fig. 6—Locations of 58042 with respect to antenna beam at times when radar reflections were observed in Finland.

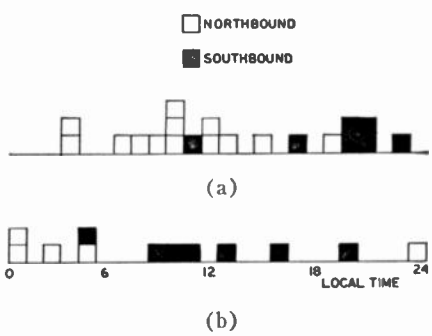


Fig. 7—Distribution of north- and southbound passes of 58042 giving radar echoes or flats as a function of time of day as observed in Finland (a) and with doploc system (b).

tion and narrow in the N-S direction. Distances are given by miles along this line east and west of Fort Sill. The subsatellite tracks are shown for the 7 NW to SE and 5 SW to NE passes of 58042, with the location of the satellite when flats occurred shown by the bars on the tracks. The relative intensity of the flats is suggested by the widths of the bars, the wider bars indicating stronger flats.

Of interest is the fact that 14 or 78 per cent of the 18 flats occurred when 58042 was south of the antenna beam overlap zone. If the location of 58042 with respect to the antenna-beam overlap zone was not important, one should expect an equal number of flats to each side of the beam, *i.e.*, 9 to the north and 9 to the south instead of 4 to the north and 14 to the south as observed. The predominance of flats south of the antenna beam appears independent of the satellite heading, 7 passes being southbound and 5 northbound.

In observations of 58042 at the State Institute of Technical Research in Finland,[7] a concentration of 100-Mc radar echoes at locations south of the antenna-beam crossing point was noted. These results are summarized in Fig. 6 and it will be noted that they are similar to those found with the doploc system. Twenty passes of 58042 are involved with the satellite position during the radar echo shown by the bars on the subsatellite tracks. In both these observations, and those reported by Lootens, the satellite tended to be south of the observing region when the echoes were observed.

The magnetic-dip angle of the earth's field at both observing locations is about 70°. If the ionization detected was along a magnetic field line through the satellite and below it (*i.e.*, at a magnetic subsatellite point) it would be to the north of the satellite at both observing locations, and hence, in the antenna beams when the satellite was to the south as observed. Accordingly, these observations suggest that the satellite may produce a disturbance which propagates down along magnetic field lines to a point in the beams. If, for example, the disturbance were a magneto-hydrodynamic (mhd) wave a reflection would be expected at the base of the ionosphere where there is a transition in conductivity of the medium from a large value above to a small value below. This transition would be analogous to an open circuit on the magnetic field "transmission line" with an attendant increase in effects at the reflection point. However, some entirely different mechanism, such as one involving a horizontal ionized column or plume may be responsible for the effect.

A further analysis of the observations made in Finland shows that the echoes obtained on northbound passes of 58042 tended to be about 12 hours earlier than the echoes detected on southbound passes. The distribution obtained is shown in Fig. 7(a) as a function of local time. Making a similar distribution as a function of local time for the 12 passes of 58042 observed with the doploc system, as discussed above, the result is as in Fig. 7(b). Here again one notes that the times at which flats were obtained on northbound passes tend to occur about 12 hours before the times at which flats were observed on southbound passes. This trend is suggestive

[7] Martti Tiuri, "Investigations of radio reflections from satellite-produced ion trails using 100-Mc CW radar," State Institute for Technical Research, Helsinki, Finland, Publ. 59; 1960.

of an influence by magnetic storm currents in the ionosphere. Such currents follow a diurnal cycle, reversing direction at a given location every 12 hours. There is a time-phase difference of about 7 hours between the two distributions. This is as might be expected in view of the different satellite headings and different latitudes at the sites used for the two sets of observations.

Conclusion

The data of Figs. 5, 6, and 7 suggest satellite-induced ionization effects at relatively short ranges from the satellite or within a few hundred kilometers. However, the data of Figs. 1, 2, 3, and 4 indicate satellite-related ionization effects from greater distances. As indicated in these figures, many satellite-related events appear to occur as much as 20 minutes before the time of passage of the satellite. Assuming a satellite velocity of 350 kilometers per minute, the satellite would be, therefore, at a distance of 7000 kilometers from the observer at the time of such events. Although mhd waves were mentioned as a mechanism in connection with the short range effects, this is only a speculative suggestion. More data are required to determine what mechanism or mechanisms may actually be responsible for both the short and longer distance effects and systematic high-power pulse-radar observations are being conducted at the OSURO for this purpose.

The Dielectric Microwave Resonator*

A. OKAYA† and L. F. BARASH‡, member, ire

Summary—**Pieces of single crystals of rutile show high Q resonances in the microwave range. A piece about $(1/7\,\lambda)^3$ has a Q as high as that of a metal-walled cavity at room temperature. Lowering the temperature increases both the Q and the resonant wavelength. Q's of 10^5 were seen at 4°K. A frequency equation of first- and second-order approximations was derived in terms of the dimensions and the anisotropic dielectric constant, ϵ', for rectangular parallelepiped resonators. Accurate values of anisotropic ϵ' were obtained.**

In the anisotropic medium there are two types of resonant modes, one of which has an outside E field similar to an electric multipole, and the other an outside magnetic field similar to a magnetic multipole. These two types of modes degenerate into one type if the dielectric is isotropic. Resonators were also made of ceramic rutile and strontium titanate, both of which had Q values of thousands. An extremely high unloaded Q, of the order of a million, was seen at 1.4°K on a KRS-5 (ThBr-ThI) single crystal at X band.

I. Introduction

METAL CAVITIES have long been used as microwave resonators, occasionally combined with dielectric materials. Dielectrics themselves have not been used in this way, but were considered only theoretically for use at high frequencies [2].

One can see that the surface between air and a dielectric will be a perfect reflector of microwaves if the angle of incidence is greater than the critical angle, $\theta_c = \sin^{-1} 1/\epsilon'^{1/2}$ where ϵ' is the imaginary part of electric susceptibility and is the dielectric constant. Hence, a piece of dielectric in air may be a microwave resonator if the microwave loss is small.

It was found that a piece of rutile, *et al.*, fulfilled the conditions for such a resonator and that the unloaded Q of a rutile resonator was as high as, or higher than, that of a metal-walled cavity [1]. The high Q results from very small dielectric losses and no wall losses in the resonator. The resonant frequency is determined by the ϵ', the dimensions, and the shape of a resonator.

This resonance effect in rutile found an effective use in electron paramagnetic resonance experiments [3] on Fe^{3+} and Cr^{3+} in rutile single crystals. The sensitivity of EPR was enormously increased by this resonance phenomenon, and resonances of Fe^{3+} were observed from 2 kMc to 120 kMc with ease. Furthermore, not only Ti^{3+}, but the spectra of sparsely populated color centers and vacancies [4] in a "pure" rutile single crystal (impurity less 0.001 per cent) were also seen.

The large zero-magnetic-field splitting of Fe^{3+} (43.3 kMc and 81.3 kMc) and Cr^{3+} (43 kMc), the fairly long relaxation times, and the microwave resonant characteristics make rutile quite useful as a maser material, especially in the millimeter wave region. (Several rutile millimeter and centimeter masers have been made recently by groups at CRL [5], M.I.T. [6], and RCA [7].)

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For these reasons and for general interest, the study of microwave field configurations and the characteristics of dielectric resonators was undertaken.

A first-order approximation for the frequency equation of modes in anisotropic media for rectangular parallelepipeds was derived, assuming perfect reflection at the resonator surface. This was compared with experimental results. Modes were identified, and fairly good agreement of frequencies at high modes was obtained. Extension of the first-order approximation to the more general case of anisotropic permeability μ and the dielectric constant ϵ' is treated in Appendix I. For the second-order approximation, more realistic boundary conditions were considered. The results were applied to the fundamental modes to obtain accurate solutions for frequency in terms of the dimensions and dielectric constant. Using measured resonant frequency, the anisotropic ϵ' was calculated accurately by this equation. Finally, the use of dielectric resonators in microwave filter circuits is discussed.

II. First-Order Approximation of the Resonant Frequency Equation

Assuming that the boundary coincides with a node or antinode of either the electric or magnetic field, we confine our discussion to dielectric slabs of rectangular cross section and solve Maxwell's equation for the dielectrics. It will be shown in Section III that this assumption is valid if the resonator dimension is larger than the effective wavelength in the crystal, $\lambda/\epsilon'^{1/2}$. We also assume that $\mu=1$, both inside and outside the dielectric, and that the dielectric constants along the three axes, taken parallel to the resonator surfaces, are ϵ_x', ϵ_y', and ϵ_z', respectively, (see Fig. 1). As the solutions of Maxwell equations

$$\nabla \times \boldsymbol{E} = -\dot{\boldsymbol{H}},$$
$$\nabla \times \boldsymbol{H} = \epsilon \dot{\boldsymbol{E}}, \tag{1}$$

we assume an electric field of the following form

$$E_x = A_x \begin{matrix}\cos\\ \sin\end{matrix} \xi x \begin{matrix}\cos\\ \sin\end{matrix} \eta y \begin{matrix}\sin\\ \cos\end{matrix} \zeta z e^{i\omega t},$$

$$E_y = A_y \begin{matrix}\sin\\ \cos\end{matrix} \xi x \begin{matrix}\sin\\ \cos\end{matrix} \eta y \begin{matrix}\sin\\ \cos\end{matrix} \zeta z e^{i\omega t},$$

$$E_z = A_z \begin{matrix}\sin\\ \cos\end{matrix} \xi x \begin{matrix}\cos\\ \sin\end{matrix} \eta y \begin{matrix}\cos\\ \sin\end{matrix} \zeta z e^{i\omega t}. \tag{2}$$

Hereafter, for convenience, ϵ' will be written as ϵ, where the boundary conditions on the surface are already taken into account.

Then we take the following three cases which are applicable to rutile:

$$\text{Case I} \quad \epsilon_z = \epsilon_\parallel \qquad \epsilon_x = \epsilon_y = \epsilon_\perp,$$
$$\text{Case II} \quad \epsilon_x = \epsilon_\parallel \qquad \epsilon_y = \epsilon_z = \epsilon_\perp,$$
$$\text{Case III} \quad \epsilon_y = \epsilon_\parallel \qquad \epsilon_x = \epsilon_z = \epsilon_\perp, \tag{3}$$

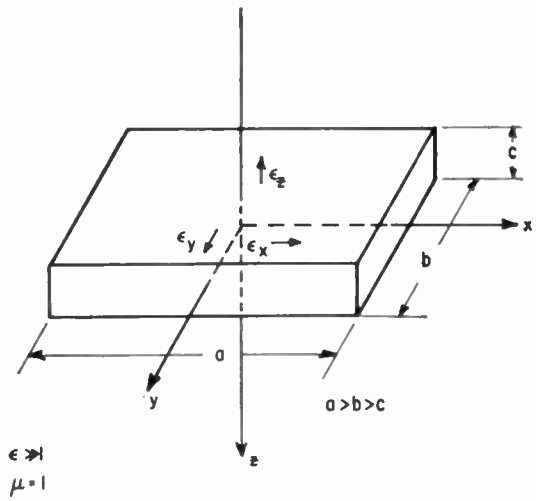


Fig. 1—Rutile dielectric microwave resonator.

where $\epsilon_\parallel$ and $\epsilon_\perp$ are the dielectric constants parallel and perpendicular to the optic axis. Dimensions a, b, and c are the crystal dimensions along the x, y, and z axes, as shown in Fig. 1. From the linear equations, which will be reduced from (1) and (2) and after considering the conditions of (3), we have solutions which satisfy the condition of nonzero E_x, E_y, and E_z field.

$$\text{Case I} \quad \left(\frac{2}{\lambda}\right)^2 = \frac{1}{\epsilon_\perp}\left[\left(\frac{l}{a}\right)^2 + \left(\frac{m}{b}\right)^2 + \left(\frac{n}{c}\right)^2\right],$$
$$\left(\frac{2}{\lambda}\right)^2 = \frac{1}{\epsilon_\parallel}\left[\left(\frac{l}{a}\right)^2 + \left(\frac{m}{b}\right)^2\right] + \frac{1}{\epsilon_\perp}\left(\frac{n}{c}\right)^2,$$
$$\text{Case II} \quad \left(\frac{2}{\lambda}\right)^2 = \frac{1}{\epsilon_\perp}\left[\left(\frac{l}{a}\right)^2 + \left(\frac{m}{b}\right)^2 + \left(\frac{n}{c}\right)^2\right],$$
$$\left(\frac{2}{\lambda}\right)^2 = \frac{1}{\epsilon_\perp}\left(\frac{l}{a}\right)^2 + \frac{1}{\epsilon_\parallel}\left[\left(\frac{m}{b}\right)^2 + \left(\frac{n}{c}\right)^2\right],$$
$$\text{Case III} \quad \left(\frac{2}{\lambda}\right)^2 = \frac{1}{\epsilon_\perp}\left[\left(\frac{l}{a}\right)^2 + \left(\frac{m}{b}\right)^2 + \left(\frac{n}{c}\right)^2\right],$$
$$\left(\frac{2}{\lambda}\right)^2 = \frac{1}{\epsilon_\parallel}\left[\left(\frac{l}{a}\right)^2 + \left(\frac{n}{c}\right)^2\right] + \frac{1}{\epsilon_\perp}\left(\frac{m}{b}\right)^2. \tag{4}$$

where

$$\xi = \frac{l}{a}\pi, \quad \eta = \frac{m}{b}\pi, \quad \zeta = \frac{n}{c}\pi$$

and l, m and n are integers.

There are two solutions for each case. The first is the same for each of the three cases and results from the isotropy of μ. The second solution, which is different for the three cases, results from the anisotropy of ϵ. Exactly the same results are obtained when the equation for the magnetic field is solved. Detailed boundary conditions are discussed in Section III. In Cases I and II, (4) is about 25 per cent in error for the lowest modes but becomes increasingly accurate for higher modes where the assumption of nodes or antinodes at the boundaries becomes more realistic. The above solutions, however, are accurate enough for identification of the modes and

for rough estimates of frequency. The solutions for the general case of completely anisotropic μ and ϵ are derived in Appendix I, and one can deduce (4) from those results as a special case. For a completely isotropic medium, there is a single-frequency equation and, therefore, only half as many modes.

III. Boundary Conditions

The boundaries of the dielectric resonator could be interpreted as follows. Suppose we have an isotropic dielectric sheet of infinite extent whose thickness in the z direction is c, as shown in Fig. 2. We shall, for simplicity, discuss this two-dimensional case first. The E field parallel to the resonator surface should be continuous across its boundary. Hence

$$E_{iz} = E_{oz}, \tag{5}$$

where E_{iz} and E_{oz} are the inside and outside electric fields parallel to the surface, respectively. Also,

$$\nabla \times E = \dot{B},$$
$$\nabla \cdot B = 0. \tag{6}$$

Hence, $\partial E_x/\partial z =$ constant.

Let us assume that, at resonant condition, the standing wave of an electric field has sinusoidal variation inside a resonator and decays exponentially outside it.

The detailed discussions of Section IV will explain that the exponential decay is a good approximation in the space close to the surface. The x component of the electric field will be given by

$$\text{Inside:} \quad E_{xi} = e_{xi} \cos \zeta_i z e^{i\omega t},$$
$$\text{Outside:} \quad E_{xo} = e_{xo} e^{-\zeta_o z} e^{i\omega t}, \tag{7}$$

where ω is the resonant angular frequency and ζ_i and ζ_o are the propagation constants along the z direction inside and outside, respectively.

From (5), (6), and (7) we have

$$\zeta_i \tan \zeta_i \frac{c}{2} = + \zeta_o, \tag{8}$$

here, c is the thickness of the resonator. Then e_{xi} is a function of the propagation constant ρ along the x direction, *i.e.*,

$$e_{xi} = e_i e^{\rho_i x_i}. \tag{9}$$

The field outside, but immediately adjacent to the surface, has to equal the field just inside and has to have the same maximum and minimum points. Therefore, the propagation constants inside and outside a resonator should be the same and $\rho_i = \rho_o$.

On the other hand, the propagation constants must satisfy the following equation [8]:

$$\rho_i^2 + \zeta_o^2 = \beta^2 \text{ outside}, \tag{10a}$$
$$\rho_i^2 + \zeta_i^2 = \epsilon\beta^2 \text{ inside}, \tag{10b}$$

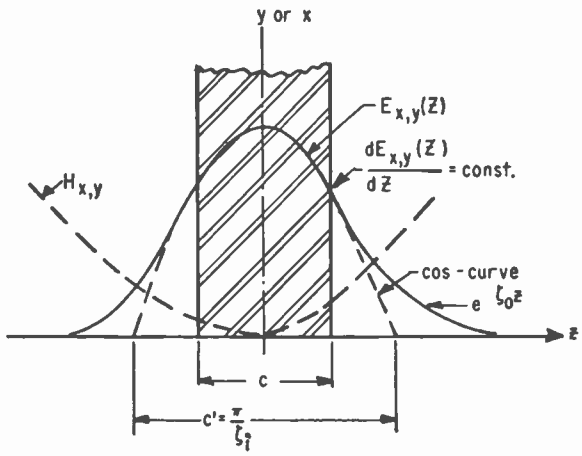


Fig. 2—Field configuration along the z axis.

where

$$\beta = \frac{2\pi}{\lambda} = \frac{\omega}{c} = \text{intrinsic phase constant.}$$

Therefore, $\zeta_i \gg \zeta_o$ for large ϵ.

Putting (10) into (8) we have

$$\left(\frac{c\zeta_i}{2}\right)^2 \tan^2 \frac{c\zeta_i}{2} = (\beta^2 - \rho_i^2)\left(\frac{c}{2}\right)^2, \tag{11}$$

where ρ_i is imaginary [as we expected from (9)]. The transverse propagation constant ρ_i is of the order of $2\sqrt{\epsilon}/\lambda$ and $\rho_i \gg \beta$. Then the right side of (11) is rewritten,

$$(\beta^2 - \rho_i^2)\left(\frac{c}{2}\right)^2 \simeq -\left(\frac{c\rho_i}{2}\right)^2 \simeq \left(\frac{c\sqrt{\epsilon}}{\lambda}\right)^2, \tag{12}$$

where $\lambda/\sqrt{\epsilon}$ is the effective wavelength inside the dielectric. Combining (12) and (11) we can see that if the thickness c is much greater than the effective wavelength in the medium $\lambda/\sqrt{\epsilon}$, then $c\zeta_i/2$ approaches $\pi/2$, and therefore $(c/\pi)\zeta_i$ approaches unity. This means that $\pi/\zeta_i = c$ will be a good approximation if $c \gg \lambda/\sqrt{\epsilon}$.

This explains that if c is much greater than $\lambda/\sqrt{\epsilon}$, π/ζ_i is very nearly equal to c, then the boundary surface can be treated as a node of the standing waves inside the medium. This is why the first-order approximation for the frequency equation, as treated in Section II, fits well for the higher order modes. The further physical meaning of these equations can be more clearly understood from Fig. 2. If the resonator is thin, as compared with $\lambda/\sqrt{\epsilon}$, the effective wavelength in the dielectrics, then more E field extends outside the resonator.

IV. A Second-Order Approximation for the H Mode

We will now confine the problem to the fundamental H mode $TM_{11\delta}$, and discuss the details of the three-dimensional field inside a resonator for the three different cases of (3).

$TM_{\parallel\delta}$ means that the E field varies a half cycle along the x and y axes, and varies only the fraction $\delta=(c/\pi)\zeta_i$ of the sine wave along the z axis inside a resonator. The E field inside the resonator can be written in the following form:

$$E_{xi} = A_x(z)\cos\xi x\sin\eta y e^{i\omega t},$$
$$E_{yi} = A_y(z)\sin\xi x\cos\eta y e^{i\omega t},$$
$$E_{zi} = 0. \tag{13}$$

After substituting the above equation into Maxwell's equation (1) we have

$$\frac{d^2A_x}{dz^2} + (\epsilon_x\beta^2 - \eta^2)A_x + \xi\eta A_y = 0,$$
$$\frac{d^2A_y}{dz^2} + (\epsilon_y\beta^2 - \xi^2)A_y + \xi\eta A_x = 0, \tag{14}$$
$$\xi\frac{dA_x}{dz} = -\eta\frac{dA_y}{dz}.$$

One can express a solution by means of two orthogonal modes with different propagation constants ζ_{i-} and ζ_{i+} in the following form:

$$A_x = C_-\sin\alpha_i e^{i\zeta_{i-}z} + C_+\cos\alpha_i e^{i\zeta_{i+}z},$$
$$A_y = C_-\cos\alpha_i e^{i\zeta_{i-}z} - C_+\sin\alpha_i e^{i\zeta_{i+}z}. \tag{15}$$

Here, α_i is the angle between the principal axis of the orthogonal mode and the x and y axes. Then the propagation constants $\zeta_{i\pm}$ along the z axis should satisfy the relation

$$\zeta_{i\pm}^2 = \tfrac{1}{2}\left|\beta^2(\epsilon_x+\epsilon_y) - (\xi^2+\eta^2)\right.$$
$$\left. \pm\left\{\left[\beta^2(\epsilon_x-\epsilon_y) - (\xi^2-\eta^2)\right]^2 + 4\xi^2\eta^2\right\}^{1/2}\right|, \tag{16}$$

and the angle α_i should satisfy

$$\tan\alpha_i = -\frac{1}{2\xi\eta}\left\{\left[\beta^2(\epsilon_y-\epsilon_x) + (\eta^2-\xi^2)\right]\right.$$
$$\left. + \left[\left\{\beta^2(\epsilon_x-\epsilon_y) - (\xi^2-\eta^2)^2 + 4\xi^2\eta^2\right\}^{1/2}\right\}, \tag{17}$$

where $\xi=\pi l/a$, $\eta=\pi m/b$ are propagation constants along the x and y axes and l and m are integers.

For the three cases defined in Section II, (16) and (17) become:

Case I.

$$(\zeta_{i-}/\pi)^2 = \left(\frac{2}{\lambda}\right)^2\epsilon_\perp - \left[\left(\frac{l}{a}\right)^2 + \left(\frac{m}{b}\right)^2\right], \tag{18a}$$

where ζ_{i+} is eliminated from the solutions, because it does not satisfy the boundary conditions as we can see from (7);

$$\tan\alpha_i = -\frac{\eta}{\xi} = -\frac{ma}{lb}. \tag{18b}$$

Case II.

$$(\zeta_{i\pm}/\pi)^2 = \frac{1}{2}\left[\left(\frac{2}{\lambda}\right)^2(\epsilon_\parallel+\epsilon_\perp) - \left(\frac{l}{a}\right)^2 - \left(\frac{m}{b}\right)^2\right.$$
$$\left. \pm\left\{\left[\left(\frac{2}{\lambda}\right)^2(\epsilon_\parallel-\epsilon_\perp) + \left(\frac{l}{a}\right)^2 - \left(\frac{m}{b}\right)^2\right]^2 + 4\left(\frac{lm}{ab}\right)^2\right\}^{1/2}\right], \tag{19a}$$

$$\tan\alpha_i = -\frac{ab}{2lm}\left[\left(\frac{2}{\lambda}\right)^2(\epsilon_\perp-\epsilon_\parallel) - \left(\frac{l}{a}\right)^2 + \left(\frac{m}{b}\right)^2\right.$$
$$\left. + \left\{\left[\left(\frac{2}{\lambda}\right)^2(\epsilon_\parallel-\epsilon_\perp) + \left(\frac{l}{a}\right)^2 - \left(\frac{m}{b}\right)^2\right]^2 + 4\left(\frac{lm}{ab}\right)^2\right\}^{1/2}\right]. \tag{19b}$$

Case III.

$$(\zeta_{i\pm}/\pi)^2 = \frac{1}{2}\left[\left(\frac{2}{\lambda}\right)^2(\epsilon_\parallel+\epsilon_\perp) - \left(\frac{l}{a}\right)^2 - \left(\frac{m}{b}\right)^2\right.$$
$$\left. \pm\left\{\left[\left(\frac{2}{\lambda}\right)^2(\epsilon_\perp-\epsilon_\parallel) + \left(\frac{l}{a}\right)^2 - \left(\frac{m}{b}\right)^2\right]^2 + 4\left(\frac{lm}{ab}\right)^2\right\}^{1/2}\right], \tag{20a}$$

$$\tan\alpha_i = \frac{-ab}{2lm}\left[\left(\frac{2}{\lambda}\right)^2(\epsilon_\parallel-\epsilon_\perp) - \left(\frac{l}{a}\right)^2 + \left(\frac{m}{b}\right)^2\right.$$
$$\left. + \left\{\left[\left(\frac{2}{\lambda}\right)^2(\epsilon_\perp-\epsilon_\parallel) + \left(\frac{l}{a}\right)^2 - \left(\frac{m}{b}\right)^2\right]^2 + 4\left(\frac{lm}{ab}\right)^2\right\}^{1/2}\right]. \tag{20b}$$

Now let the field outside the resonator be such that the inside and outside fields are matched on the $x-y$ plane at $z=\pm c/2$.

Immediately outside the resonator the E field must have the same periodicity as the field just inside the resonator. Hence, we express the E field in the form

$$E_{xo} = B_x(z)\cos\xi x\sin\eta y e^{i\omega t},$$
$$E_{yo} = B_y(z)\sin\xi x\cos\eta y e^{i\omega t},$$
$$E_{zo} = 0. \tag{21}$$

Note that these equations are true only very close to the boundary.

The wave equation for the outside field is the same as (14) except for $\epsilon = 1$, and its solutions are

$$\begin{aligned} B_x &= D_- \sin \alpha_o e^{i\zeta_{o-}z} + D_+ \cos \alpha_o e^{i\zeta_{o+}z} \\ B_y &= D_- \cos \alpha_o e^{i\zeta_{o-}z} - D_+ \sin \alpha_o e^{i\zeta_{o+}z}, \end{aligned} \tag{22}$$

where

$$\begin{aligned} (\zeta_{o-}/\pi)^2 &= \left(\frac{2}{\lambda}\right)^2 - \left[\left(\frac{l}{a}\right)^2 + \left(\frac{m}{b}\right)^2\right], \\ (\zeta_{o+}/\pi)^2 &= \left(\frac{2}{\lambda}\right)^2, \\ \tan \alpha_o &= -\frac{\eta}{\xi} = -\frac{ma}{lb}. \end{aligned} \tag{23}$$

$i\zeta_{o-}$ is real and shows an exponentially damped wave outside of the resonator. On the other hand, $i\zeta_{o+}$ is imaginary and shows propagation without attenuation.

From (6) we have the boundary condition

$$\begin{aligned} A_x &= B_x\big|_{z=c/2} \qquad A_y = B_y\big|_{z=c/2} \\ \frac{\partial A_x}{\partial z} &= \frac{\partial B_x}{\partial z}\bigg|_{z=c/2} \qquad \frac{\partial A_y}{\partial z} = \frac{\partial B_y}{\partial z}\bigg|_{z=c/2}. \end{aligned} \tag{24}$$

Substituting (15) and (22) into (24), we have four linear equations for the C's and D's. To have a nonzero solution of C's and D's, the determinant derived from them must be equal to zero. After much painstaking algebra assuming

$$\left|\frac{\zeta_{o+}}{\zeta_{o-}}\right|^2 \simeq \frac{\left(\frac{2}{\lambda}\right)^2}{\left(\frac{l}{a}\right)^2 + \left(\frac{m}{b}\right)^2} \simeq \frac{1}{\epsilon} \simeq 0,$$

we have

$$\begin{aligned} \zeta_{o-}\left(\alpha\zeta_{i-}\tan\frac{c}{2}\zeta_{i-} + \beta\zeta_{i+}\tan\frac{c}{2}\zeta_{i+}\right) \\ \simeq \gamma\zeta_{i+}\zeta_{i-}\tan\frac{c}{2}\zeta_{i+}\tan\frac{c}{2}\zeta_{i-}, \end{aligned} \tag{25}$$

where

$$\begin{aligned} \alpha &= \sin^2(\alpha_i - \alpha_o), \\ \beta &= \cos^2(\alpha_i - \alpha_o), \\ \gamma &= 1. \end{aligned}$$

For Case I, (25) is much simpler and becomes

$$\zeta_{o-} = \zeta_{i-}\tan\zeta_{i-}\frac{c}{2}. \tag{26}$$

Of course this result agrees with (14) of the two-dimensional case, as one expected.

Comparison with the experimental results will be discussed in Section VII.

V. A Second-Order Approximation for the E Mode

There is another set of modes which also satisfies the boundary conditions. In this case, the roles of E and H are interchanged from the H-mode case and the outside field is similar to an electric dipole. Hence, we call this case the E or TE mode. One can see that the field configuration inside the resonator is similar to the rectangular metal-wall cavity. The fundamental mode is called the $TE_{11\delta}$ for reasons analogous to the TE_{111} case. The calculations are similar to the H-mode case and are carried out in a similar way.

For the fundamental $TE_{11\delta}$ mode, we assume the H field, in analogy to (13), is of the form

$$\begin{aligned} H_{xi} &= L_x(z)\cos\xi x \sin\eta y e^{i\omega t}, \\ H_{yi} &= L_y(z)\sin\xi x\cos\eta y e^{i\omega t}, \\ H_{zi} &= 0, \end{aligned} \tag{27}$$

where

$$\begin{aligned} L_x &= (M_-\cos\alpha_i\cos\zeta_{i-}z + M_+\sin\alpha_i\cos\zeta_{i+}z) \\ L_y &= (M_-\sin\alpha_i\cos\zeta_{i-}z - M_+\cos\alpha_i\cos\zeta_{i+}z). \end{aligned} \tag{28}$$

The solutions for the ζ_i and the angle α_i for the three cases are:

Case I.

$$(\zeta_{i-}/\pi)^2 = \epsilon_\perp\left(\frac{2}{\lambda}\right)^2 - \frac{\epsilon_\perp}{\epsilon_\parallel}\left[\left(\frac{l}{a}\right)^2 + \left(\frac{m}{b}\right)^2\right],$$

or

$$\begin{aligned} (2/\lambda)^2 &= \frac{1}{\epsilon_\parallel}\left[\left(\frac{l}{a}\right)^2 + \left(\frac{m}{b}\right)^2\right] + \frac{1}{\epsilon_\perp}(\zeta_{i-}/\pi)^2, \\ (\zeta_{i+}/\pi)^2 &= \epsilon_\perp\left(\frac{2}{\lambda}\right)^2. \end{aligned}$$

$$\tan\alpha_i = -\frac{\eta}{\xi} = -\frac{ma}{lb}. \tag{29a}$$

Case II.

$$\begin{aligned} (\zeta_{i\pm}/\pi)^2 = \frac{1}{2}\Bigg\{(\epsilon_\perp + \epsilon_\parallel)\left(\frac{2}{\lambda}\right)^2 - \left(\frac{m}{b}\right)^2 - \frac{\epsilon_\parallel}{\epsilon_\perp}\left(\frac{l}{a}\right)^2 \\ \pm\Bigg[\left\{\left(\frac{m}{b}\right)^2 - \frac{\epsilon_\parallel}{\epsilon_\perp}\left(\frac{l}{a}\right)^2 + \left(\frac{2}{\lambda}\right)^2(\epsilon_\perp - \epsilon_\parallel)\right\}^2 \\ + \frac{4\epsilon_\parallel}{\epsilon_\perp}\left(\frac{lm}{ab}\right)^2\Bigg]^{1/2}\Bigg\} \end{aligned}$$

$$\begin{aligned} \tan\alpha_i = \frac{1}{2}\frac{ab}{lm}\Bigg\{\frac{\epsilon_\parallel}{\epsilon_\perp}\left(\frac{l}{a}\right)^2 - \left(\frac{m}{b}\right)^2 + \left(\frac{2}{\lambda}\right)^2(\epsilon_\parallel - \epsilon_\perp) \\ - \Bigg[\left\{\frac{\epsilon_\parallel}{\epsilon_\perp}\left(\frac{l}{a}\right)^2 - \left(\frac{m}{b}\right)^2 + \left(\frac{2}{\lambda}\right)^2(\epsilon_\parallel - \epsilon_\perp)\right\}^2 \\ + \frac{4\epsilon_\parallel}{\epsilon_\perp}\left(\frac{lm}{ab}\right)^2\Bigg]^{1/2}\Bigg\}. \end{aligned} \tag{29b}$$

Case III.

$$(\zeta_{i\pm}/\pi)^2 = \frac{1}{2}\left\{(\epsilon_\perp + \epsilon_\parallel)\left(\frac{2}{\lambda}\right)^2 - \frac{\epsilon_\parallel}{\epsilon_\perp}\left(\frac{m}{b}\right)^2 - \left(\frac{l}{a}\right)^2 \pm \left[\left\{\frac{\epsilon_\parallel}{\epsilon_\perp}\left(\frac{m}{b}\right)^2 - \left(\frac{l}{a}\right)^2 + \left(\frac{2}{\lambda}\right)^2(\epsilon_\parallel - \epsilon_\perp)\right\}^2 + \frac{4\epsilon_\parallel}{\epsilon_\perp}\left(\frac{lm}{ab}\right)^2\right]^{1/2}\right\}$$

$$\tan \alpha_i = \frac{1}{2}\frac{ab}{lm}\left\{\left(\frac{l}{a}\right)^2 - \frac{\epsilon_\parallel}{\epsilon_\perp}\left(\frac{m}{b}\right)^2 + \left(\frac{2}{\lambda}\right)^2(\epsilon_\perp - \epsilon_\parallel) - \left[\left\{\left(\frac{l}{a}\right)^2 - \frac{\epsilon_\parallel}{\epsilon_\perp}\left(\frac{m}{b}\right)^2 + \left(\frac{2}{\lambda}\right)^2(\epsilon_\perp - \epsilon_\parallel)\right\}^2 + 4\frac{\epsilon_\parallel}{\epsilon_\perp}\left(\frac{lm}{ab}\right)^2\right]^{1/2}\right\}. \quad (29c)$$

Then, using boundary conditions analogous to (24) and assuming the outside field is the same as (22) except that the E and H fields are interchanged, we have a boundary equation of the same form as (25), but with the following coefficients:

Case II.

$$\begin{aligned}\alpha &= (\epsilon_\perp \sin\alpha_i \sin\alpha_o - \epsilon_\parallel \cos\alpha_i \cos\alpha_o)\\ &\quad\cdot(\sin\alpha_i \sin\alpha_o - \cos\alpha_i \cos\alpha_o),\\ \beta &= (\epsilon_\parallel \cos\alpha_i \cos\alpha_o + \epsilon_\perp \sin\alpha_i \cos\alpha_o)\\ &\quad\cdot(\cos\alpha_i \sin\alpha_o + \sin\alpha_i \cos\alpha_o),\\ \gamma &= \epsilon_\parallel \cos^2\alpha_i + \epsilon_\perp \sin^2\alpha_i. \end{aligned} \quad (30)$$

VI. Physical Picture of the Field

Let us consider the $TM_{11\delta}$ mode of Case I. The propagation constants along the z axis for inside and outside fields are given by (18) and (23). For Case I, $\alpha_i = \alpha_o$ and (14) and (15), we have

$$\frac{C_+}{C_-} = \frac{\zeta_{i-}(\xi \sin\alpha_i - \eta\cos\alpha_i)}{\zeta_{i+}(\eta \sin\alpha_i + \xi\cos\alpha_i)} e^{i(\zeta_{i-}-\zeta_{i+})z}$$

but since $\tan\alpha_i = -\eta/\xi$, we have $C_+ = 0$. Similarly, for the outside field,

$$\frac{D_+}{D_-} = \frac{\zeta_{o-}(\xi \sin\alpha_o - \eta\cos\alpha_o)}{\zeta_{o+}(\eta \sin\alpha_o - \xi\cos\alpha_o)} e^{i(\zeta_{o-}-\zeta_{o+})z}.$$

Hence $D_+ = 0$, also. Therefore, only the C_- inside field couples to the D_- outside field, the attenuated wave, and not to the propagation wave D_+. On the other hand, in Cases II and III the situation becomes more complicated because of the anisotropy in the x-y plane. Both the C_- and C_+ waves of different propagation constants couple to the outside attenuation D_- wave and the propagation D_+ wave. An elliptically polarized wave is excited inside the resonator. Larger anisotropy induces more excitation of the propagation D_+ wave outside the resonator because $(\alpha_i - \alpha_o)$ becomes larger.

Coupling schemes of the resonator as a whole can be classified in the following two cases. The first is the coupling of the inside resonator field to the outside free-space field or to the waveguide field via the radiation field (dipole field, etc.) of the resonator. Hence, the coupling of this case depends upon both the coupling between the inside resonator field and the radiation field and the coupling between the radiation field and the free-space field. The first coupling is due to the perpendicular component of the Poynting vector on the resonator surface, which will be induced by a) distortion of the basic mode field and b) the presence of local higher-order modes. The second is the coupling of the inside field directly to the outside field by the C_+ and propagation D_+ modes. Let us consider the coupling of the dipole radiation field to the outside free-space field.

In the case of a magnetic dipole, the radiation impedance [9] of a loop in free space is given by

$$R_E = 7.9 \times 10^2 \left(\frac{R_1}{\lambda}\right)^2 \text{ ohms.} \quad (31a)$$

In our case, the radius of the loop R_1 corresponds to the resonator dimensions. For example, a rutile resonator measuring $0.123 \times 0.247 \times 0.470$ inches has a resonant wavelength of 6.83 cm. Then we have

$$R_H \cong 75 \text{ ohms.}$$

This impedance is about the same order of magnitude as free-space or waveguide impedance. On the other hand, the E dipole impedance is expressed as [9]

$$R_E \cong 3 \times 10^{-27}\left(\frac{\lambda}{R_1}\right)^4 \text{ ohms.} \quad (31b)$$

This is a very small number which makes coupling to free space almost impossible.

The exact calculation of the outside radiation field is quite complicated and is beyond the scope of our discussion in this paper. However, one can draw a picture of fields for either the Case I or isotropic resonator, depending upon the following statements:

a) Only a monotonically decreasing mode D_- exists at the outside of the resonator. It will be transferred smoothly to the outside dipole field.
b) The size of a resonator is quite small compared to the wavelength; hence, a resonator can be considered as nearly a point source radiator.
c) The direction of the E or H field at the origin of the radiator should be parallel to the z axis because of the boundary condition.

For these reasons, one can conclude that the field is nearly a dipole E or H field, as shown in Fig. 3. On the other hand, the radiation fields of higher modes are quite complicated but can be understood by comparison to the radiation pattern of a multipole antenna, which is somewhat similar to that case.

If the nonattenuating mode, which is due to the D_+

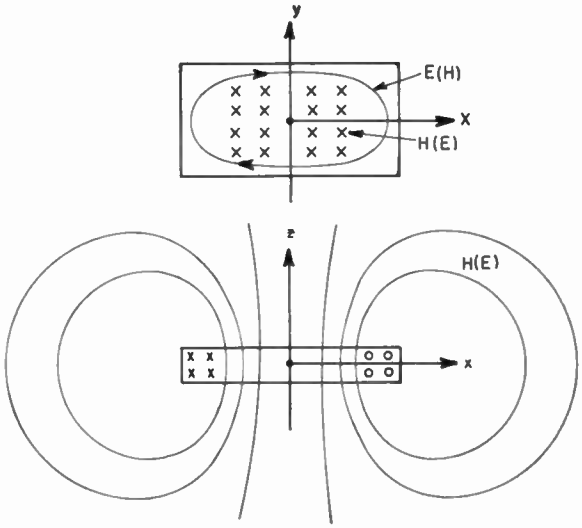


Fig. 3—Dipole radiation field of a resonator which resonates at fundamental anode.

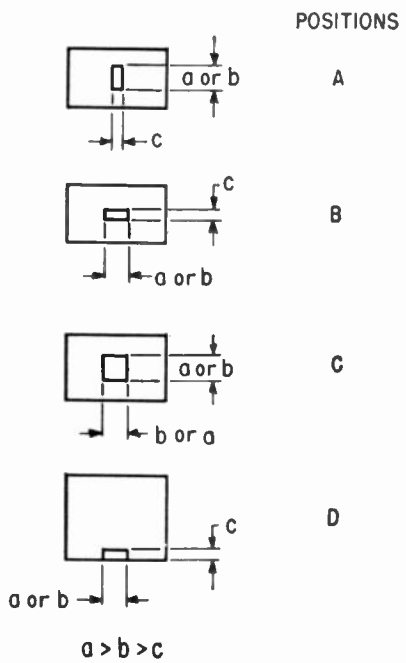


Fig. 4—Variable position of a resonator in the waveguide.

term, is superimposed on the above mentioned radiation mode, the situation becomes more complicated. This situation was observed as a change of coupling of a resonator when the Case II and Case III resonators in position A in Fig. 4 were rotated about the z axis. Other aspects of the resonator, such as Q values, will be discussed in another paper [12].

VII. Experimental Results

Experimental verification of the theory and accurate measurement of the dielectric constant ϵ, particularly in single crystal rutile, were made. The measurements of $\epsilon_{\parallel}$ in an anisotropic medium, such as rutile in the microwave region, are not reliable if one uses the conventional techniques. If the wave travels several wavelengths in the medium, the polarization plane of the incoming wave rotates about the direction of propagation and also changes its feature so that the effective path length is much greater than the physical length. This is especially important for the case in which the optical axis is perpendicular to the direction of propagation.

Before discussing the experimental results, the experimental procedure will be described. A piece of dielectric of known dimensions, resonant at microwave frequencies, was placed in a waveguide and supported in polyfoam. This polyfoam support had negligible loss and a dielectric constant very close to unity so that it did not disturb the fields surrounding the resonator.

The resonator was well isolated from both generator and detector by attenuators. The coupling of the resonator was adjusted by changing its orientation in the waveguide, about which more will be said later. It was also kept as far from the waveguide walls as possible to avoid distortion of the radiation field, and to compare with theoretical results based on a dielectric resonator in free space. As a check, the resonators were put in an oversized waveguide, or between waveguide horns. Only small frequency differences for either case were observed. The set-up is shown in Fig. 5(a). Crystal detectors #1 and #2 were used to monitor the reflected and transmitted power. The standing-wave detector was used to check the matching at resonance.

Fig. 5(b) shows the set-up used to measure the temperature effect on ϵ. One might expect a large standing wave to be caused by the shorting plunger, but in the experiments the QL of the resonators was so much higher than that of the standing waves due to the short that the two were easily distinguishable. The short served as a tuner to help match the resonator, whose position was fixed in the low temperature bath. This system was used at frequencies from 8 to 40 kMc and at temperatures from room temperature down to 78°K. At liquid helium temperatures, the set-up shown in Fig. 5(c) was used for frequencies below 8 kMc. ϵ vs temperature curves for rutile and $SrTiO_3$ are shown in Fig. 6.

For the Q measurements, detector #2 and an attenuator in Fig. 5(a) were replaced by a sliding short. The crystal resonator was critically coupled by adjusting the short, and the Q_o was measured at this point. The reflected wave from the crystal resonator was observed at detector #1. The position of the crystal resonator in the waveguide for maximum coupling was more or less opposite for the E and H modes so that the two types of modes were distinguished in this way. One can observe a variety of absorption or reflection patterns due to E or H mode resonance for high-order modes.

Sometimes our Q measurements were inaccurate because the waveguide wall, which induces skin loss, was not far enough away from the crystal resonator.

The rutile resonators were cut from single-crystal boules, which were optically oriented to an accuracy of

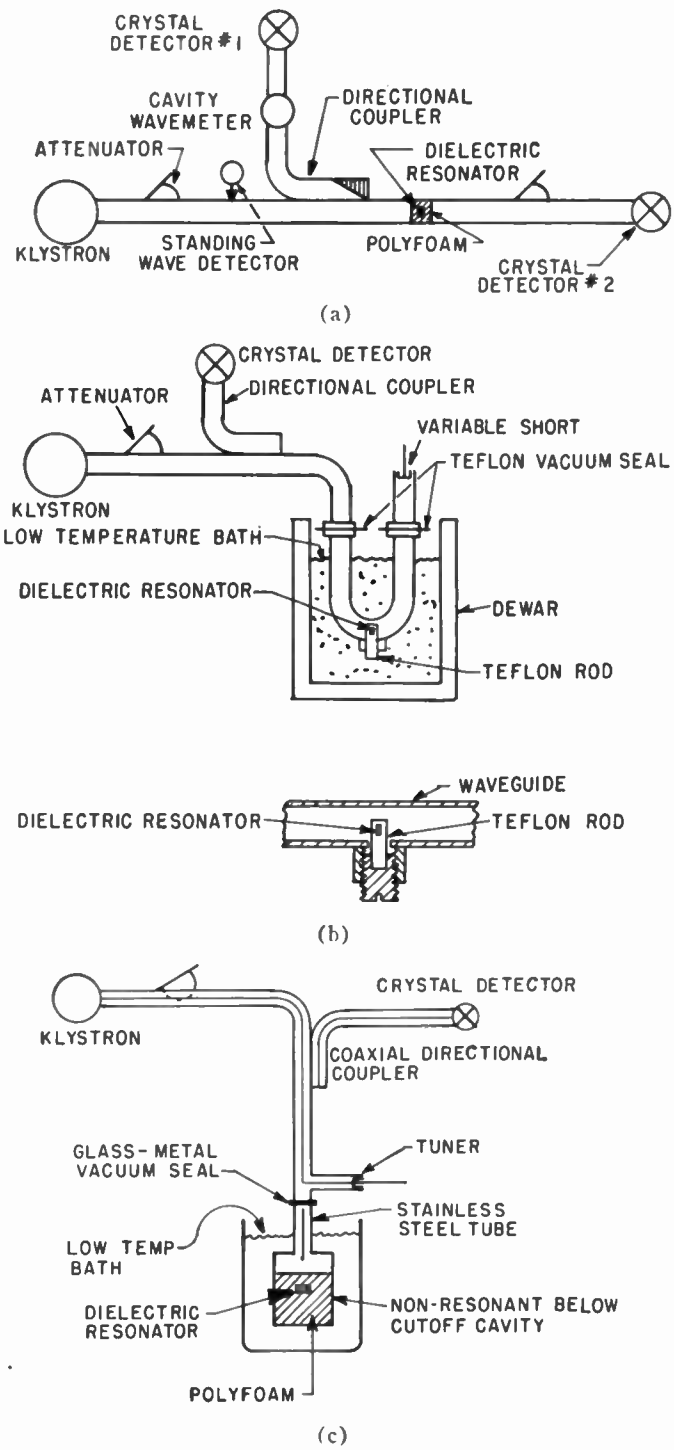


Fig. 5—(a) Microwave set-up for characteristics measurements. (b) Microwave set-up for temperature effect measurements. (c) Microwave set-up for temperature effect measurements.

$\pm 1.5°$ and annealed in an air oven at about 700°C for half an hour. Usually this annealing was quite effective in increasing the Q values. Quite often "pure rutile" has shown different Q values from piece to piece because even a small population of vacancies or dislocations influence the Q values. The $SrTiO_3$ resonators were also cut from the optically oriented boules. Even though the local crystal field is nearly cubic, a birefringent effect [4] was observed at room temperature. $SrTiO_3$ resonators showed large natural relaxation characteristics, and their Q values were increased by approximately a factor two, several days after machining.

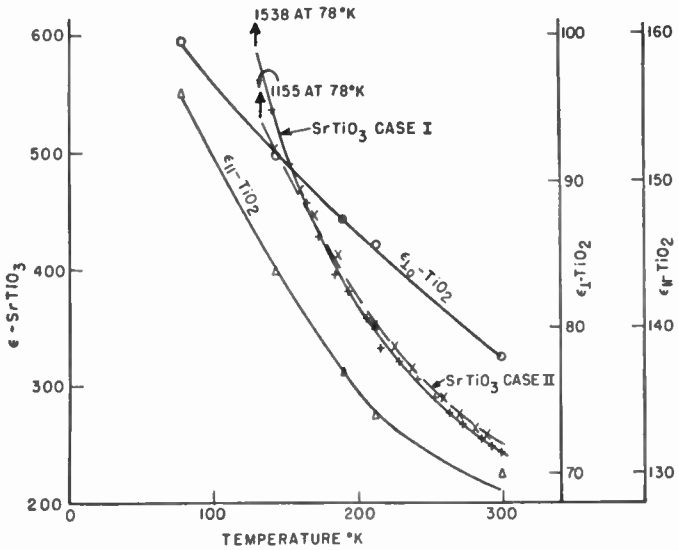


Fig. 6—Temperature vs frequency curve of TiO_2 and $SrTiO_3$.

To compare the theoretical and experimental values the following two experiments were conducted:

a) Accurate measurements of the resonant frequencies were made on accurately finished C band Case I, II and II resonators of size $0.1233'' \times 0.4075'' \times 0.2466'' \pm 0.0003''$. In this case the accuracy of the orientation of the optical axis, referred to the geometrical axis, was better than $\pm 1.5°$.

b) The resonant frequency vs the resonator thickness curve was obtained on X band Case I and II resonators of approximate size $0.20'' \times 0.10'' \times c$.

In experiment a) of the Case I resonator, one can see that the second-order equation (18a) completely agrees with the first-order equation (4) and the boundary equation. Then $\epsilon_\perp$ was obtained from the H-mode frequency, (18a) which depends only upon $\epsilon_\perp$. $\epsilon_\parallel$ was obtained from the E-mode frequency (29a) using $\epsilon_\perp$ and ζ_{i-} values. All twelve measured frequency values agreed with the values calculated by accurate frequency equations using the obtained $\epsilon_\perp$, $\epsilon_\parallel$ and ζ_{i-} constants.

Discrepancies were less than 1 per cent. $\epsilon_\perp$ and $\epsilon_\parallel$ were 78.3 and 127, respectively. These values were also confirmed by applying them to the X band Case I resonators of different frequency and dimensions. The change of $\epsilon_\perp$ and $\epsilon_\parallel$ values between the X and C bands was not noticed. It will be noted that some of the E modes, such as the $E_{11\delta}$, $E_{31\delta}$, and $E_{51\delta}$, were not observed in the experiment. This lack of observation seems to be due to the lack of coupling between the inside and outside resonator fields or to the suppression of these resonance modes by the waveguide field.

The curve obtained from experiment b) of the Case I resonator is shown in Fig. 7. Good agreement of experimental results with the values obtained from the bound-

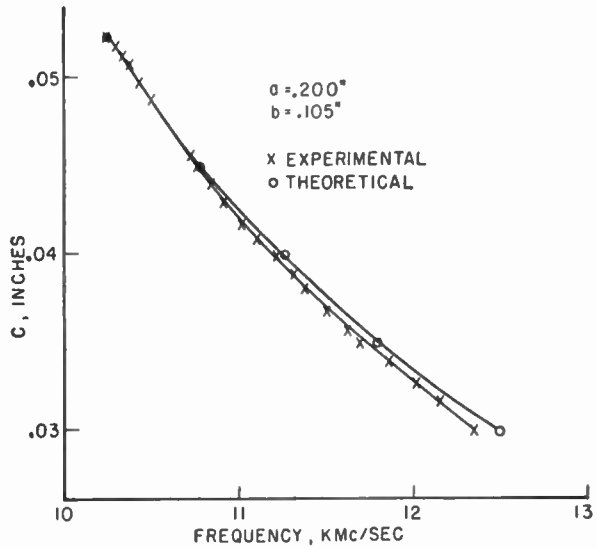


Fig. 7—Thickness vs frequency curve of TiO_2 Case I resonator.

TABLE I
CASE I

Experimental			Theoretical	
Outside field*	Wavelength cm	Q_0(300°K)	Mode	Wavelength cm†
H	9.31	14,000	$H_{11\delta}$	9.31
(E)	(9.31)		($E_{21\delta}$)	(9.31)
H	6.05	16,000	$H_{21\delta}$	6.05
			($E_{31\delta}$)	(6.00)
E	5.24	28,000	$E_{22\delta}$	5.26
E	5.15	5,000	$E_{41\delta}$	5.16
	5.05		$H_{31\delta}$	5.10
E	5.04		E_{212}	4.98
	4.79		$E_{32\delta}$	4.85
	4.77		$H_{12\delta}$	4.75
	4.45		$H_{22\delta}$	4.45
			($E_{51\delta}$)	(4.46)
	4.40		$E_{42\delta}$	4.36
	4.35		$E_{41\delta}$	4.32

* E means electric mode; H means magnetic mode.
† $\epsilon_\perp = 79$, $\epsilon_\parallel = 127$, obtained experimentally, were used.
The dimensions of the crystals are: $a = 0.471$ inch, $b = 0.247$ inch, $c = 0.123$ inch.

ary (26) shows the validity of the assumption of the boundary field on both the E and H modes of Case I resonators (Table I).

The results of experiment a) of the H mode on the much more complicated Cases II and III are listed in Tables II and III. The error at those fundamental modes was 2 per cent.

The curve of experiment b) of the H-fundamental mode of Case II is shown in Fig. 8. This also shows the validity of the boundary and frequency (19) and (20) of this case. On the other hand, the results for the fundamental E mode were not nearly as good as those obtained for the H mode and showed a discrepancy of 20 per cent. The reason for this seems to be the oversimplified boundary field which we used. The effective thickness, $c' = \pi/\zeta_i$ (see Fig. 2), calculated from the boundary equations (25) and (30), is much larger than the c' value which satisfies the experimental results. This

TABLE II
CASE II

Experimental			Theoretical		
Outside field*	Wavelength cm	Q_0(300°K)	Mode	Wavelength cm First-order approx.	Second order approx.
H	8.71		$H_{11\delta}$	7.77	8.53
E	6.86	10,000	$E_{11\delta}$	6.94	
E	6.56	6,000	$E_{21\delta}$	6.75	
H	6.25	16,000	$H_{21\delta}$	6.58	
E	5.90	8,000	$E_{31\delta}$	5.54	
H	5.46	6,000	$H_{22\delta}$	5.13	
H	5.41	18,000	$H_{31\delta}$	5.41	
E	5.14	6,000	E_{212}	5.18	

Mode	E	H
$(\zeta_i/\pi)^2(1/cm^2)$	1.63	4.80

TABLE III
CASE III

Experimental			Theoretical		
Outside field*	Wavelength cm	Q_0(300°K)	Mode	Wavelength cm First-order approx.	Second-order approx.
H	7.35	4,000	$H_{11\delta}$	6.76	7.50
E	6.95	18,000	$E_{11\delta}$	7.94	
E	6.27	10,000	$E_{21\delta}$	6.65	
H	6.17	5,000	$H_{21\delta}$	6.20	
E	5.59	10,000	$E_{31\delta}$	5.46	
H	5.44	4,000	$H_{31\delta}$	5.49	
E	4.97		$E_{21\delta}$	4.98	

Mode	E	E
$(\zeta_i/\pi)^2(1/cm^2)$	1.86	6.51

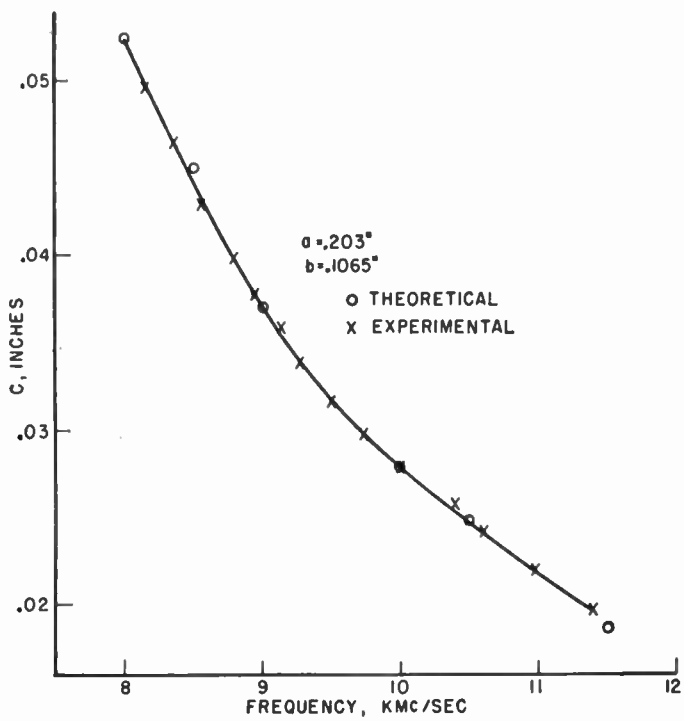


Fig. 8—Thickness vs frequency curve of TiO_2 Case II resonator.

shows that the *H* field is much more concentrated in the resonator than we assumed.

It should also be noted that these second-order equations are accurate only at a fundamental mode, because the boundary field for high modes can be expected to differ from the field of (27) and (28), except in Case I for which it is nearly true. The modes of high-mode resonances of Cases II and III were identified by the first-order frequency (4).

The constants $\zeta_i - /\pi$ for each *E* and *H* mode were chosen such that (4) fits the experimental values, especially at higher modes, as explained in Section III.

The modes were also identified by the difference of coupling characteristics between the *H* and *E* modes during the measurements. These results are shown in Tables II and III. The mode frequency patterns of all three cases are shown in Fig. 9.

The simpler case of isotropic media, *e.g.*, single crystals of $SrTiO_3$ and KRS-5 (ThBr-ThI), was also examined. The frequency equation for this case for low modes is

$$\left(\frac{2}{\lambda}\right)^2 = \frac{1}{\epsilon}\left[\left(\frac{l}{a}\right)^2 + \left(\frac{m}{b}\right)^2 + \left(\frac{n}{c'}\right)^2\right], \qquad c' = \frac{\pi}{\zeta_i}$$

coupled with the boundary equation

$$\zeta_o = \zeta_i \tan \zeta_i \frac{c'}{2},$$

where

$$(\zeta_o/\pi)^2 = \frac{1}{a^2} + \frac{1}{b^2} - \left(\frac{2}{\lambda}\right)^2.$$

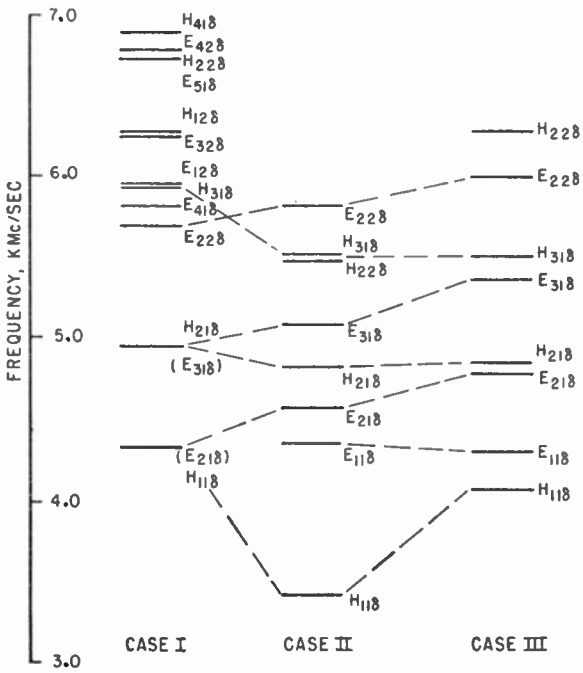


Fig. 9—Mode, frequency pattern of the resonator.

For the higher modes we have

$$\left(\frac{2}{\lambda}\right)^2 = \frac{1}{\epsilon}\left[\left(\frac{l}{a}\right)^2 + \left(\frac{m}{b}\right)^2 + \left(\frac{n}{c}\right)^2\right].$$

The mode behavior of the outside fields was all *H* type with no exception. This is because the *H*-pole field couples so much more strongly to the waveguide field than does the *E*-pole field, as discussed before. The experimentally obtained values at room temperature were

	ϵ	$\tan \delta$	Frequency	Manufacturer
$SrTiO_3$	250.0	0.001	8–25 kMc	National Lead Co.
KRS-5	27.5	0.002	4 kMc	Harshaw Chemical Co.

For $SrTiO_3$, ϵ is strongly temperature dependent and also shows a slight, but not negligible, anisotropy at room temperature. This is shown in Fig. 6. There is also a nonlinear power effect at the microwave frequencies [4] at which the measurements were made. An extremely high *Q* resonance, of the order of one million, was observed in a KRS-5 resonator at *X* band range in a 1.4°K helium bath.

VIII. Band-Pass and Band-Reject Filters Using Dielectric Resonators

Dielectric resonators can be used as the resonant elements as shown in Fig. 4, simply by placing them in the waveguide at various positions. If a piece of high dielectric constant material, such as rutile, is supported inside a waveguide, a small VSWR, less than 1.3, results for any orientation of the crystal. In the guide size appropriate to low mode resonant frequencies, the crystal at nonresonant frequencies acts merely as a small piece of low-loss insulator. If the crystal is adjusted for critical coupling at resonance, it will absorb all the incident power at resonance, making an absorptive band-reject filter. The crystal can also be adjusted for large mismatches at resonance frequencies, of the order of VSWR equal to 25, such that most of the power is reflected at resonance. In Fig. 4, position *A* shows maximum coupling for *H* modes, position *B* shows maximum coupling for *E* modes, minimum for *H* modes, and position *C* gives intermediate coupling for both. If the crystal is allowed to rest on the waveguide wall as shown in Fig. 4(D), the resonant frequency is changed (see Appendix II) and also, the loss of the waveguide wall at resonance makes Q_o lower. But this is the simplest filter, and the reflection from the resonator is very small.

The *Q* value of the reject band of the filter is normally of the order of several thousand for rutile at room temperature, but this can be decreased by doping the crystal, reducing it in a hydrogen atmosphere, or by using more than one crystal and stagger tuning. An increase of *Q* can be achieved by choosing a stress-free crystal or by lowering the temperature of the resonator. The *Q* and resonant frequency of a resonator of rutile doped

with a paramagnetic salt such as Cr^{3+}, Fe^{3+}, or Mn^{2+} can be changed with an external magnetic field by the electron paramagnetic resonance effect, if the input power is of the order of 0.1 mw, but this requires low temperature operation (78°K).

A type of band-pass filter is illustrated in Fig. 10. Sections of crossed input and output rectangular waveguides are tapered to a short section of a square waveguide. The crystal, in a polyfoam support, is placed along a diagonal of the square section so as to achieve critical coupling to both input and output guides at resonance. The radiation field of the resonator is symmetric along the diagonal of the square waveguide and, hence, couples to both input and output equally. The coupling was adjusted by rotating and moving the crystal along the diagonal. A piece of attenuator card is placed in the output section, as shown, to absorb the incident power-off resonance. One such filter roughly constructed at X band transmitted 60 per cent of the incident power at a resonance of $Q=10{,}000$ with a VSWR no worse than 1.1 anywhere. Off resonance the transmitted power was down at least 35 db. These figures could easily be improved by careful adjustment.

Appendix I

Maxwell's equations,

$$\nabla \times \boldsymbol{E} \equiv -\mu \dot{\boldsymbol{H}}, \qquad \nabla \times \boldsymbol{H} = \epsilon \dot{\boldsymbol{E}}, \tag{32}$$

will be solved for the case in which μ and ϵ are anisotropic, diagonal tensors. The solutions are assumed to be of the following forms:

$$H_x = h_x \begin{matrix}\sin\\ \cos\end{matrix} \xi x \begin{matrix}\sin\\ \cos\end{matrix} \eta y \begin{matrix}\cos\\ \sin\end{matrix} \zeta z e^{i\omega t},$$

$$H_y = h_y \begin{matrix}\cos\\ \sin\end{matrix} \xi x \begin{matrix}\cos\\ \sin\end{matrix} \eta y \begin{matrix}\cos\\ \sin\end{matrix} \zeta z e^{i\omega t},$$

$$H_z = h_z \begin{matrix}\cos\\ \sin\end{matrix} \xi x \begin{matrix}\sin\\ \cos\end{matrix} \eta y \begin{matrix}\sin\\ \cos\end{matrix} \zeta z e^{i\omega t}. \tag{33}$$

The same sort of equations and solutions can be derived for the E field.

From the condition that the determinant obtained from (32) and (33) must be equated to zero for a solution to exist, the following frequency equation is obtained.

$$\left(\frac{2}{\lambda}\right)^2 = \left(\frac{1}{a}\right)^2\left(\frac{1}{\mu_y\epsilon_z}+\frac{1}{\mu_z\epsilon_y}\right)^2+\left(\frac{1}{c}\right)\left(\frac{1}{\mu_x\epsilon_z}+\frac{1}{\mu_z\epsilon_x}\right)+\left(\frac{1}{b}\right)^2\left(\frac{1}{\mu_x\epsilon_y}+\frac{1}{\mu_y\epsilon_x}\right)$$

$$\pm\frac{1}{2}\Bigg[\left(\frac{1}{a}\right)^4\left(\frac{1}{\mu_y\epsilon_z}-\frac{1}{\mu_z\epsilon_y}\right)^2+\left(\frac{1}{c}\right)^4\left(\frac{1}{\mu_z\epsilon_x}-\frac{1}{\mu_x\epsilon_z}\right)^2+\left(\frac{1}{b}\right)^4\left(\frac{1}{\mu_y\epsilon_x}-\frac{1}{\mu_x\epsilon_y}\right)^2$$

$$+2\left(\frac{1}{a}\right)^2\left(\frac{1}{c}\right)^2\left(\frac{1}{\mu_z\epsilon_x}-\frac{1}{\mu_x\epsilon_z}\right)\left(\frac{1}{\mu_z\epsilon_y}-\frac{1}{\mu_y\epsilon_z}\right)+2\left(\frac{1}{a}\right)^2\left(\frac{1}{b}\right)^2\left(\frac{1}{\mu_y\epsilon_x}-\frac{1}{\mu_x\epsilon_y}\right)$$

$$\times\left(\frac{1}{\mu_y\epsilon_z}-\frac{1}{\mu_z\epsilon_y}\right)+2\left(\frac{1}{c}\right)^2\left(\frac{1}{b}\right)^2\left(\frac{1}{\mu_z\epsilon_x}-\frac{1}{\mu_x\epsilon_z}\right)\left(\frac{1}{\mu_y\epsilon_x}-\frac{1}{\mu_x\epsilon_y}\right)\Bigg]^{1/2}.$$

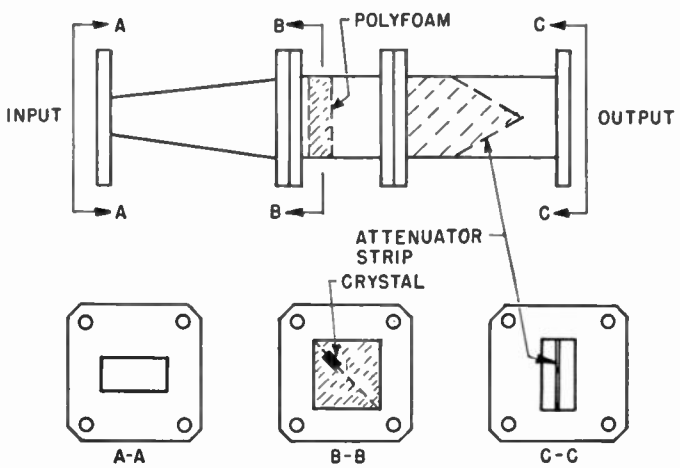


Fig. 10—Band-pass filter.

Here a, b, and c are the resonator dimensions along the x, y, and z directions, respectively. The same result was obtained from the equation of the E field. This is the zero-order solution, and it is assumed that the surface of the resonator coincided with the node or antinode of the field. Its limitations are discussed in the main article.

Appendix II

The Effect of a Metal Wall on a Dielectric Resonator

The effect of a metal wall on a dielectric resonator can be solved readily only for an isotropic, spherical resonator in a concentric metal sphere. This work was done by J. A. Giordmaine [10], and the results are shown in Fig. 11. One can see from Fig. 11 that, if the distance

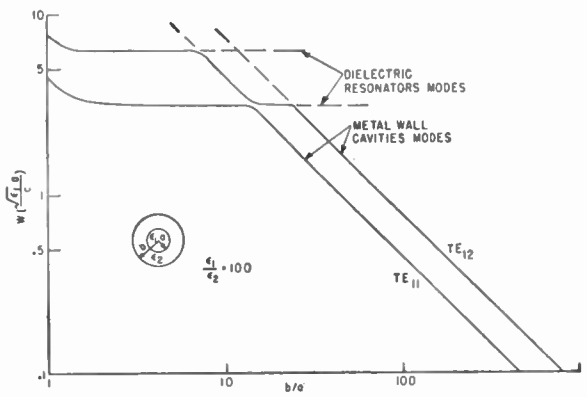


Fig. 11—The effect of a metal wall on the dielectric resonator.

between the surface of a cavity and a dielectric sphere is small, the resonant frequency of the dielectric sphere is influenced by the metal wall. However, if the gap between them becomes larger than a certain value, the dielectric sphere behaves as an independent resonator and has a constant frequency. If the resonant frequencies of two resonators are nearly the same, interference becomes significant and their frequencies shift by an amount proportional to the ratio ϵ_2/ϵ_1. Hence, the clear separation between the dielectric resonator modes and metal-cavity modes can be seen for large ratios of ϵ_1/ϵ_2. If this ratio is extremely large, the interaction is much smaller and the intersection of the mode lines becomes nearly a point.

Similar considerations would hold for rectangular dielectric resonators in a rectangular waveguide, although the results cannot be as readily calculated as in the spherical case.

Acknowledgment

The authors wish to thank Prof. C. H. Townes for his encouragement and the members of the Columbia Radiation Laboratory for useful discussions. We are also thankful to A. Marshall, whose assistance was indispensable in preparing the samples.

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CORRECTION

J. Ralph Johler, author of "Propagation of the Low-Frequency Radio Signal," which appeared on pages 404–427 of the April, 1962, issue of PROCEEDINGS, has called the following to the attention of the *Editor*.

On page 408, the line immediately after (31) should read:

"where $\tau=\tau_s$ comprises the special roots of the differential equation of Riccati, noted by Bremmer . . . ."

On page 411, (42) should read:

$$\begin{aligned} -\operatorname{Im} F_s(t) &= \operatorname{Re} i \exp(-\nu t) \\ &= \exp(-c_1 t)\sin \omega_c t \qquad (0 < t < \infty) \\ &= 0 \qquad (t < 0). \end{aligned}$$

On page 415, the second term of (57) should read:

$$\frac{-i}{4}\exp(-\nu_1 t).$$

Generalized Resolution in Radar Systems*

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AND J. F. KOVAL†, MEMBER, IRE

Summary—A generalized theory of radar resolution has been developed to facilitate understanding of the fundamental resolution limitations of radar systems. Previous work by Woodward and Elspas[1,2] to determine limitations on radar resolution led to the concept of an ambiguity function which is a quantitative measure of radar resolution in range and range rate. This theory has been extended to include simultaneous resolution in range, range rate, azimuth and elevation and led to the derivation of a four-dimensional ambiguity function. Resolution constants derived from the ambiguity function show clearly the trade-offs between system parameters and resolution.

A new concept, "angular dispersion and compression," has been evolved from the theoretical development. An angular compression system, analogous to a pulse compression system, employs a pseudo-randomly dispersed pattern which is compressed in angle at the receiver with a correlation technique, to produce the effect of a narrow beam without having a physically narrow beam.

The signal bandwidth's effect on the pattern of an antenna and on its angular resolution has been found to be slight, except for very large bandwidths. It has been concluded that trade-offs between signal complexity and antenna complexity have no practical advantage.

Because of the presence of noise, radar measurement of target parameters is essentially equivalent to statistical estimation. Woodward and Elspas have shown that the range, range-rate ambiguity function is the natural quantity to use in making a maximum likelihood estimate of range and range rate. The technique of maximum likelihood estimation has been extended to angular measurement. It is shown that the angular ambiguity function, described above, is the natural quantity to use in obtaining a maximum likelihood estimate of angle, although the angular ambiguity function was derived without considering noise.

GLOSSARY

θ = Angle of a target with respect to broadside in the plane containing the x axis and the line intersecting the phase center of the array and the target

γ = Angle of a target with respect to broadside in the plane containing the y axis and the line intersecting the phase center of the array and the target

$u = (\sin \theta_1)/c$

$u+v = (\sin \theta_2)/c$

$p = (\sin \gamma_1)/c$

$p+q = (\sin \gamma_2)/c$

c = Speed of propagation

$s(t)$ = Transmitted signal

$a(t) = \alpha(t)\exp[-j\phi(t)]$ = Modulating function

$A(\omega)$ = Fourier transform of $a(t)$

$I(x)$, $I(x, y)$ = Illumination functions

$G(u\omega_0)$ = Antenna pattern

$R(\tau)$ = Autocorrelation function

J = Ambiguity function

A_{eff} = Effective area of an aperture

l_{eff} = Effective length of an aperture

τ = Delay difference between two targets

ω_d = Doppler frequency (rad./sec)

ω_0 = Carrier frequency (rad./sec)

β_x = Delay introduced at an element at position x

W_e = Effective video bandwidth (cps)

δ = Fractional bandwidth

$n(t, x_i)$ = Noise at the ith aperture element

$\tilde{n}(t, x_i)$ = Complex envelope of $n(t, x_i)$

$p[n(t, x_i)]$ = Probability density function of $n(t, x_i)$

I. INTRODUCTION

THE RESOLUTION capabilities of a radar system indicate its ability to separate target returns and reject clutter. Thus the resolution of a radar system is an extremely important aspect of its over-all performance and these considerations often dictate system design philosophy.

The pioneer work of Woodward and others[1,2] has led to a quantitative concept of range and range-rate resolution in which the inherent resolution properties of signals are clearly demonstrated. Woodward has shown, for example, that range resolution depends only on signal bandwidth, that velocity resolution depends only on signal time duration and that combined resolution properties in range and range rate can be described in terms of a two-dimensional ambiguity function which expresses the trade-off between resolutions in both quantities. This trade-off is expressed in terms of an ambiguity principle which shows the limitations on simultaneous resolution in range and range rate.

This paper describes the formulation of an ambiguity function which demonstrates the fundamental limitations of a radar system in resolving targets differing in range, azimuth angle, elevation angle and range rate. The implications of the theory are presented.

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† Philco Scientific Laboratory, Blue Bell, Pa.

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II. Radar Ambiguity Functions

A. Criterion of Separability

Resolution refers to the separability of objects or targets which differ in dimensions, characteristics or location. The interest here is in inherent resolution properties, *i.e.*, properties which limit resolution in themselves as distinguished from limitations imposed by fluctuation noise, randomly varying transmission media, and other forms of disturbance.

A first requirement in describing resolution is a criterion of separability. Since separability depends upon differences, a measure of difference between signals is needed. For example, in the case of range resolution of two radar echoes it is necessary to distinguish between an echo delayed by t_0 and an echo of the same form delayed by $t_0+\tau$. The smaller the delay difference, τ, for which the two echoes can be distinguished, the better the range resolution will be.

Giving the smallest value of τ for which echoes can be distinguished is only part of the story; there may exist other values of τ, rather large, for which two echoes may be indistinguishable. These values of τ represent *ambiguities*. Ambiguities of this type occur in the familiar case of a pulsed radar employing a periodic repetition period. What is needed is a measure of difference, as a function of τ, which will display all of the resolution properties. A measure which meets these requirements is the integrated squared difference between the two echoes. This measure has been used by Woodward[1] and Elspas.[2] If the transmitted signal is $s(t)$, the signals received from the two targets, assumed to be stationary and of equal strength, will be $s(t-t_0)$ and $s(t-t_0-\tau)$. The measure of difference is given by

$$\epsilon^2 = \int_{-\infty}^{+\infty} | s(t - t_0) - s(t - t_0 - \tau) |^2 dt. \quad (1)$$

Under the assumption of equal energy returned from the targets,

$$\epsilon^2 = 2 \times \text{received energy} - 2 \operatorname{Re} \int_{-\infty}^{+\infty} s^*(t)s(t - \tau)dt \quad (2)$$

where the asterisk indicates the conjugate and the real part of the integral is to be taken.

The resolution properties are tied up in the expression

$$\operatorname{Re} \int_{-\infty}^{+\infty} s^*(t)s(t - \tau)dt. \quad (3)$$

For good resolution this should be as small as possible for all $\tau \neq 0$. The important features of (3) can be shown much better by writing $s(t)$ so that it shows explicitly its modulation structure. The analytic expression for the signal is[1,2]

$$\begin{aligned} s(t) &= \alpha(t) \exp [j(\omega_0 t - \phi(t))] \\ &= a(t) \exp [j\omega_0 t] \end{aligned} \quad (4)$$

where $a(t) = \alpha(t)\exp[j\phi(t)]$ expresses both the amplitude and phase modulations. Upon substitution of (4) into (3) the resolution properties are expressed by

$$\operatorname{Re} \{\exp [j\omega_0\tau] J_\tau(\tau)\}$$

where

$$J_\tau(\tau) = \int_{-\infty}^{\infty} a^*(t)a(t - \tau)dt. \quad (5)$$

$J_\tau(\tau)$ can be recognized as the complex self-convolution of the signal modulation. It is also called the complex autocorrelation function of $a(t)$, but a better term is the *range* or *delay ambiguity function*. There is no need to explore the properties of $J_\tau(\tau)$, for they have been explored thoroughly.[1,2] Here the concern is merely to lay the background for similar considerations in angular resolution.

Doppler frequency shift, due to radial velocity, can be accommodated in the same formulation. Two targets at the same range causing Doppler shifts f_1 and f_1+f_d, respectively will give rise to echoes having spectra given by $S(f-f_1)$ and $S(f-f_1-f_d)$, respectively. Written as time functions, the signals are $s(t)\exp[-j\omega_1 t]$ and $s(t)\exp[-j(\omega_1+\omega_d)t]$. The measure of difference is

$$\epsilon^2 = \frac{1}{2\pi} \int_{-\infty}^{\infty} | S(\omega - \omega_1) - S(\omega - \omega_1 - \omega_d) |^2 d\omega \quad (6)$$

or

$$\epsilon^2 = \int_{-\infty}^{\infty} | s(t) \exp [-j\omega_1 t] - s(t) \exp [-j(\omega_1 + \omega_d)t] |^2 dt. \quad (7)$$

Eqs. (6) and (7) are exactly the same. Each leads to the Doppler ambiguity function or complex frequency-correlation function given by

$$\begin{aligned} J_{\omega_d}(\omega_d) &= \frac{1}{2\pi} \int_{-\infty}^{\infty} A^*(\omega) A(\omega - \omega_d) d\omega \\ &= \int_{-\infty}^{\infty} | a(t) |^2 \exp [j\omega_d t] dt \end{aligned} \quad (8)$$

where $A(\omega)$ is the Fourier transform of $a(t)$ and $\omega_d = 2\pi f_d$.

A consideration of both delay and Doppler shift leads to the combined delay-Doppler (range, range-rate) ambiguity function

$$\begin{aligned} J_{\tau,\omega_d}(\tau, \omega_d) &= \int_{-\infty}^{\infty} a(t)a^*(t - \tau) \exp [j\omega_d t] dt \\ &= \frac{1}{2\pi} \int_{-\infty}^{\infty} A^*(\omega) A(\omega - \omega_d) \exp [j\omega\tau] d\omega. \end{aligned} \quad (9)$$

Desirable properties for the ambiguity function can be stated and then a synthesis problem is presented: given a desirable range and/or velocity ambiguity function, what signal form will achieve it?

It is not intended that this question be answered here. The reader is referred to work by Lerner,[3,4] Key,[5,6] Fowle,[5,6,7] and Haggarty,[5,6] Wilcox,[8] Sussman,[9] Widnall,[10] and Klauder.[11]

In the discussion to follow, an analogous development for angular resolution is sought. The signal description in this case will be somewhat different from that used in the analysis of range resolution. There the signal is described as a scalar function of a single variable—time. This was true even when Doppler shift and range difference were considered together. When the signals are considered as waves propagated in space, a different sort of description is necessary. To make the discussion definite, attention will be confined to a single plane (the azimuthal plane, for instance), neglecting the polarization involved, assuming that a scalar wave in a plane is being treated and that the sources are so far removed that the wavefronts are straight lines. Furthermore, the discussion will be confined to linear apertures.

B. Formulation of the Angular Ambiguity Function

Before formulating the angular ambiguity function, a brief discussion of the nature of the received signal will be helpful. Consider a receiving line aperture and choose a reference point on the aperture to be labeled $x=0$ (see Fig. 1). Using the subscripts 1 and 2 to indicate signals arriving at the aperture from two targets differing in angle

$$s_1(t) = s\left(t - \frac{x}{c}\sin\theta_1\right) \tag{10}$$

and

$$s_2(t) = s\left(t - \frac{x}{c}\sin\theta_2\right) \tag{11}$$

where $s(t)$ is the transmitted signal, $s_1(t)$ and $s_2(t)$ are the signals received at position x from targets 1 and 2, respectively, and c is the rate of propagation. Let

$$u = \frac{\sin\theta_1}{c} \tag{12}$$

and

$$u + v = \frac{\sin\theta_2}{c}. \tag{13}$$

Then

$$s_1(t) = s(t - ux) \tag{14}$$

and

$$s_2(t) = s(t - ux - vx). \tag{15}$$

It can be seen immediately that the signals at the aperture are functions of position along the aperture as well as functions of time. The two signals have the same structure but are displaced in time by an amount depending upon the difference in angle and distance from the reference point of the aperture. Because each signal for fixed u and v is a function of both t and x, each can be plotted as a surface sitting on the (t, x) plane as illustrated in Fig. 2. The two signals coincide exactly at

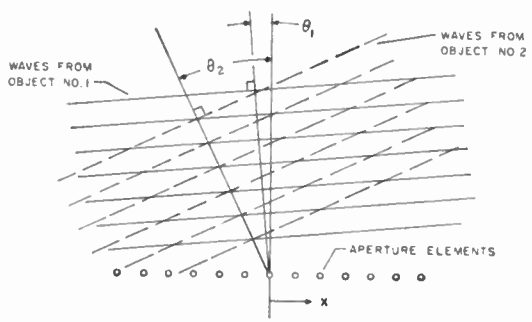


Fig. 1—Waves impinging upon a linear receiving aperture.

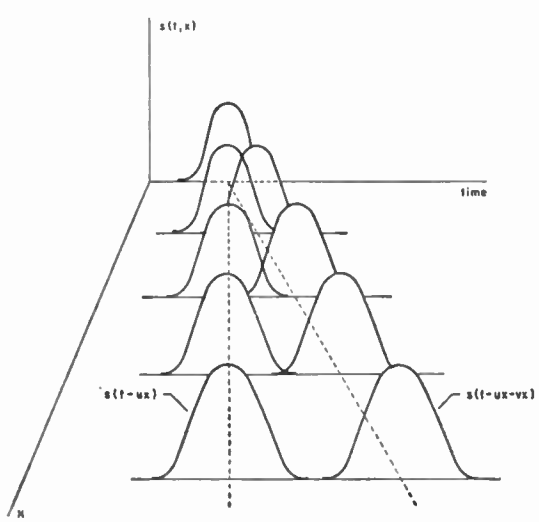


Fig. 2—Representation of two signals arriving at an aperture from different angles.

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$x=0$ and spread farther and farther apart for increasing x. However, each signal is an ensemble of subsignals, one subsignal arising from each part of the aperture. Each subsignal of the ensemble is characterized by a time difference referred to the reference point of the aperture and by the value of the illumination function at the particular value of x. Different targets give rise to different ensembles of subsignals; to determine "inherent" resolution properties of the ensemble, a measure of the "difference" between two ensembles must be defined. This difference will have v as a parameter and, of course, will depend upon the signal structure and illumination function.

The two signal ensembles as received by the aperture are designated by $s_1(t, x)$ and $s_2(t, x)$, and can be written as

$$\begin{aligned} s_1(t, x) &= |I(x)|\, s(t - ux + \beta_x) \\ s_2(t, x) &= |I(x)|\, s(t - ux - vx + \beta_x) \end{aligned} \tag{16}$$

where $|I(x)|$ is the magnitude of $I(x)$, the complex illumination function and β_x is the equivalent time delay associated with $I(x)$. Expressed in analytic form, the ensembles are

$$\begin{aligned} s_1(t, x) &= |I(x)|\, a(t - ux + \beta_x) \exp [j\omega_0(t - ux + \beta_x)] \\ s_2(t, x) &= |I(x)|\, a(t - ux - vx + \beta_x) \\ &\quad \cdot \exp [j\omega_0(t - ux - vx + \beta_x)]. \end{aligned} \tag{17}$$

We note that $\omega_0\beta_x$ is the phase angle associated with $I(x)$ and we assume it to be of little consequence in affecting the modulation. Thus it may be omitted from the argument of $a(t)$. (There is no loss of generality here; β_x would have disappeared from $a(t)$, just as ux does.) Because

$$I(x) = |I(x)| \exp (j\omega_0\beta_x), \tag{18}$$

it follows that

$$\begin{aligned} s_1(t, x) &= I(x)a(t - ux) \exp [j\omega_0(t - ux)] \\ s_2(t, x) &= I(x)a(t - ux - vx) \exp [j\omega_0(t - ux - vx)]. \end{aligned} \tag{19}$$

The measure of difference between the two signal ensembles will be taken as the integrated square difference between them:

$$\begin{aligned} d(s_1, s_2) &= \iint |s_1(t, x) - s_2(t, x)|^2 dxdt \\ &= \iint |s_1(t, x)|^2 dxdt + \iint |s_2(t, x)|^2 dxdt \\ &\quad - 2 \operatorname{Re} \iint s_1(t, x)s_2^*(t, x)dxdt. \end{aligned} \tag{20}$$

The first two terms are energy terms and are essentially independent of u and v. The last term is the one of interest and for $d(s_1, s_2)$ to be large, this must be small.

Define the angular ambiguity function $J_v(v)$ as

$$\begin{aligned} J_v(v) &= \iint s_1(t, x)s_2^*(t, x)dxdt \\ &= \iint |I(x)|^2 a(t - ux)a^*(t - ux - vx) \\ &\quad \cdot \exp (j\omega_0 vx)dxdt. \end{aligned} \tag{21}$$

Noting that

$$\begin{aligned} &\int a(t - ux)a^*(t - ux - vx)dt \\ &\qquad = \frac{1}{2\pi} \int |A(\omega)|^2 e^{j\omega vx} d\omega, \end{aligned} \tag{22}$$

it follows that

$$J_v(v) = \frac{1}{2\pi} \iint |I(x)|^2 |A(\omega)|^2 \exp [j(\omega_0 + \omega)vx]dxd\omega. \tag{23}$$

The space pattern of the aperture, at frequency ω_0, is defined as

$$G(\omega_0 u) = \frac{1}{2\pi} \int I(x) \exp (j\omega_0 ux)dx. \tag{24}$$

It can be shown that

$$\begin{aligned} &\int |I(x)|^2 \exp (j\omega_0 vx)dx \\ &\qquad = 2\pi \int G^*(\omega_0 u)G(\omega_0[u + v])du. \end{aligned} \tag{25}$$

The left hand side of (25) is the self-convolution of the space pattern. Letting

$$R_v(\omega_0 v) = 2\pi \int G^*(\omega_0 u)G(\omega_0[u + v])du, \tag{26}$$

(23) becomes

$$J_v(v) = \int |A(\omega)|^2 R_v[(\omega_0 + \omega)v]d\omega. \tag{27}$$

Also, from (22)

$$J_v(v) = \int |I(x)|^2 J_\tau(vx) \exp (j\omega_0 vx)dx \tag{28}$$

where J_τ is the range-delay ambiguity function.

When the narrow band approximation (approximately 20 per cent or less) can be made, $\omega_0 + \omega \approx \omega_0$ and

$$J_v(v) = J_\tau(0)R_v(\omega_0 v). \tag{29}$$

Eq. (29) gives the angular ambiguity function in its most useful form. Note that the angular ambiguity function, apart from a multiplying constant, is the self-convolution (crudely speaking, the autocorrelation function) of the space pattern.

C. Angular Resolution Constant

The angular resolution constant Δv may be defined as

$$\Delta v = \frac{\int |J_v(v)|^2 dv}{J_v^2(0)} = \frac{\int |R_r(\omega_0 v)|^2 dv}{R_r^2(0)}. \tag{30}$$

Using (25) and (26), this becomes

$$\Delta v = \frac{2\pi}{\omega_0} \frac{\int |I(x)|^4 dx}{\left[\int |I(x)|^2 dx\right]^2}. \tag{31}$$

Defining the effective aperture length l_{eff} by

$$l_{eff} = \frac{\left[\int |I(x)|^2 dx\right]^2}{\int |I(x)|^4 dx}, \tag{32}$$

then

$$\Delta v = \frac{1}{f_0 l_{eff}}. \tag{33}$$

For small angular differences $\Delta v \approx \Delta\theta/c$.

$$\Delta\theta \approx \frac{\lambda_0}{l_{eff}} \equiv \text{effective beamwidth} \tag{34}$$

where λ_0 is the carrier wavelength.

The implications of (34) are important. In a manner analogous to the range case, the actual shape of the space pattern does not determine the angular resolution. It is the self-convolution of the pattern which is important. The angular resolution depends on what we have called the effective length as given by (32). No matter how the phase of the illumination varies across the aperture, these can be unscrambled to display the best resolution. An application of this principle is discussed later.

D. Combined Angle and Delay Resolution

The discussion above is easily extended to handle two targets differing in angle and by a range delay τ. Under these circumstances (18) becomes

$$\begin{aligned} s_1(t, x) &= I(x)a(t - ux)\exp[j\omega_0(t - ux)] \\ s_2(t, x) &= I(x)a(t - \tau - ux - vx) \\ &\quad \cdot \exp[j\omega_0(t - \tau - ux - vx)]. \end{aligned} \tag{35}$$

Then straightforward analogy results in the range-angle ambiguity function

$$J_{\tau,v}(\tau, v) = \frac{1}{2\pi}\int\int |I(x)|^2 |A(\omega)|^2 \cdot \exp[j(\omega_0 + \omega)vx + j\omega\tau] dx d\omega. \tag{36}$$

If the narrow band approximation is used this becomes

$$\begin{aligned} J_{\tau,v}(\tau, v) &= \frac{1}{2\pi}\int\int |I(x)|^2 |A(\omega)|^2 \cdot \exp[j(\omega_0 vx + \omega\tau)] dx d\omega \\ &= J_\tau(\tau) R_r(\omega_0 v) \end{aligned} \tag{37}$$

where $J_\tau(\tau)$ is the range ambiguity function.

A sketch of a range-angle ambiguity function for a $(\sin x/x)^2$ energy density spectrum and a uniformly illuminated linear aperture is shown in Fig. 3. A combined angle-Doppler ambiguity function may be defined in a precisely similar manner. Fig. 4 shows a sketch of such an ambiguity function for a signal whose envelope is proportional to $|\sin x/x|$ and a uniformly illuminated linear array.

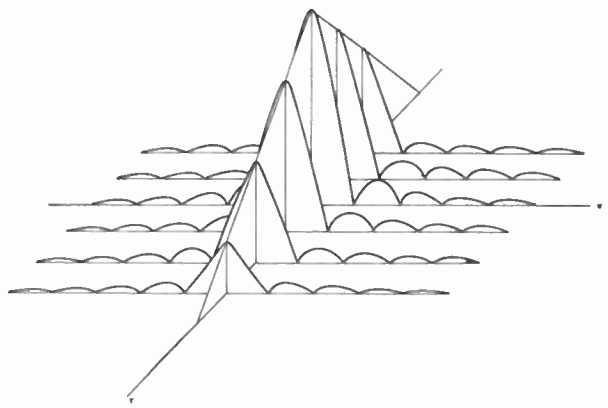


Fig. 3—Ambiguity function in range and angle.

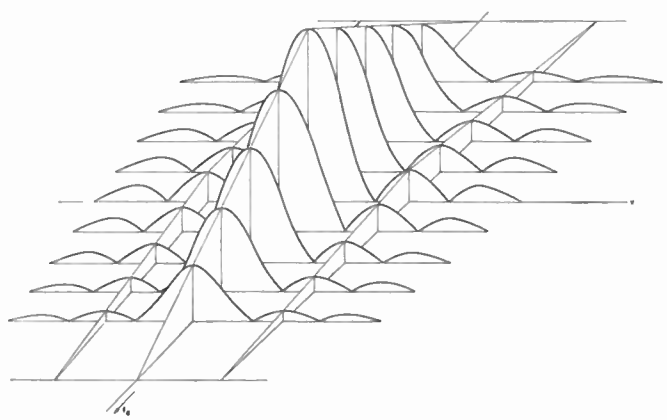


Fig. 4—Ambiguity function in angle and Doppler.

E. Resolution in Azimuth and Elevation

When both azimuth and elevation are considered, (19) becomes

$$s_1(t, x, y) = I(x, y)a(t - ux - py)\exp[j\omega_0(t - ux - py)]$$
$$s_2(t, x, y) = I(x, y)a(t - ux - vx - py) \cdot \exp[j\omega_0(t - ux - vx - py - qy)]. \quad (38)$$

Then the "difference" between these signals is defined by

$$d(s_1, s_2) = \iiint |s_1(t, x, y) - s_2(t, x, y)|^2 dxdydt. \quad (39)$$

The ambiguity function which results is

$$J_{v,q}(v, q) = \frac{1}{2\pi}\iiint |I(x, y)|^2 |A(\omega)|^2 \cdot \exp[j(\omega_0 + \omega)(vx + qy)]dxdyd\omega. \quad (40)$$

The two-dimensional pattern is defined by

$$G(\omega_0 u, \omega_0 p) = \frac{1}{4\pi^2}\iint I(x, y)\exp[j\omega_0(ux + py)]dxdy. \quad (41)$$

Defining the self-convolution of the pattern by

$$R_{v,q}(\omega_0 v, \omega_0 q) = \iint G^*(\omega_0 u, \omega_0 p)G(\omega_0[u + v], \omega_0[p + q])dudp, \quad (42)$$

then

$$J_{v,q}(v, q) = \frac{1}{2\pi}\int |A(\omega)|^2 R_{v,q}[(\omega_0 + \omega)v, (\omega_0 + \omega)q]d\omega. \quad (43)$$

When the narrow-band approximation is used,

$$J_{v,q}(v, q) = J_\tau(0)R_{v,q}(v, q). \quad (44)$$

The resolution constant in azimuth and elevation is defined by

$$\Delta(v, q) = \frac{\iint J_{v,q}(v, q)^2\, dvdq}{J_{v,q}^2(0, 0)} \quad (45)$$

$$= \frac{4\pi^2}{\omega_0^2}\cdot\frac{\iint I(x, y)^4\, dxdy}{\left[\iint I(x, y)^2\, dxdy\right]^2}. \quad (46)$$

Define the effective area of the aperture by

$$A_{\text{eff}} = \frac{\left[\iint I(x, y)^2\, dxdy\right]^2}{\iint (x, y)^4\, dxdy}. \quad (47)$$

Then

$$\Delta(v, q) = \frac{1}{f_0^2 A_{\text{eff}}}, \quad (48)$$

and, if θ is the azimuth angle and ϕ is the elevation angle, for small differences we have

$$\Delta(\theta, \phi) = \frac{\lambda_0^2}{A_{\text{eff}}}. \quad (49)$$

In the special case where the illumination function is separable we have

$$I(x, y) = I(x)I(y) \quad (50)$$

and

$$\Delta(v, q) = \Delta v\Delta q = \frac{1}{f_0^2 l_{\text{eff}} H_{\text{eff}}} \quad (51)$$

where H_{eff} is the effective aperture height. Also

$$\Delta(\theta, \phi) = \Delta\theta\Delta\phi = \frac{\lambda_0^2}{l_{\text{eff}} H_{\text{eff}}}. \quad (52)$$

F. Combined Ambiguity Function

It is fairly clear that a combined ambiguity function describing the combined resolution in range, range rate, and the two angles is obtained easily if the narrow band approximation is made. The combined ambiguity function is given by

$$J(\tau, \omega_d, v, q) = J_{\tau,\omega_d}(\tau, \omega_d)R_{v,q}(\omega_0 v, \omega_0 q) \quad (53)$$

where $J_{\tau,\omega_d}(\tau, \omega_d)$ is given by (9) and $R_{v,q}$ is given by (42). The combined resolution constant is

$$\Delta(\tau, \omega_d, v, q) = \Delta(\tau, \omega_d)\Delta(v, q) = \Delta(v, q)2\pi = \frac{2\pi}{f_0^2 A_{\text{eff}}}. \quad (54)$$

G. Resolution Constants

The resolution constants for resolution in one, two, three or four dimensions are tabulated in Table I. Both the general expressions and the narrow-band approximations are shown. The general expressions for the resolution constants can be derived in a straightforward manner (see Appendix).

Examining first the narrow-band approximation to the resolution constants, observe that for all cases that involve range and range rate (Cases 4, 6 and 8) there is no dependence upon the bandwidth of the transmitted signal (*i.e.*, up to fractional bandwidths of about 0.2). For cases involving range alone or range combined with one or both angles, the resolution constant is inversely proportional to the bandwidth.

To see the influence of signal bandwidth on the resolution constants more clearly, refer to Figs. 5–8. These four figures show plots of the resolution constants for Cases 5 to 8, inclusive, as functions of fractional band-

TABLE I

RESOLUTION CONSTANTS

	Narrow Band Approx. $\delta < 0.2$	General
1. $\Delta(\tau)$	$\frac{1}{2W_e}$	$\frac{1}{2W_e}$
2. $\Delta(\omega_d)$	$\frac{2\pi}{T}$	$\frac{2\pi}{T}$
3. $\Delta(v)$	$\frac{\lambda_0}{cl_{eff}}$	$\frac{\lambda_0}{cl_{eff}}$
4. $\Delta(\tau, \omega_d)$	2π	2π
5. $\Delta(\tau, v)$	$\frac{\lambda_0}{2cl_{eff}W}$	$\frac{4\pi^2}{l_{eff}} \frac{\int \frac{\lvert A(\omega)\rvert^4}{\lvert \omega + \omega_d \rvert} d\omega}{\left[\int \lvert A(\omega)\rvert^2 d\omega\right]^2}$

	Narrow Band Approx. $\delta < 0.2$	General
6. $\Delta(\tau, \omega_d, v)$	$\frac{2\pi\lambda_0}{cl_{eff}}$	$\frac{4\pi^2}{l_{eff}} \frac{\int \frac{\lvert A(\omega)\rvert^2 \lvert A(\omega-\omega_d)\rvert^2}{\lvert \omega + \omega_0 - \omega_d \rvert} d\omega d\omega_d}{\left[\int \lvert A(\omega)\rvert^2 d\omega\right]^2}$
7. $\Delta(\tau, v, q)$	$\frac{\lambda_0^2}{2c^2 A_{eff} W_e}$	$\frac{8\pi^3}{A_{eff}} \frac{\int \frac{\lvert A(\omega)\rvert^4}{\lvert \omega + \omega_0 \rvert^2} d\omega}{\left[\int \lvert A(\omega)\rvert^2 d\omega\right]^2}$
8. $\Delta(\tau, \omega_d, v, q)$	$\frac{2\pi\lambda_0^2}{c^2 A_{eff}}$	$\frac{8\pi^3}{A_{eff}} \frac{\iint \frac{\lvert A(\omega)\rvert^2 \lvert A(\omega-\omega_d)\rvert^2}{\lvert \omega + \omega_0 - \omega_d \rvert^2} d\omega d\omega_d}{\left[\int \lvert A(\omega)\rvert^2 d\omega\right]^2}$

W_e = Effective Video Bandwidth
T = Effective Signal Duration
l_{eff} = Effective Aperture Length
A_{eff} = Effective Aperture Area

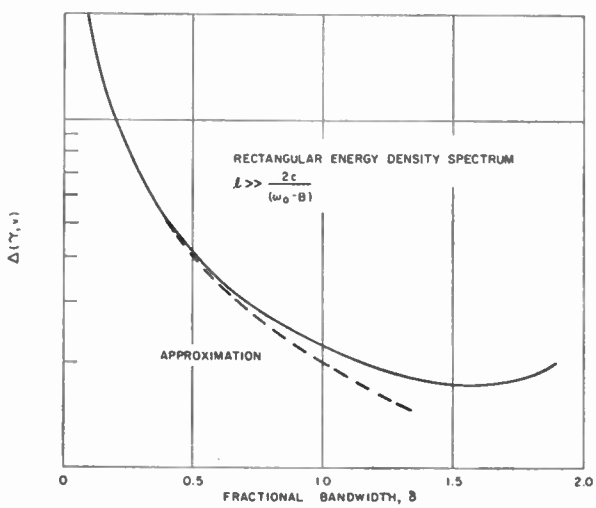


Fig. 5—Resolution constant or ambiguity area for range and angle.

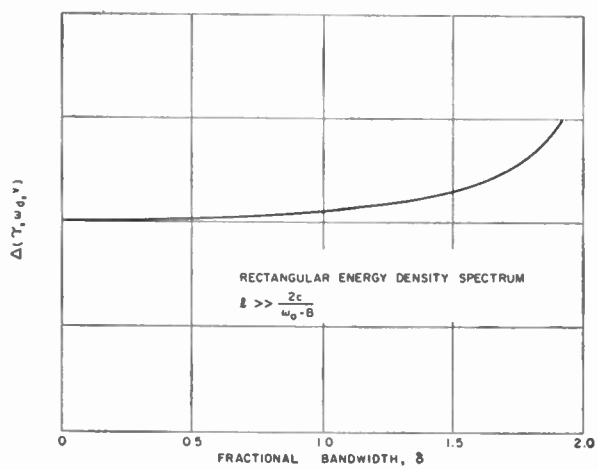


Fig. 6—Resolution constant or ambiguity area for range, range rate and angle.

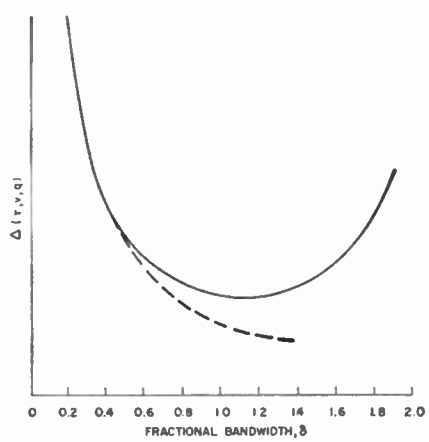


Fig. 7—Resolution constant for range, azimuth and elevation angles.

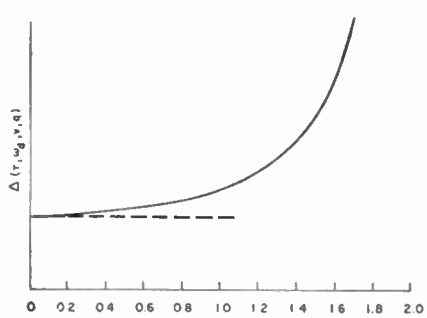


Fig. 8—Resolution constant for range, range rate, azimuth and elevation angles.

width, δ. A rectangular energy density spectrum has been assumed. The approximate expressions are plotted as dashed curves. Note that the curves do not depart from the approximation for $\delta<0.2-0.3$. For simultaneous resolution in range and angle,there is a lower limit to the resolution constant occurring at $\delta=1.59$. A lower limit also is found for simultaneous resolution in range and the two angles at $\delta=1.15$. A larger dependence upon bandwidth is observed for the resolution constants which involve both angles as compared with those involving only one angle.

In general it is concluded that the trade-off between antenna complexity and signal complexity is a poor one. Although the sidelobes of an antenna pattern can be reduced by increasing the bandwidth,[12,13] it is seen that when considering simultaneous resolution in range, range rate and angle, the resolution constant increases with bandwidth. At best, the sidelobe reduction is small except for very large bandwidths ($\delta>0.5$) and the increased bandwidth has a degrading effect on the resolution constants.

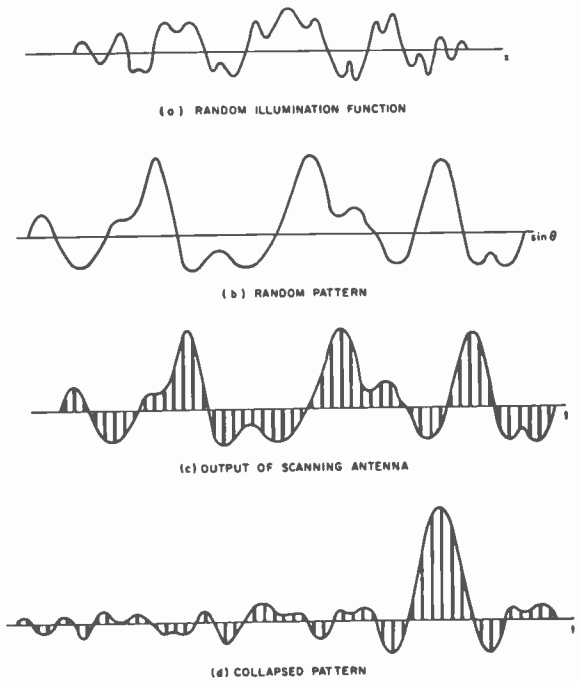


Fig. 9—Angular dispersion and compression.

III. Angular Dispersion and Compression

The angular dispersion and compression technique is a spatial analog of pulse dispersion and compression techniques. If an aperture is excited with a random amplitude and phase illumination with a zero average, its far-field pattern has a random appearance without indication of a main beam (see Fig. 9). Consider the return on a single receiving element from a target which has been illuminated by this antenna. The amplitude and phase of the output depends upon the angle of the target. If the array is allowed to rotate electronically or mechanically, the waveform observed at the receiving element output will be the same as the random antenna pattern. In the case of pulsed radar the output is a sampled replica of the pattern. A filter matched to this random pattern will form a correlation peak at its output; its time position is proportional to the angle of the target. When two targets separated in angle exist, the output will consist of two correlation peaks. The time separation between the correlation peaks is proportional to the angular separation of the two targets; thus targets have been resolved in angle without using a strong directive beam.

To illustrate this concept further, refer to Fig. 10. Part (a) of this figure represents a space pattern which is dispersed in both time and angle. No main beam exists in either dimension. After range compression, the space pattern is modified as shown on part (b) of the figure. Resolution is now evident in range but not in

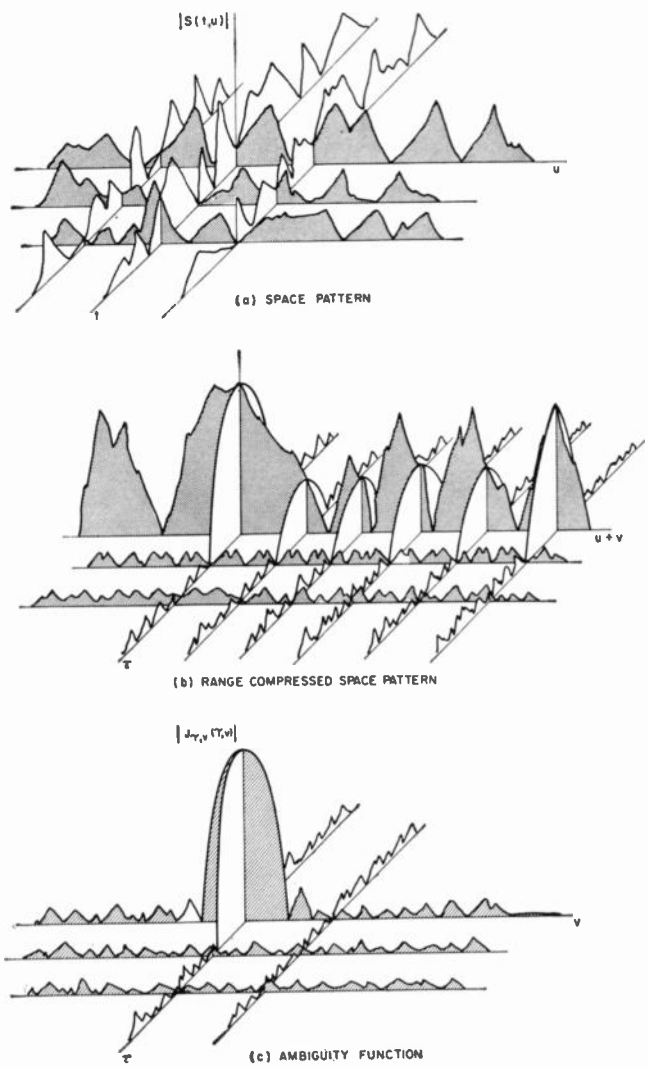
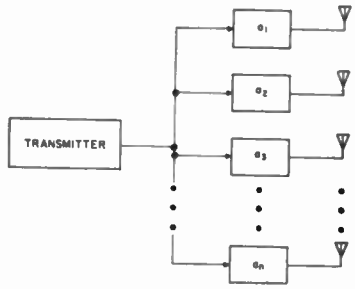


Fig. 10—Dispersion in angle.

[12] L. R. Daussin, K. E. Niebuhr, N. J. Nilsson, "The Effects of Wide-Band Signals on Radar Antenna Design," Rome Air Dev. Ctr., Griffiss AFB, N. Y., Rept. No. RADC-TN-59-178, ASTIA Doc. No. AD 216023; July, 1959.

[13] C. Hauer, J. Koval, L. Procopio and H. Urkowitz, "Synthetic Aperture Study," Philco Corp. Res. Div., Final Rept., Contract No. AF 30(602)-2324 (RADC); May, 1961.



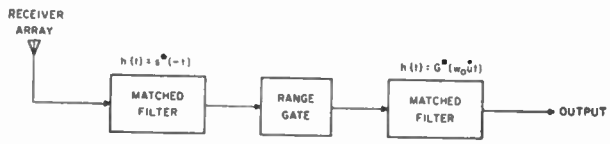


Fig. 11—Angular compression technique.

angle. After the signal is compressed in angle, a sharp main beam is formed (part (c)). This surface represents the range-angle ambiguity function.

A functional block diagram for angular dispersion and compression is shown in Fig. 11. The random illumination function is generated by random adjustment of the coefficients $a_1, a_2, \cdots, a_n$. Antenna rotation is necessary to convert the pattern into a time function at the receiver. The received signal is first fed to a filter matched to transmitted signal. The output of this filter is given by

$$e(t) = R_\tau(t - \tau)G(\omega_0 \dot{u}t - v) \tag{55}$$

where $R_\tau(\tau)$ is the autocorrelation function of the transmitted signal and $\dot{u}$ is the rate of change of u.

The range gate following the filter selects the range element of interest, or if range is unknown, an estimate of range must be made first. The range gate output is

$$e(t) = R_\tau(0)G(\omega_0 \dot{u}t - v). \tag{56}$$

The second matched filter has an impulse response given by

$$h(t) = G^*[-\omega_0 \dot{u}t]. \tag{57}$$

The output of the filter in response to the signal of (56) is

$$e_0 = \int R_\tau(0)G[\omega_0(\dot{u}t_1 - v)]G^*[\omega_0 \dot{u}t_1 - t)]dt_1 \tag{58}$$

or

$$e_0 = R_\tau(0) \iiint I(x)I^*(y) \exp j\omega_0(\dot{u}t_1 - v)x \cdot \exp(-j\omega_0 \dot{u}(t_1 - t)y)dydxdt_1. \tag{59}$$

Examining all terms involving t_1, one gets

$$\int \exp[j\omega_0 \dot{u}(x - y)t_1]dt_1 = \frac{2\pi}{\omega_0 \dot{u}} \delta(x - y). \tag{60}$$

Therefore

$$e_0 = KR_\tau(0) \int |I(x)|^2 \exp[j\omega_0(\dot{u}t - v)x]dx \tag{61}$$

where K is a constant. e_0 may be written as

$$e_0 = KR_\tau(0)R_v[\omega_0(\dot{u}t - v)] \tag{62}$$

which is precisely the angular ambiguity function.

IV. Estimation of the Target Parameters

A. Introduction

Up to this point the inherent resolution properties of the aperture of a radar system have been described in terms of an ambiguity function which shows not only the resolution properties in each of the coordinates (range, range rate, azimuth, and elevation), but also the relationships among these properties. The formulation suggests that the angular accuracy is related to the angular ambiguity function in a way similar to the way that range accuracy is related to the range ambiguity function of Woodward.[1]

The purpose here is to examine the fundamental limitations of angular accuracy obtainable with an antenna aperture. Ideally one must make as few assumptions as possible about the signal processing which is done at the receiver. Swerling[14] and Bernstein[15] have studied angular accuracy, but their work is based on the assumption that the antenna sums all the signals from each of its elements and that the resultant sum has been demodulated by envelope detection. Needless to say, one cannot be sure that the inherent accuracy has been adequately described.

A more fundamental analysis was made by Manasse[16] who considered the actual wavefront at an aperture. He recognized that the inherent accuracy depends upon the actual wave function across the aperture. By considering the waveform sampled at points along the aperture and sampled in time, Manasse obtained a probability distribution (assuming Gaussian noise) from which a maximum likelihood estimate could be formed. Manasse's paper is limited to a single-frequency pulse signal and neglects the effect of any weighting at the aperture (illumination function).

A similar analysis was made by Hiller[17] for sinusoidal received signals as well as estimating the angle of arrival of a noise wave. Again it was assumed that the illumination function was uniform.

[14] P. Swerling, "Maximum angular accuracy of a pulsed search radar," Proc. IRE, vol. 44, pp. 1146–1155; September, 1956.
[15] R. Bernstein, "An Analysis of Angular Accuracy in Search Radar," IRE Nat'l. Convention Record, pt. 5, pp. 61–78; 1955.
[16] R. Manasse, "An Analysis of Angular Accuracies from Radar Measurements," M.I.T. Lincoln Lab., Lexington, Mass., Group Rept. 32–46; December 6, 1955.
[17] E. R. Hiller, Jr., "Available Source Information in Receiving Apertures," E. E. thesis, Dept. of Elec. Engrg., M.I.T., Cambridge, Mass., 1956; also Res. Lab. of Electronics, M.I.T., Tech. Rept. No. 3, Contract No. NOrd 14362, September 1, 1956.

One may argue that the receiver illumination function should not be involved in determining the *inherent* accuracy obtainable. This work is based on the belief that the measurement is performed by an aperture. The wavefront is never observed directly. Only the portion of the wavefront which is intercepted by the aperture weighted by the illumination function is seen. The result, therefore, is a function of the illumination.

The close connection between accuracy and resolution is recognized in this paper. We have said that resolution measures the separability of target returns. Accuracy refers to the precision of measurement, *i.e.*, how close an estimate of a parameter is to the actual value. An outline is given of a quantitative relationship between angular resolution and accuracy of angular measurement. This relationship is similar to the work of Woodward[1] and Elspas[2] who treat range (and Doppler) resolution as a quantitative measure of difference between target returns; they show that the best statistical estimate of range (and Doppler) is obtained by operating on the function which displays the inherent range (and Doppler) resolution properties of the received waveform, where resolution is defined independently of accuracy.

B. Description of the Received Signal Ensemble

Consider a linear aperture with a plane wave impinging on it as in Fig. 12. If the signal at the center of the aperture is $s(t)$, then, at a point x, the signal will be $s(t-ux)$. Of course, $s(t)$ is never observed directly; it is always weighted by the illumination function of the aperture.

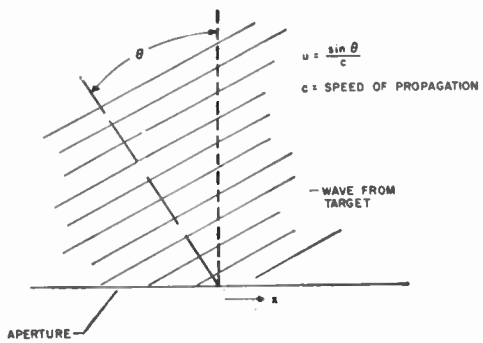


Fig. 12—Linear receiving aperture.

The signal form will be assumed to be known and it will be convenient to express it in analytic (complex) form

$$s(t) = a(t) \exp (j\omega_0 t) \tag{63}$$

where $a(t)$ is the complex modulation (or complex envelope) function. We further assume that the range delay to and from the target is known so that the signal into the aperture, expressed as a time-space function is

$$s(t, x) = | I(x) | a(t - ux + \beta_x) \exp [j\omega_0(t - ux + \beta_x) + j\phi] \tag{64}$$

where ϕ is a phase angle which is unknown in general and may sometimes be interpreted as the "fine" structure of the range delay. Now β_x may be omitted from the argument of the modulation function as being of no consequence in affecting the modulation. Thus

$$s(t, x) = I(x)a(t - ux) \exp [j\omega_0(t - ux) + j\phi]. \tag{65}$$

C. Noise

The two general sources of noise are external (*e.g.*, sky noise) and internal. We consider only internal noise although external noise can be treated similarly.[17] Each aperture element can be considered to be a source of noise which is Gaussian with zero mean. The power density spectrum is equal to a constant N_0 over a band which vanishes outside the frequency interval $|f-f_0| < W/2$, where $f_0=\omega_0/(2\pi)$. The noise may be expressed in analytic form as

$$n(t, x_i) = \tilde{n}(t, x_i)e^{j\omega_0 t} \tag{66}$$

where x_i represents the ith aperture element and the tilde indicates the complex envelope of $n(t, x_i)$. The probability density function of a sample of noise T seconds in duration is[1]

$$p[n(t, x_i)] = k \exp \left\{ - \frac{1}{2N_0} \int_0^T | n(t, x_i) |^2 dt \right\} \tag{67}$$

where k is a proportionality constant.

Consider now the collection of noise waveforms from all of the aperture elements. If there are M such elements we assume that the M noise waveforms are statistically independent with the same power density spectra. Then the probability density function for the set is

$$p[n(t, x)] = k \exp \left\{ - \frac{1}{2N_0} \sum_{i=1}^{M} \int_0^T | n(t, x_i) |^2 dt \right\}. \tag{68}$$

All proportionality constants have been absorbed by k. If M is large or if the aperture is continuous, the summation may be approximated by an integral

$$p[n(t, x)] = k \exp \left\{ - \frac{1}{2N_0} \int_0^T \int_{-L/2}^{L/2} | n(t, x) |^2 dx dt \right\} \tag{69}$$

where L is the length of the aperture.

D. The Maximum Likelihood Estimate

In the presence of noise the received wave is

$$y(t, x) = n(t, x) + s(t, x) \tag{70}$$

$$= \tilde{y}(t, x) \exp [j\omega_0(t - ux)] \tag{71}$$

where the tilde again indicates the complex envelope. Then

$$n(t, x) = y(t, x) - s(t, x) \tag{72}$$

and when this is substituted into (69) the likelihood function of $y(t, x)$ is obtained:

$$p[y(t, x)]$$

$$= k \exp \left\{ - \frac{1}{2N_0} \int_0^T \int_{-L/2}^{L/2} | y(t, x) - s(t, x) |^2 dx dt \right\}, \tag{73}$$

expanding the integrand results in three terms. The two extreme terms will represent the total energy in the received wave and in the signal waveform, both of which are fixed. Then the likelihood function may be written as

$$p[y(t, x)]$$

$$= k \exp \left\{ - \frac{1}{N_0} Re \int_0^T \int_{-L/2}^{L/2} y(t, x) s^*(t, x) dx dt \right\} \tag{74}$$

where k has absorbed all the proportionality constants and "Re" indicates the real part.

If u_0 is the true "angle" of arrival, the maximum likelihood estimate of u_0 is obtained by finding the value of u which maximizes

$$\text{Re } Q = \text{Re} \int_0^T \int_{-L/2}^{L/2} y(t, x) s^*(t, x) dx dt. \tag{75}$$

$y(t, x)$ can be written explicitly as a function of the true value u_0:

$$y(t, x; u_0) = \tilde{y}(t, x; u_0) \exp [j\omega_0(t - u_0 x)]. \tag{76}$$

Substituting (65) and (69) in (75)

$$\text{Re } Q = \text{Re} \left\{ e^{j\phi} \int\int I^*(x) a^*(t - ux) \right.$$

$$\left. \cdot \exp [j\omega_0(u - u_0)x] \tilde{y}(t, x; u_0) dx dt \right\}. \tag{77}$$

Now for complex z,

$$\text{Re} \{ z e^{j\phi} \} = \text{Re} \{ |z| \exp j[\arg z + \phi] \}$$

attains its maximum when $\phi = -\arg z$ and the maximum is $|z|$. Therefore

$$|Q| = \left| \int\int I^*(x) a^*(t - ux) \right.$$

$$\left. \cdot \exp [j\omega_0(u - u_0)x] \tilde{y}(t, x; u_0) dx dt \right| \tag{78}$$

The construction of $|Q|$ may be accomplished in the receiver by performing the operations indicated in (78) or in (75). The resulting receiver structure may be in the form of correlators or matched filters. Fig. 13 shows the matched filter receiver structure for a discrete aperture. The impulse response of the matched filter is just $s^*(-t)$. The delays depend upon the particular value of u under examination. To get the output $|Q(u)|$ as a function of u so that the maximum value may be determined, the delays must be changed. This is the same as electronic scanning. Alternatively, the aperture may be mechanically scanned so that if all the delays are zero the maximum value of $|Q|$ occurs when $u = 0$. A range gate should be included in the figure because the range to the target is known and only that section of the input containing a target return should be processed.

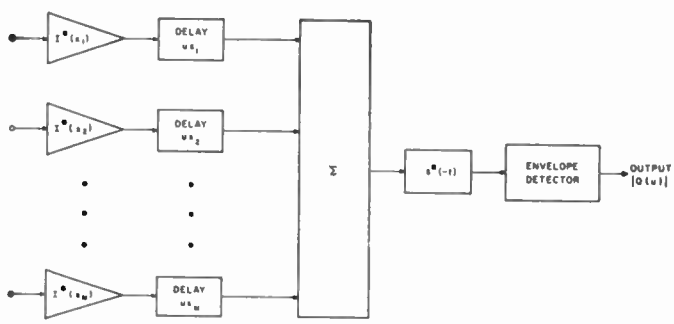


Fig. 13—Angle estimating filter using a matched filter.

E. Relation to the Ambiguity Function

$\tilde{y}(t, x; u_0)$ can be written as

$$\tilde{y}(t, x; u_0) = I(x) a(t - u_0 x) + \tilde{n}(t, x). \tag{79}$$

It is implied, in writing (79), that the statistical characteristics are unaffected by u_0 or any phase shift. Substitution into (78) yields

$$Q(u) = \int\int |I(x)|^2 a(t - u_0 x) a^*(t - ux)$$

$$\cdot \exp [j\omega_0(u - u_0)x] dx dt + \tilde{n}'(u) \tag{80}$$

where $\tilde{n}'(u)$ is a random variable obtained by the operation upon noise. The first term in (80) will be recognized as the angular ambiguity function $J(u)$. That is,

$$Q(u) = J(u - u_0) + \tilde{n}'(u). \tag{81}$$

Thus it is concluded that the maximum likelihood estimation of angle involves the construction (or attempted construction) of the angular ambiguity function and finding the value of u which maximizes the argument.

F. Remarks

It has been shown how the angular ambiguity function fits naturally into the framework of maximum likelihood estimates. It is not difficult to accept the thesis that the multidimensional ambiguity function fits, in the same way, into the framework of estimating all radar parameters simultaneously. Extensions to the other radar parameters and the derivation of the accuracies to be expected in estimating these parameters will be presented in a future paper.

V. Summary and Conclusions

It has been shown that Woodward's philosophy of resolution can be extended to include the radar coordinates of azimuth and elevation as well as those of range and Doppler. In particular the concept of an angular ambiguity function has been obtained by analogy with the range ambiguity function by defining a physically meaningful and mathematically tractable definition of a difference function between the two sets of signals produced at the elements of a receiving aperture by two targets differing only in angle. The resulting angular ambiguity function shows that the inherent angular resolution properties of an aperture are not dependent upon the specific beam pattern *per se*, but upon the space correlation function of this beam pattern. The sharpness of this ambiguity function is determined by the squared magnitude of the complex illumination function as well as, but to a much lesser extent, the energy density spectrum of the signal. In the narrow-band case, the angular resolution is independent of the signal spectrum.

By defining an effective aperture length in terms of the squared magnitude of the complex illumination function, the angular resolution constant is given simply by a constant times the reciprocal of the effective aperture length.

A combined range-angle ambiguity function is defined. It is separable, for narrow-band signals, into the product of the range ambiguity function and angular ambiguity function. The combined range-angle resolution constant is proportional to the reciprocal product of effective bandwidth and effective aperture length.

A two-dimensional angular ambiguity function covering azimuth and elevation simultaneously is defined in an analogous manner to the one-dimensional ambiguity function. The effective area now determines the simultaneous two-dimensional angular resolution constant and is defined in terms of the squared magnitude of the two-dimensional complex illumination function. If the two-dimensional illumination function is separable, the combined azimuth-elevation ambiguity function becomes the product of the individual ambiguity functions.

A four-dimensional ambiguity function covering range, Doppler, azimuth, and elevation is defined by straightforward extension. Again, for narrow-band signals this four-dimensional ambiguity function is separable into the product of the range-Doppler ambiguity function and the azimuth-elevation ambiguity function. The four-dimensional resolution constant shows that the simultaneous resolution in all four coordinates depends upon up the effective area of the aperture. This is principally because the combined range-Doppler resolution constant is independent of the signal complexity.

More exact formulas, allowing for very broad-band signals, are derived for combined ambiguity functions. In general, the bandwidth does affect the resolution constant, but the effect is small. Improved angular resolution is obtained with broad-band signals, but the improvement is slight for increasing bandwidth, showing a poor trade-off between signal complexity and antenna complexity.

The nature of the angular ambiguity function suggests that good angular resolution may be obtained even with a dispersed beam pattern showing no pronounced main beam. The resolution is obtained by collapsing or compressing the dispersed beam pattern in a manner analogous to range or time dispersion and compression. The technique involves a combination of range compression and angle compression using the equivalent of a range correlator or matched filter followed by an angle correlator or matched filter.

Finally the angular measurement is treated from the point of view of statistical estimation. In particular, a maximum likelihood estimate of a target's angle is made in the presence of noise where the other radar coordinates are presumed known. The receiver required to form the maximum likelihood estimate is identical with the one which would be used to form the angular ambiguity function. In other words, the natural quantity which is sought in maximum likelihood estimation of angle is the angular ambiguity function. The angle at which the peak occurs is the angle estimate.

Appendix

Resolution Constant for Range and Angle

The ambiguity function for range and angle is given by

$$J_{\tau,v}(\tau, v) = \frac{1}{2\pi}\int\int |I(x)|^2 |A(\omega)|^2 \cdot \exp\left[j(\omega_0 + \omega)vx + j\omega\tau\right]dx d\omega. \quad (82)$$

The resolution constant for range and angle is defined as

$$\Delta(v, \tau) = \frac{\int |J_{\tau,v}(\tau, v)|^2 dv d\tau}{|J_{\tau,v}(0, 0)|^2}. \quad (83)$$

Taking first the squared magnitude of the ambiguity function,

$$|J_{\tau,v}(\tau, v)|^2 = \frac{1}{4\pi^2}\int\int\int\int |I(x)|^2|I(y)|^2|A(\omega_1)|^2|A(\omega_2)|^2 \cdot \exp[j(\omega_1-\omega_2)\tau - j(\omega_2 y - \omega_1 x)v + j\omega_0 v(x-y)]d\omega_1 d\omega_2 dx dy, \quad (84)$$

the unnormalized area under the ambiguity function [(numerator of (83)] is

$$A' = \int\int |J_{\tau,v}(\tau, v)|^2 d\tau dv. \quad (85)$$

The only factor involving τ is

$$\int \exp[j(\omega_1-\omega_2)\tau]d\tau = 2\pi\delta(\omega_2-\omega_1). \quad (86)$$

Therefore ω_1 has value only at ω_2. Hence

$$A' = \frac{1}{2\pi}\int\int\int\int |I(x)|^2|I(y)|^2|A(\omega)|^4 \cdot \exp[j(\omega+\omega_0)v(x-y)]d\omega dx dy dv. \quad (87)$$

Now

$$\int_{-1/c}^{+1/c}\exp[j(\omega+\omega_0)v(x-y)]dv = \frac{2\sin(\omega_0+\omega)(x-y)/c}{(\omega_0+\omega)(x-y)} = \frac{2}{(\omega_0+\omega)}\frac{\sin 2\pi\dfrac{(x-y)}{\lambda}}{(x-y)} \quad (88)$$

where $\lambda = 2\pi c/(\omega+\omega_0)$.

If the ratio l/λ is reasonably large, x can be considered to have value only at y; *i.e.*, the $(\sin x)/x$ term approaches a delta function:

$$\frac{2\sin(\omega+\omega_0)(x-y)/c}{(x-y)} \approx 2\pi\delta(x-y). \quad (89)$$

Therefore

$$A' = \int\int |I(x)|^4 \frac{|A(\omega)|^4}{|\omega+\omega_0|}dx d\omega. \quad (90)$$

The squared magnitude of the ambiguity function for $\tau=0$, $v=0$ is

$$|J_{\tau,v}(0,0)|^2 = \frac{1}{4\pi^2}\left[\int\int |I(x)|^2|A(\omega)|^2 d\omega dx\right]^2 \quad (91)$$

The normalized ambiguity area or resolution constant is

$$\Delta(\tau, v) = 4\pi^2 \frac{\int |I(x)|^4 dx}{\left[\int |I(x)|^2 dx\right]^2}\frac{\int \dfrac{|A(\omega)|^4}{|\omega+\omega_0|}d\omega}{\left[\int A|(\omega)|^2 d\omega\right]^2} \quad (92)$$

or

$$\Delta(\tau, v) = \frac{4\pi^2}{l_{\text{eff}}}\frac{\int \dfrac{|A(\omega)|^4}{|\omega+\omega_0|}d\omega}{\left[\int |A(\omega)|^2 d\omega\right]^2} \quad (93)$$

where l_{eff} is the effective length of the aperture as defined by (32) in the body of this paper.

For small bandwidth ($\delta < 0.2$), the resolution constant becomes

$$\Delta(\tau, v) \approx \frac{4\pi^2}{l_{\text{eff}}\omega_0}\frac{\int |A(\omega)|^4 d\omega}{\left[\int |A(\omega)|^2 d\omega\right]^2} \quad (94)$$

or

$$\Delta(\tau, v) = \frac{2\pi}{l_{\text{eff}}\omega_0 W_e} = \frac{\lambda_0}{2cl_{\text{eff}}W_e} \quad (95)$$

where W_e is the effective bandwidth of the energy density spectrum in cycles per second and is defined by

$$W_e = \frac{\left[\int |A(\omega)|^2 d\omega\right]^2}{2\pi\int |A(\omega)|^4 d\omega}. \quad (96)$$

Consider the case of a rectangular energy density spectrum. Eq. (93) becomes

$$\Delta(\tau, v) = \frac{4\pi^2}{l_{\text{eff}}}\frac{\int_{\omega_0-B}^{\omega_0+B}\frac{1}{\omega}d\omega}{\int_{-B}^{+B}d\omega} = \frac{4\pi^2}{l_{\text{eff}}\omega_0^2\delta^2}\ln\frac{1+\delta/2}{1-\delta/2} \quad (97)$$

or

$$\Delta(\tau, v) = \frac{\lambda_0}{l_{\text{eff}}\delta c W_e}\tan^{-1}\delta/2 \quad (98)$$

where B is the video bandwidth and $\delta = 2B/\omega_0$.

For small fractional bandwidths (98) reduces to

$$\Delta(\tau, v) = \frac{\lambda_0}{2cl_{\text{eff}}W_e} \quad (99)$$

which agrees with the previous result (95).

The general expressions for the resolution constants for resolution involving other radar parameters can be derived in a similar manner.

Correspondence

Piezo-Junctions: Elements of a New Class of Semiconductor Devices*

Recently large effects of anisotropic stress on the resistance of *p-n* junctions were observed.[1] The magnitude of these effects is far larger than that previously reported.[2,3] Moreover, it was found that stress can cause considerable, and hitherto unreported, changes in the breakdown characteristics. The basic phenomena are illustrated in Fig. 1 where (a) the schematic stressing arrangement and (b) a family of isobaric *I-V* characteristics are shown. The radius *R* of the stressing element ordinarily used was of the order of 20 μ. The contact diameters *d* were of the order of 2 to 3 μ and local pressures obtained under reversible conditions were calculated[4] to be as high as 8×10^4 kg/cm^2. Due to the anisotropies of the stress field and the electrical field various mechanisms, which are presently being investigated, contribute to the effects described. Within the elastic range the effects are completely reversible and provide a basic element, for brevity referred to as the piezo-junction, of a new class of devices.

p-n junctions constitute basic building blocks of many semiconductor devices from diodes to tetrodes. In any such device the control of one junction will affect the performance of the whole device. For example, by stressing the emitter junction of a transistor the output across the collector junction can be controlled. Such an application is illustrated in Fig. 2 which shows a set of collector characteristics of a *p-n-p* Ge mesa transistor. Except for the indicated variations in stress on the emitter, all other experimental conditions remained identical and the changes of h_{fe} from about 40 to 5 were completely reversible. By proper selection of junction depth and the value of *R* of such a piezo-triode responses by more than an order of magnitude larger than those shown have been obtained. An interesting aspect of the drastic changes in transistor characteristics under controlled stress is the clear demonstration of the previously suspected[5] effects which uncontrolled built-in stresses may have on semiconductor devices.

Ge and Si piezo-diodes have been incorporated in small experimental microphone and hydrophone structures. Fig. 3 shows, for example, such a piezo-diode microphone

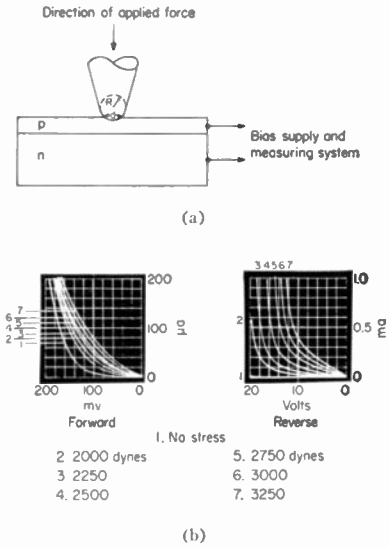


Fig. 1—(a) Schematic stressing arrangement. (b) Forward and reverse bias characteristics of a piezo-diode.

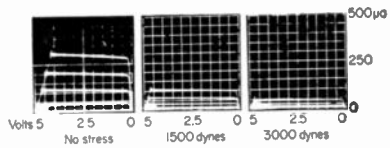


Fig. 2—h_{fe} of a piezo-triode with $I_e = 2$ μa/step.

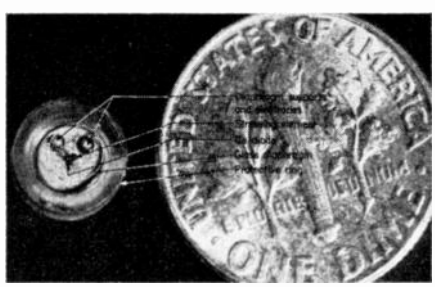


Fig. 3—An experimental piezo-diode microphone.

with the pertinent structural details. Outputs of up to about 2 v were obtained on some units in sound fields of about 60 db above 200 μ dynes. Preliminary measurements indicate that high sensitivities are possible with piezo-triodes since in the latter the inherent gain of the transistor is exploited. The frequency response has been found to extend from dc to over 100 kc, with various peaks which can be attributed to mechanical resonances.

The devices already realized in practice constitute only a fraction of the potential piezo-junction accelerometers, displacement transducers and other piezo-junction devices. In this connection some features of piezo-junctions are of particular interest. For example, the volume of the active semiconductor region extends over only a few cubic microns. Thus small dimensions and small masses and correspondingly high mechanical resonance frequencies are possible. The stress-dependent breakdown characteristics provide potential sensitivities limited mainly by noise and stability considerations.

The high sensitivity displayed suggests the possibility of coupling piezo-junctions with electromechanical transducers to provide novel four-terminal devices. This can, for example, be achieved by applying the signal input to electrostatic, magnetostrictive or piezoelectric elements in suitable mechanical contact with piezo-junctions which provide the output terminals.

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WWV and WWVH Standard Frequency and Time Transmissions*

The frequencies of the National Bureau of Standards radio stations WWV and WWVH are kept in agreement with respect to each other and have been maintained as constant as possible since December 1, 1957, with respect to an improved United States Frequency Standard (USFS).[1] The corrections reported here were arrived at by means of improved measurement methods based on transmissions from the NBS stations WWVB (60 kc) and WWVL (20 kc). The values given in the table are 5-day running averages of the daily 24-hour values for the period beginning at 1800 UT of each day listed.

The time signals of WWV and WWVH are also kept in agreement with each other. Since these signals are locked to the frequency of the transmissions, a continuous departure from UT2 may occur. Corrections are determined and published by the U. S. Naval Observatory. The time signals are maintained in close agreement with UT2 by properly offsetting the broadcast frequency from the USFS at the beginning of each year when necessary. This new system was commenced on January 1, 1960.

Subsequent changes were as follows:

FREQUENCY OFFSET, WITH REFERENCE TO THE USFS

January 1, 1960,	−150 parts in 10^{10}
January 1, 1962,	−130 parts in 10^{10}

* Received August 27, 1962.

[1] Refer to "National Standards of Time and Frequency in the United States," PROC. IRE, vol. 48, pp. 105–106; January, 1960.

* Received July 30, 1962.

[1] W. Rindner, "Anisotropic strain effects in *p-n* junctions," *Bull. Am. Phys. Soc.*, vol. 7, pp. 65–66; 1962.

[2] H. Hall, J. Bardeen, and G. Pearson, "The effects of pressure and temperature on the resistance of *p-n* junctions in Ge," *Phys. Rev.*, vol. 84, pp. 129–132; 1951.

[3] S. Miller, M. Nathan, and A. Smith, "Pressure dependence of the current-voltage characteristics of Esaki diodes," *Phys. Rev. Lett.*, vol. 4, pp. 60–62; 1960.

[4] W. Rindner and I. Braun, "On the Effects of Elastic and Plastic Deformation of *p-n* Junctions in Semiconductors," *Proc. Internat'l Conf. on the Physics of Semiconductors*, Exeter, England; July, 1962, to be published.

[5] W. G. Pfann, "Improvement of semiconductor devices by elastic strain," *Solid State Electronics*, vol. 3, pp. 261–267; November, 1961.

TIME ADJUSTMENTS, WITH REFERENCE TO THE TIME SCALE UT2

December 16, 1959,	retardation,	20 milliseconds
January 1, 1961,	retardation,	5 milliseconds
August 1, 1961,	advancement,	50 milliseconds

Adjustments were made at 0000 UT on the foregoing dates; an advancement means that the signals were adjusted to occur at an earlier time than before.

WWV FREQUENCY WITH RESPECT TO U. S. FREQUENCY STANDARD

1962		Parts in 10^{10}
July	1	−129.2
	2	−129.2
	3	−129.0†
	4	−130.0
	5	−129.9
	6	−129.9
	7	−129.9
	8	−129.9
	9	−129.9
	10	−129.8
	11	−129.7
	12	−129.6
	13	−129.6
	14	−129.5
	15	−129.5
	16	−129.4
	17	−129.4
	18	−129.4
	19	−129.3
	20	−129.3
	21	−129.3†
	22	−130.7
	23	−130.6
	24	−130.5
	25	−130.3
	26	−130.1
	27	−130.0
	28	−129.9
	29	−129.9
	30	−130.0
	31	−130.0

† A minus sign indicates that the broadcast frequency was below nominal. The uncertainty associated with these values is $\pm 5 \times 10^{-11}$.

WWV frequency adjusted as follows:
July 3, -1.1 parts $\times 10^{-10}$ at 1900 UT.
July 21, -1.4 parts $\times 10^{-10}$ at 1900 UT.

NATIONAL BUREAU OF STANDARDS
Boulder, Colo.

Miniature Electrically Tunable YIG Band-Pass Filter*

Early designs of electrically tunable band-pass filters using yttrium-iron-garnet (YIG) spheres have concentrated on single-resonator filters.[1,2] However, the performance of single-resonator filters is limited by spurious responses, (about 15 db below the main response), caused by magnetostatic modes.[3] Two-resonator filters have reduced this spurious response level to about 35 db below the main response; however, such filters are ordinarily large. Those two-resonator filters already developed for use in S band[4] and *X* band[1] are typical examples. In many applications, it is highly desirable to miniaturize these YIG filter structures because a reduction in filter size also reduces the magnet size and tuning-power requirements.

We have developed a two-resonator YIG filter (Fig. 1) that tunes from 2.4 to 4.8 Gc, which has a residual response level of 60 to 80 db, and has spot-frequency spurious resonances below 44 db. It has a volume, excluding the magnet, of less than 0.25 cubic inch. The filter is a miniaturized version of a basic structure used by Carter.[4] It consists of two garnets coupled to two strip transmission lines with a slot in the common ground plane. The slot is the coupling element between the resonators. The small size is achieved primarily by using strip-transmission-line ground-plane separations of 0.162 inch. Previously fabricated structures have had large ground-plane separations (0.6 inch).[4] Power is coupled from one line to the other when the YIG spheres are at their ferrimagnetic resonant frequencies. The filter has a two-pole response with a nominal 3-db bandwidth of 35 Mc. The filter is tuned by varying the field of an electromagnetic about a permanent-magnet biasing field. The biasing field was set for 1140 oersteds, corresponding to a frequency of 3.2 Gc.

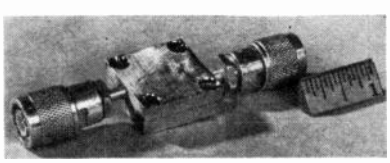


Fig. 1—Miniature two-resonator YIG band-pass filter.

The filter characteristics have been evaluated between 2.4 to 4.8 Gc. Fig. 2 shows a typical frequency response curve, and Fig. 3 shows the characteristics of the filter over the complete tuning range. The curve in Fig. 2 is centered at 3.8 Gc with a 34-Mc 3-db bandwidth and a 1.1-db insertion loss at the peak of the response. It has a frequency response equivalent to a two-pole Chebyshev filter having a 0.8-db pass-band ripple up to about 40 db of insertion loss. It then continues to a residual loss level of 83 db except for the two narrow-band responses that are 50 and 62 db down. The strongest spurious level is therefore 50 db down when the filter is tuned to 3.8 Gc. Fig. 3 is a plot of the strongest spurious-response level, the midband insertion loss, and the 3-db bandwidth vs frequency. The spurious-response level is always more than 44 db down, the 3-db bandwidth averages about 35 Mc, and the midband insertion loss is under 2 db over most of the range. The insertion loss starts to increase slightly at the low-frequency end because the unloaded *Q*'s of the YIG spheres are degraded. The filter is capable of tuning over the full 2.4 to 4.8 Gc range with 3 watts of power.

These data were measured after the resonant frequencies of the two garnets were very carefully aligned. The alignment, essential to low insertion loss, was achieved by rotating the axes of the garnets until they both had the same resonant frequencies.

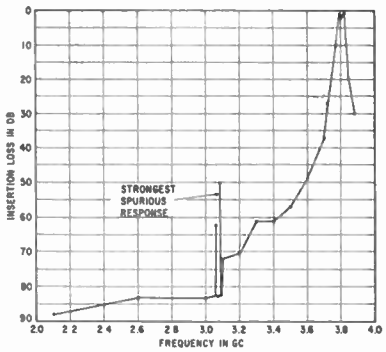


Fig. 2—Typical response curve.

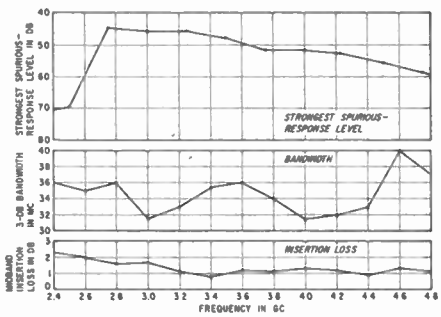


Fig. 3—Characteristics of miniature two-resonator YIG band-pass filter.

This procedure is effective because the insertion loss is only slightly higher (0.1 to 0.3 db) over the entire tuning range than when the filter was peaked to any specific frequency by individually tuning each sphere. Therefore, the filter can be accurately tuned by only one adjustment—the variation of the magnetic field.

In many receiver applications, the presence of the few, narrow-band spurious-response levels appearing at any given filter center frequency do not necessarily determine the strongest spurious response of the entire receiver. By careful choice of the IF amplifier center frequency, it is often possible to reject the few strong filter spurious levels present. Where this is feasible, the YIG filter described would have an effective spurious response level equal to its residual level of 60 to 80 db.

Work is continuing on techniques to reduce the strongest spurious response levels to more than 60 db below the main response level and to increase the tuning range. The present filter tunes from 2.1 to 5.6 Gc with slightly degraded performance between 2.1 to 2.4 Gc and 4.8 to 5.6 Gc. The results to date indicate that miniature YIG filters with two resonators can be constructed and tuned over more than one octave with a single voltage control. The rapid and remote tuning capabilities of these filters can be used extensively in microwave receiving systems.

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* Received June 25, 1962; revised manuscript received, July 2, 1962.

[1] P. S. Carter, Jr., "Design of magnetically-tunable microwave filters using single-crystal yttrium iron-garnet resonators," IRE TRANS. ON MICROWAVE THEORY AND TECHNIQUES, vol. MTT-9, pp. 252–260; May, 1961.
[2] K. L. Kotzebue, "Broadband electronically-tuned microwave filters," 1960 IRE WESCON CONVENTION RECORD, pt. 1, pp. 21–27.
[3] L. R. Walker, "Magnetostatic modes in ferromagnetic resonance," *Phys. Rev.*, vol. 105, pp. 390–399; January 15, 1957.
[4] P. S. Carter, Jr., *et al.*, "Microwave Filters and Coupling Structures," Stanford Res. Inst., Menlo Park, Calif., Rept. No. 3, Contract DA-36-039-sc-87398, SRI Proj 3527; October, 1961.

Use of Electro-Optical Shutters to Stabilize Ruby Laser Operation*

Emission from a ruby laser is normally characterized by extreme fluctuations, apparently arising from an instability of stimulated emission oscillations in a resonant cavity, and further complicated by the existence of many degrees of freedom (modes) which can be excited.

The use of a feedback control to stabilize oscillation level has apparently not been previously attempted. However, the theory and practice of feedback controls is well established.

We have used a Kerr electro-optical shutter as the control element in a feedback loop to stabilize the operation of a ruby laser. The stabilizing effect is dramatic; output is changed from irregular "spiking" to essentially steady emission which generally follows the intensity of the pumping illumination. The two oscillograms of Fig. 1 compare the output of the laser with and without feedback stabilization.

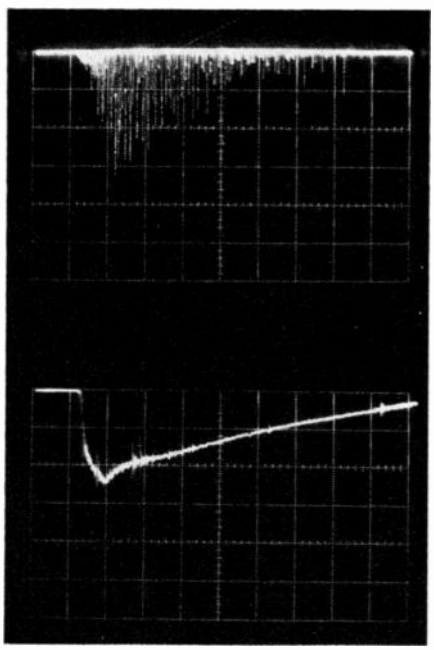


Fig. 1—Upper trace oscillogram of 1: Ruby laser output without stabilization. Lower trace, oscillogram of ruby laser with feedback stabilization. Sweep speed: 200 microseconds per major division in both traces.

The experimental arrangement is shown in Fig. 2. Both the Kerr cell and a polarizer[1] are included, together with the ruby, in the optical resonator. The mirrors are adjustable for parallelism, and the ruby is pumped with a helical xenon flash tube.

Feedback is accomplished by directing a portion of the laser output onto a high current photodiode. The photodiode drives the Kerr cell directly, the load impedance being chosen to provide suitable gain and phase characteristics. The Kerr cell is specifically designed for laser modulation. A bias voltage is applied in order to bring the Kerr cell to a more sensitive portion of its characteristic.

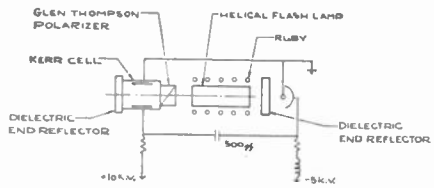


Fig. 2—Schematic of Kerr electro-optical stabilized ruby laser.

We believe that the value of this development lies in two broad areas. First, it is now feasible to control a ruby laser, and also to introduce amplitude modulation for many obvious applications. Second, this development makes the ruby laser operation much more amenable to theoretical study and analysis. Improvement in line width, coherence, and beam angle is anticipated to result from the more stable operation.

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* Received August 1, 1962.

[1] The polarizer is not essential, since the dichroism of the ruby will generally suffice.

Elimination of Re-Entry Radio Blackout*

Mr. H. Hodara has suggested[1] that the propagation of an EM wave through a collisionless plasma can be greatly improved by the use of a longitudinal magnetic field; in particular he mentioned that the propagation characteristics of a right-handed (RH) circularly polarized wave are much better than those of a left-handed (LH) circularly polarized wave. Although this is true for an infinite medium, it would appear that in the more realistic case of transmission through finite slabs this advantage is vitiated.

To see this consider the propagation of a VHF signal ($\omega = 1.6 \times 10^9$) through a 1-cm thick homogeneous plasma when a 570-gauss ($\omega_b = 10^{10}$) longitudinal magnetic field is employed. Fig. 1 shows the transmission coefficient as the plasma frequency is varied. For small values of the plasma frequency ($\omega_p \sim 10^{10}$), there is substantially complete transmission for both right- and left-hand signals. As ω_p increases, however, the transmission coefficient for the LH signal decreases monitonically at a rate which becomes proportional to

$$e^{-[2x/c\sqrt{\omega_b/\omega)+1}]\,\omega_p}$$

(x is the slab thickness and c is the velocity of light).

The RH signal does not have this "decaying exponential" behavior since for large

* Received March 7, 1962.

[1] H. Hodara, "The use of magnetic fields in the elimination of the re-entry radio blackout," PROC. IRE, vol. 49, pp. 1825–1830; December, 1961.

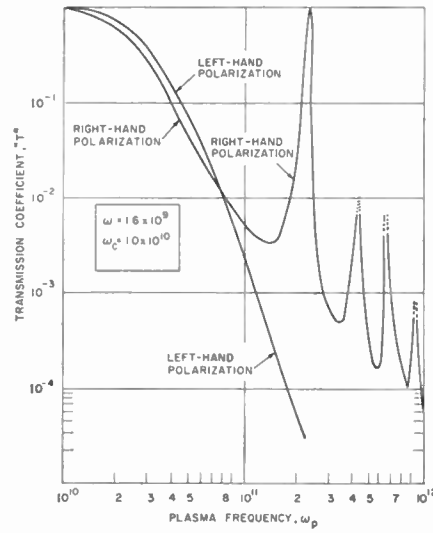


Fig. 1.

plasma frequencies its index of refraction

$$\left(n = \left[1 + \frac{\omega_p^2}{\omega(\omega_b - \omega)}\right]^{1/2}\right)$$

is real. Thus, the problem is simply one of transmission through a homogeneous slab of real index of refraction as either the thickness or index of refraction is increased.

As a practical matter, however, Fig. 1 shows that the RH polarization is no improvement over the LH wave. The transmission coefficient is simply

$$T = \left[1 - \frac{n^2 - 1}{4n^2} \sin^2 kxn\right]^{-1},$$

and for large index of refraction ($\omega_p^2 \gg \omega\omega_b$) there will be, on the average, a large reflection. In fact in this case the RH signal reaches a 10-db attenuation point slightly before the LH signal.[2] The fact that aperiodically the RH signal has complete transmission is of little help since the plasma conditions are continuously changing and one would have sporadic transmission at best. For an inhomogeneous plasma the peaks are lowered and the valleys raised so that in a more realistic case one would not even get these sharp bursts of transmission.

Even though an RH signal has no advantage over an LH signal, there is still the question of whether a magnetic field has any advantage at all. The answer to this question can be seen by examining Fig. 2 where transmission has been computed for a 5700-gauss ($\omega_b = 10^{11}$) longitudinal magnetic field. By increasing ω_b by a factor of 10 one is able to stand a three-fold increase in ω_p before reaching a 10-db attenuation. Thus one can get through a plasma by using a strong magnetic field, but as a practical matter the necessary field strengths are often prohibitively large.

[2] It might be objected that one might be willing to live with 20 or 30 db attenuation by going to high power systems in which case the RH system is better than the LH system. This has a disadvantage both in that high powers tend to induce antenna breakdown and also that there will be strong reflection and thus mismatch, particularly for the low-frequency systems intimated by the use of magnetic fields. (A magnetic field is of no advantage unless the signal frequency is less than the cyclotron frequency.)

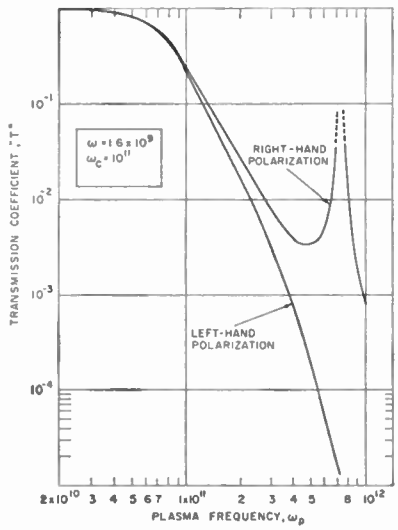


Fig. 2.

The conclusion is thus two-fold:

1) There is no advantage of one polarization over another.
2) The only way to use a magnetic field to get through a *dense* plasma is to use a *strong* magnetic field—there is no royal road to propagation.

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Author's Comments[3]

I would like to thank Mr. Mullin for the interest shown in my work.

Mr. Mullin's comments deserve to be discussed and I believe I can answer the questions he raises, if not to his satisfaction, at least to mine. There are essentially two statements to his thesis, first that there is no advantage in using an RH wave over an LH wave, second that in order to propagate through a dense plasma, strong static magnetic fields are required. Both of his conclusions are stated with much generality and their value can only be determined when expressed quantitatively. It is the purpose of this reply to determine this value.

Let us discuss the first statement. According to Mr. Mullin it is implied in my paper[1] that an RH wave is "much better" than an LH wave. Nowhere in the paper is there such a statement as it does not convey any quantitative information. It is stated, however, on page 1829 of the paper, that for a signal frequency below cyclotron resonance, 3 db of additional power penetrates the plasma medium if the incident wave is RH circularly polarized instead of linearly polarized. Whether 3 db is "much better" than nothing is another matter. On the other hand for the collisionless semi-infinite plasma under discussion at signal frequencies below cyclotron resonance, an RH wave is partially transmitted while an LH wave is totally reflected in which case the ratio of transmitted RH to LH wave is infinite; for this case indeed an RH wave is "much better" than an LH wave. In reality, as Mr. Mullin correctly points out, the plasma sheath surrounding a re-entering space vehicle has a finite width and is a far cry from the semi-infinite medium approximation.

A propagation analysis based on a homogeneous finite width plasma slab may be just as far a cry from reality since it is well known that the electron density exhibits a sharp discontinuity at the skin of the vehicle and decays approximately exponentially towards the shock layer boundary (the statement "sharp discontinuity" may be as vague to Mr. Mullin as the statement "much better" mentioned above). It would seem that the actual model of an inhomogeneous plasma sheath lies somewhere in between the two approximations: the homogeneous finite width slab and the homogeneous semi-infinite medium. As pointed out by Mr. Mullin, in the case of an inhomogeneous slab, the valleys in his curve of Fig. 1 are raised and the advantages of an RH wave over an LH wave are improved, provided of course it is possible to transmit with the attenuation levels corresponding to the valleys of the curve Fig. 1 rather than to the peaks. This brings up Mr. Mullin's point of contention, namely that it is impractical or at least disadvantageous to transmit with such high attenuation levels because of the deleterious effects brought about by the additional required power as described in the footnote[2] to his comments. Since this statement applies not only the the pros and cons of an RH wave over an LH wave but also to the relative advantages or disadvantages of propagation with and without magnetic field, let us discuss then Mr. Mullin's second conclusion before answering the statement in question.

Let us briefly analyze how the power transmission coefficient varies for a homogeneous plasma slab of finite width L for the cases of no magnetic field, transverse magnetic field, and longitudinal magnetic field. The transmission coefficients for these cases are respectively:

$$T_0 = \left| \frac{1}{c_0 + \frac{j}{2} s_0 \left(b_0 + \frac{1}{b_0} \right)} \right|^2 \quad \text{Isotropic case } (B_0 = 0), \tag{1}$$

$$T_T = \left| \frac{1}{c_T + \frac{j}{2} s_T \left(b_T + \frac{1}{b_T} \right)} \right|^2 \quad \text{Transverse static magnetic field case} \tag{2}$$

$$T_L = \left| \frac{1/2}{c_1 + \frac{j}{2} s_1 \left(b_1 + \frac{1}{b_1} \right)} + \frac{1/2}{c_2 + \frac{j}{2} s_2 \left(b_2 + \frac{1}{b_2} \right)} \right|^2 + \left| \frac{1/2}{c_1 + \frac{j}{2} s_1 \left(b_1 + \frac{1}{b_1} \right)} - \frac{1/2}{c_2 + \frac{j}{2} s_2 \left(b_2 + \frac{1}{b_2} \right)} \right|^2 \tag{3a}$$

Longitudinal magnetic field case; incident wave linearly polarized,

$$T_{1,2} = \left| \frac{1}{c_{1,2} + \frac{j}{2} s_{1,2} \left(b_{1,2} + \frac{1}{b_{1,2}} \right)} \right|^2 \quad \text{Longitudinal magnetic field case with incident wave circularly polarized to the right } (T_1) \text{ or to the left } (T_2). \tag{3b}$$

The following notation is used in the above equations: the subscripts 0, T, L refer, respectively, to isotropic, transverse, and longitudinal static magnetic field case. The subscripts 1 and 2 refer to an RH mode and an LH mode and the various symbols are defined below:

$$c_n = \cos b_n l \qquad (n = 0, T, 1, 2)$$
$$s_n = \sin b_n l$$
$$l = \frac{L}{\lambda\!\!\!^{-}_v}$$
$$\lambda\!\!\!^{-}_v = \frac{\lambda_v}{2\pi} \quad (\lambda = \text{wavelength of signal in free space})$$
$$b_0 = \sqrt{\epsilon_1}$$
$$b_T = \sqrt{\epsilon_1 + \frac{\epsilon_2^2}{\epsilon_1}}$$
$$b_{1,2} = \sqrt{\epsilon_1 \mp j\epsilon_2} \quad (- \text{ sign corresponds to subscript 1, } + \text{ sign to subscript 2})$$
$$\left.\begin{aligned} \epsilon_1 &= 1 + \frac{\omega_p^2}{\omega_b^2 - \omega^2} \\ \epsilon_2 &= j \frac{\omega_b}{\omega} \frac{\omega_p^2}{\omega_b^2 - \omega^2} \end{aligned}\right\} \text{(collisionless case: } \nu = 0).$$

The case of practical interest in the blackout problem is

$$\omega^2 \ll \omega_b^2 \ll \omega_p^2 \tag{4}$$

and (1), (2), and (3b) reduce to

$$T_0 \sim 16 \frac{\omega^2}{\omega_p^2} e^{-2(\omega_p/\omega)(L/\lambda\!\!\!^{-}_v)} \tag{5}$$

$$T_T \sim 16 \frac{\omega^2}{\omega_p^2} e^{-2(\omega_p/\omega)(L/\lambda\!\!\!^{-}_v)} \tag{6}$$

$$\bar{T}_1 \approx 4 \frac{\omega_b \omega}{\omega_p^2} \tag{7}$$

$$T_2 \approx 4 \frac{\omega_b \omega}{\omega_p^2} e^{-2(\omega_p/\sqrt{\omega_b \omega})(L/\lambda\!\!\!^{-}_v)}. \tag{8}$$

Eq. (7) describes the locus of the minima of the transmission coefficient for the RH wave; it describes the worst possible condition for this case.

Although some of the parameters used by Mullin in this discussion do not quite fulfill

[3] Received July 26, 1962.

the inequality (4), the discrepancy between the results using (5)–(8) and Mr. Mullin's results is not higher than 20 per cent. This error is tolerable for the present discussion since it simplifies the analysis and does not limit the validity of the conclusions to be reached. No generality is lost and the exact formulas (1), (2), and (3) are available if desired.

Under condition (4), comparison of (5) and (6) shows no reduction in the transmission coefficient with a transverse magnetic field. On the other hand with an RH wave and a longitudinal magnetic field, although both $\bar{T}_1$ and T_0 decrease inversely with the square of the plasma frequency, T_0 decreases exponentially with the slab width while T_1 is not affected. In particular, the ratio

$$\frac{\bar{T}_1}{T_0} = \frac{\bar{T}_1}{T_T} = \frac{1}{4}\frac{\omega_b}{\omega} e^{2(\omega_p/\omega)(L/\lambda_v)} \qquad (9)$$

shows clearly the relative improvement to be obtained with a longitudinal static magnetic field and an RH wave. As ω_p and L increase, the relative decibel improvement over no magnetic field increases rapidly as seen from (9). For example if $\omega_p = 1.6 \times 10^9$, $\omega_b = 1.0 \times 10^{10}$, $L = 1$ cm, then $10 \log_{10} \bar{T}_1/T_0 = 48$ db. Thus if a "strong" longitudinal magnetic field is required to propagate through the plasma, for the same attenuation loss, a "much stronger" (48 db stronger) transverse magnetic field is necessary. If L is doubled, the relative db improvement is then $10 \log_{10} \bar{T}_1/T_0 = 94$ db. Such improvements occur at attenuation levels, of the order of 25 db; this brings up the unanswered question and the very same point of contention brought out by Mr. Mullin's footnote,[2] namely, is it advantageous or practical to transmit with as much attenuation? If one wants to assume, as Mr. Mullin does, that it is disadvantageous to raise the transmitter power to overcome the 25-db attenuation (in spite of the fact that with no magnetic field the attenuation would be 94 db for a 2-cm thick plasma sheath), the answer lies in the fact that it may be possible not to raise the transmitting power if the ground receiving system has sufficient sensitivity. In other words a criterion for transmission is not only the plasma attenuation but also the minimum detectable SNR at the receiver. As an example,[4] SNR are plotted in Fig. 3 for the re-entry path shown in Fig. 3, under two conditions: no static magnetic field, $B_0 = 0$ and $B_0 = 1000$ gauss. The curves are self-explanatory; the results show that 1000 gauss seem to be adequate to insure a minimum SNR of 25 db throughout re-entry. In making these calculations the variations of plasma sheath thickness, plasma frequency and collision frequency as a function of altitude have been taken into account. These results by themselves do not guarantee the feasibility of the magnetic field approach. Much more experimental data and flight testing are necessary before any conclusion can be drawn as to the validity or invalidity of this approach. Further theoretical studies are necessary to take into account the characteristics of the radiating structure. Recent results[5] shown in Fig. 5 illustrate the pattern to be expected from a narrow rectangular slot covered with a homogeneous gyroplasma sheath. Each curve in Fig. 5 corresponds to a different thickness and the variations in amplitudes of the patterns match qualitatively the oscillating behavior of the transmission coefficient in Mr. Mullin's Fig. 1. The impedance of a radiation structure covered with gyroplasma is also being investigated at this laboratory. It is believed that the results of this impedance analysis coupled with recent experimental data will shed additional light on the feasibility of the static magnetic field approach.

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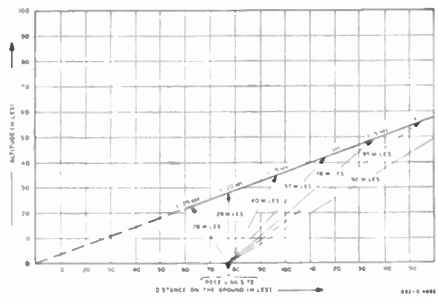


Fig. 3—Range vs altitude at a selected receiving site for typical R/V.

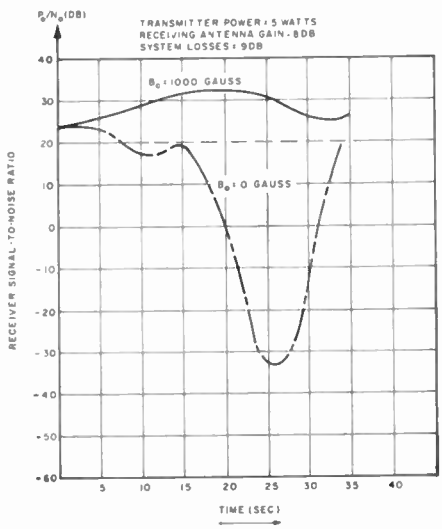


Fig. 4—SNR at selected receiving site with transmitter located aft of the R/V.

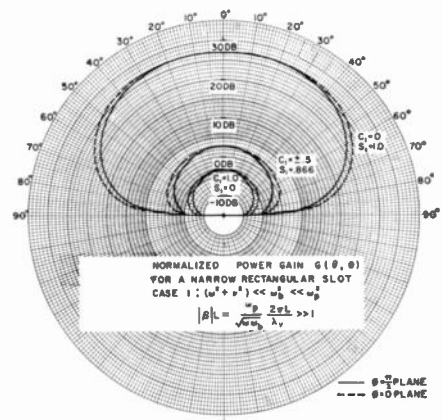


Fig. 5.

Comments on "Broad-Band Parametric Amplifiers by Simple Experimental Techniques"*

Supplementary technical information is currently available which did not appear in the original paper.[1] This is included in the following miscellaneous comments:

1) All results were achieved using RG-52/U waveguide diode mounts (propagating to both pump and idle frequencies in the TE_{10} mode) and coaxial circuitry for the signal. A photograph of a typical breadboard model of the paramp is shown in Fig. 1. This paramp configuration is a modified version of structures previously developed at RCA.[2,3] The signal circuit contains a coaxial low-pass filter and a coaxial double quarter-wave transformer. A Melabs Model X-344 coaxial ferrite circulator was employed. This circulator had a useful bandwidth from 2.0 kMc to 3.0 kMc. The triple screw tuner shown in the photograph was

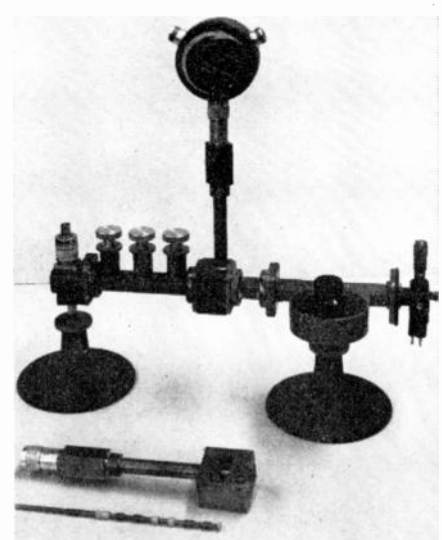


Fig. 1.

* Received March 23, 1962.

[1] B. B. Bossard and R. Pettai, "Broad-band parametric amplifiers by simple experimental techniques," PROC. IRE (*Correspondence*), vol. 50, pp. 328–329; March, 1962.

[2] R. Pettai, B. Bossard, and S. Weisbaum, "Single-diode parametric upconverter with large gain-bandwidth product," PROC. IRE (*Correspondence*), vol. 48, pp. 1323–1324; July, 1960.

[3] R. M. Kurzrok, "Parametric amplifier device," U. S. Patent No. 2, 970, 275; January 31, 1961.

[4] H. Hodara, "Calculations Pertinent to the Re-Entry Blackout," The Hallicrafters Co., Chicago, Ill., Tech. Rept., January, 1961.

[5] H. Hodara, "Radiations from a gyroplasma sheathed aperture," to be published in IRE TRANS. ON ANTENNAS AND PROPAGATION, January, 1963.

used to trim the idle circuit; however, the gain response shown in Fig. 1 of the original paper was achieved using three commercially available slide screw tuners in the idle circuit. More recent results at S band have entailed a gain of 11.2 db, a 1-db bandwidth of 580 Mc and a 3-db bandwidth of 700 Mc.

2) Best paramp performance was obtained using a partial height capacitive post for impedance. Similar techniques have also been used in broad-band waveguide conventional mixers.

3) External idler resistive loading was employed. Although paramp noise figures for a given varactor diode are no longer optimum, the noise figures obtained are only about 0.3 db worse for the case of no external idler loading. This should be satisfactory for most applications. Other significant broad-band microwave parametric amplifiers reported in the literature[4-7] *do not* use external idler loading.

4) A silicon cartridge-type diode with avalanche breakdown at -5 volts was used in achieving the best results. Typical amplifier gain saturation for this diode is as follows:

a) 1-db deviation from small signal gain at signal input level of -13 dbm
b) 3-db deviation from small signal gain at signal input level of -7 dbm.

The large dynamic range of the paramp yields excellent intermodulation performance. For example, a small signal of -90 dbm in the presence of a large interfering signal of -10 dbm caused no discernible intermodulation products.

5) The diode used apparently did not exhibit an abnormally high value of $C_1/2C_0$ due to anamolous minority carrier-charge storage effect; however, this has not been conclusively proven. For broad-band parametric amplifier operation at microwave frequencies, diodes must have carefully controlled values of junction capacitance, case capacitance, and lead inductance. Availability of suitable diodes is not a major problem in the advanced development of S-band paramps with bandwidths of 400 Mc; however, selected diodes are required for S-band paramps having bandwidths of 700 Mc to 900 Mc.

6) Preliminary measurements of signal and idle pass bands are accomplished using swept frequency techniques. If a particular diode looks promising, detailed point-by-point data is subsequently taken. Point-by-point measurement of the signal pass band entails use of a high-level signal input (about zero dbm) to obtain sufficient rectification. Point-by-point measurement of the idle pass band entails use of a low-level idle input (about -30 dbm). Both of these pass-band measurements can be made under static operating conditions (*i.e.*, without pump power).

7) The broad-band paramp can be tuned over a 1000 Mc frequency range by simply changing the pump frequency. Instantaneous single-tuned bandwidths of 60 Mc were obtained for amplifier gain levels of 16 db. Further details of this experimentation will be reported subsequently.

The experimental work of B. S. Perlman is acknowledged.

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[4] J. Kliphuis, "C-band nondegenerate parametric amplifier with 500-Mc bandwidth," Proc. IRE (*Correspondence*), vol. 49, p. 961; May, 1961.
[5] B. T. Vincent, "A C-band parametric amplifier with large bandwidth," Proc. IRE (*Correspondence*), vol. 49, p. 1682; November, 1961.
[6] K. M. Johnson, "Broad-band S-band parametric amplifier," Proc. IRE (*Correspondence*), vol. 49, p. 1943; December, 1961.
[7] K. M. Johnson, "900-Mc, non-degenerate X-band amplifier," Proc. IRE (*Correspondence*), vol. 50, p. 332; March, 1962.

New Helium-Neon Optical Maser Transitions*

This letter is intended as a summary of our optical maser investigations in RF discharges containing mixtures of helium and neon. For the purpose of making these investigations, a special Fabry-Perot interferometer was constructed. A confocal system[1] was chosen for ease in alignment and Brewster angle windows were incorporated to permit the use of external mirrors.[2] The use of external mirrors is desirable to eliminate damage to the high-reflectance dielectric coatings during bakeout of the vacuum system and to permit changing mirrors without opening the system to air. In order to increase the optical gain, a long discharge length was used; a 225-cm length (quartz) discharge tube was found to be a reasonable compromise between the various practical considerations involved. Since previous gain studies[3] in the helium-neon system and in pure neon showed that the gain varied roughly as the reciprocal of the tube diameter, a small diameter was employed. A 7-mm inside diameter was found to be a realistic compromise between gain and diffraction loss (which is primarily determined by tube-alignment errors with these dimensions). A more detailed account of the maser and of the vacuum techniques will be given elsewhere.[4]

Oscillation in the helium-neon system has been reported previously[5] on the neon transitions at 1.1177, 1.1523, 1.1614, 1.1985 and 1.2066 microns and these lines were easily observed with the present maser. Of these five transitions, the strongest oscillation is obtained at 1.1523 microns. We have obtained maser oscillation on five additional $2s$-$2p$ transitions (Paschen notation) in this system and a summary of the present results is contained in Table I. The 1.5231 micron transition is second in intensity only to the previously reported line at 1.1523 microns. The partial pressures of helium and neon were approximately in the ratio of 10:1 with the total pressure $\cong$1 to 2 mm Hg for these measurements.

TABLE I
New Helium-Neon Optical Maser Transitions (Paschen Notation)

Transition	λ_{air} (microns)
$2s_3-2p_7$	1.0798
$2s_2-2p_6$	1.0844
$2s_4-2p_8$	1.1143
$2s_5-2p_8$	1.1390
$2s_2-2p_2$	1.1767
$2s_2-2p_1$	1.5231

It is reasonable to expect population inversions on all of the thirty allowed $2s$-$2p$ transitions of neon (ranging from about 0.9 to 1.8 microns) in this system.[5,6,7] Needless to say, the optical gain must exceed the loss in the maser for oscillation to take place on a given transition. The gain obtainable will, of course, depend on the oscillator strength for the transition as well as on the population difference; and one expects oscillation to occur most readily on the stronger lines. Koster and Statz[8] have calculated approximate values for these relative transition probabilities. Although there is no pronounced correlation between the observed gain and the calculated transition probabilities, all but two transitions ($2s_3$-$2p_7$ at 1.0798 microns) and $2s_2$-$2p_6$ at 1.0844 microns on which oscillation has been obtained have at least been reported by Koster and Statz to have nonzero probabilities. The two discrepancies presumably arise from a failure of the jl-coupling approximation used in the calculation. In this connection, it should be noted that many of the "forbidden" transitions listed by Koster and Statz are conspicuously present (though generally weak) in the spontaneous emission spectrum.

We investigated the maser output in the vicinity of the strong maser line at 1.1523 microns with a high resolution (1:300,000) grating spectrometer and found no sign of oscillation on the closely neighboring $2s_4$-$2p_7$ transition at 1.1525 microns. A similar result was obtained previously[9] with a one meter (lower gain), plane parallel maser. The difference-wavelength fringes that would be provided by simultaneous oscillation on these two neighboring lines would be of considerable practical value in long-distance interferometry.

Attempts to obtain oscillation on the shortest wavelength $2s$-$2p$ transitions—in particular, the $2s_4$-$2p_{10}$, $2s_5$-$2p_{10}$, and $2s_3$-$2p_{10}$ transitions at 0.949 μ, 0.967 μ, and 0.899 μ

* Received July 31, 1962.
[1] G. D. Boyd and J. P. Gordon, "Confocal multimode resonator for millimeter through optical wavelength masers," *Bell Sys. Tech. J.*, vol. 25, pp. 489–509; March, 1961.
[2] W. W. Rigrod, H. Kogelnick, B. J. Brangaccio, and D. R. Herriott, "Gaseous optical maser with external concave mirrors," *J. Appl. Phys.*, vol. 33, pp. 743–744; February, 1962.
[3] W. R. Bennett, Jr., "Recent progress in experiments with helium-neon masers," *Bull. Am. Phys. Soc. II*, vol. 7, p. 15; January, 1962.
[4] C. K. N. Patel, "Optical power output in He-Ne and pure neon masers," *J. Appl. Phys.*, to be published.
[5] A. Javan, W. R. Bennett, Jr., and D. R. Herriott, "Population inversion and continuous optical maser oscillation in a gas discharge containing a He-Ne mixture," *Phys. Rev. Lett.*, vol. 6, pp. 106–110; February, 1961.
[6] A. Javan, "Optical maser oscillations in a gaseous discharge," in "Advances in Quantum Electronics," J. Singer, Ed., Columbia University Press, New York, N. Y., pp. 18–27; 1961.
[7] W. R. Bennett, Jr., "Radiative lifetimes and collision transfer cross-sections of excited atomic states," in "Advances in Quantum Electronics," J. Singer, Ed., Columbia University Press, New York, N. Y., pp. 28–43; 1961.
[8] G. F. Koster and H. Statz, "Probabilities for the neon laser transitions," *J. Appl. Phys.*, vol. 32, pp. 2054–2055; June, 1961.
[9] W. R. Bennett, Jr., "Hole burning effects in a He-Ne optical maser," *Phys. Rev.*, vol. 126, pp. 580–593; April, 1962.

initially proposed by Javan[10]—have been unsuccessful so far. This failure may have its origin in a larger population of the $2p_{10}$ in respect to the other $2p$ levels—for example, through collision transfer effects in the $2p$ group. However, there is no reason to believe that these transitions are not inverted at low discharge intensities and it seems probable that oscillation might be obtained by going to still longer discharge lengths and/or smaller tube diameters. Our results indicate that the discharge is at least highly transparent at 0.967 microns, and also on several of the $3d$-$2p$ transitions. The latter is particularly significant since the $3d$ levels fall too far in energy above the $He(2^3S_1)$ to be excited by this helium metastable with any appreciable degree of probability. On the other hand, many of the $3d$ levels should have large cross sections for excitation by electron impact. This general property of $3d$ levels has been confirmed by the authors[11] by obtaining maser oscillation on the $3d$-$2p$ transitions in other pure noble gases.

The above discussion has, of course, dealt entirely with the infrared transitions of neon and primarily with the $2s \rightarrow 2p$ transitions. It should be mentioned that oscillation has also been obtained [12] recently in a helium-neon maser on the $3s_2 \rightarrow 2p_4$ transition at 0.6328 microns. The processes involved in the excitation of this visible transition, of course, differ in detail from those considered here.

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[10] A. Javan, "Possibility of production of negative temperature in gas discharges," *Phys. Rev. Lett.*, vol. 3, pp. 87–89; July, 1959. "Quantum Electronics," C. H. Townes, Ed., Columbia University Press, New York, N. Y., pp. 564–571; 1960.
[11] C. K. N. Patel, W. R. Bennett, Jr., W. L. Faust, and R. A. McFarlane, "Infrared spectroscopy using stimulated emission techniques," *Phys. Rev. Lett.*, vol. 9, pp. 102–109; August, 1962.
[12] A. D. White and J. D. Rigden, "Continous gas maser oscillation in the visible," Proc. IRE, (*Correspondence*), vol. 50, p. 1697; July, 1962.

The Measurement of the Velocity of Light*

I was very interested in Dr. Rapier's note.[1] Although I am not working in that field, I have put considerable thought into the problem of the measurement of the velocity of light. There seems to be some evidence for light having a constant velocity relative to its source and also for there being physical reasons for this possibility. However, it seems to me that none of the experiments to show that light has a constant velocity relative to the receiver has actually shown that it has or that it hasn't. I believe that this is true because in all these experiments an important factor has been neglected. In order to explain what this factor is, let me use an analogy in the radio-frequency part of the electromagnetic spectrum.

Suppose that we want to measure the velocity of a radio signal sent back from a rocket receding from the earth. We could measure the difference in arrival time (of different parts of some identifiable wavefront, for instance) at two spaced receivers. However, if the velocity of the signal were constant relative to the rocket and independent of the receiver, and if we interposed a receiver and reradiator in the path between the rocket and the measuring receivers, we would only measure the velocity of the wave from the reradiator and would have no information as to whether the velocity of the signal was constant relative to the rocket or not. This would be true even if this reradiator were a parasitic dipole or a conducting metal sheet, for instance. The electric vector of the wave from the rocket induces currents in the reflecting dipole or sheet. These currents then radiate energy as though this reflector had been the antenna on a transmitter and, hence, this dipole or sheet acts as a new source.

The light that is reflected from a mirror is reradiated in the same way, hence the mirror acts as a new source. If light has a velocity which is constant relative to its source, the light after reflection will have a constant velocity relative to the mirror and not to its primary source. A lens also acts in the same way. Light does not go through it without change but interacts with the molecules of the glass, and each molecule influences its neighbor till new light is radiated from the opposite side of the lens. The lens therefore acts as a new source. At microwave frequencies, lenses have been built with artificial dielectric made of pieces of metal. An analysis of their operation shows very little of the incident radiation as going through the structure. Most of what leaves this kind of lens is considered as reradiated from the components of the lens which make up the artificial dielectric.

There may be cases, such as for X and γ rays, where the radiation actually passes through matter without interaction with it, and similar cases for light for short distances through gas; but any experiment which attempts to measure the velocity of light directly from a source moving relative to the receiver when there is a possibility that this velocity is constant relative to its source or even has a component that is, cannot use mirrors or lenses or other transparent matter and must depend on the timing or phasing of mechanical shutters (toothed wheels, etc.) in vacuum.

I am afraid that the proposed experiment using a Čerenkov flash would have the same drawback, since even though the original light as generated had the velocity that is suggested, it has to pass through transparent *matter;* and therefore the velocity of the light that would be measured would be constant relative to the structure and not to the electrons of the source.

One could determine whether the velocity of light was constant relative to the receiver or source by measuring the difference in velocity of the light from the sun as it arrived at a satellite or at the moon, as the satellite or moon approached and receded from the sun.

Another way of determining whether or not light has a constant velocity relative to its source alone is that if light has a constant velocity relative to its source and if the velocity of light in glass is constant relative to the glass, independent of the velocity of the impinging light, then light from a source receding from a piece of glass will be refracted less than light from a source moving at the same velocity as the glass. Therefore, if light has a constant velocity relative to its source alone the relative velocity of the source of light could be measured by determining the focal length of a lens of which a curved surface was the first matter that intercepted the light from that source. Also, since the light reflected from a mirror would have a different velocity than that incident from a moving source, the angle of reflection would *not* be equal to the angle of incidence and therefore a similar experiment could be performed with mirrors.

A very sensitive determination of the relative velocities of light from a moving source and a stationary one can be made, if the above conditions apply, by interposing a thin piece of glass or a mirror in the path of a portion of the light from a moving source. Either the glass or the mirror will produce light coherent with the light from the moving source but now coming from a stationary source (the piece of glass or the mirror). If this portion and the unchanged portion are allowed to travel some distance and then combined after traversing equal optical path lengths, they will interfere with each other. By slight adjustment of the length of one path, cancellation of reinforcement can be obtained. If the lengths are adjusted for cancellation with a stationary source, then cancellation will not occur for a moving source. This apparatus may be considered to be a modified Rayleigh interferometer. For a distance of travel of one meter, one cycle shift in the cancellation point will be caused by difference in velocity of approximately 5×10^{-7} c for green light. This is about 300 mph or about 500 ft/sec. This is much too sensitive to be used on the moon or a satellite for measuring light from the sun, but since it should be easy to detect 1/10 of a cycle shift, a simple vibrating source (which might be the glass condensing lens of the interferometer through which light passes or a mirror from which it is reflected, as well as a direct source) driven by a loud-speaker element might be used. 1000 cps of $\frac{1}{4}$-inch amplitude gives approximately 130-ft/sec peak velocity. The output of a photocell placed in the interference area of the device would have an ac output depending on the movement of the source. As with any interferometer a visual pattern may be obtained. Then motion of the source will displace the pattern and a vibrating source will visually smear the pattern. The size of such a device is such that it could be quite reasonably placed in the vacuum of a conventional vacuum system.

An experiment of this type could add to our knowledge of the behavior of light. A positive result would indicate definitely that

* Received March 26, 1962.
[1] P. M. Rapier, "A proposed test for the existence of a Lorentz-invarient aether," Proc. IRE, vol. 50, pp. 229–230; February, 1962.

the velocity of light was constant relative to the source alone and was independent of the receiver. A negative result would be more ambiguous but would at least eliminate several possibilities.

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Intermodulation Noise in FM Troposcatter Links*

The purpose of this note is to give an approximate formula for the intermodulation noise produced by tropospheric scattering.

Measurements[1] on one path show that for small frequency deviations the signal-to-intermodulation ratio S/I as a function of peak deviation P and the highest baseband frequency F can be represented by

$$S/I = \frac{k}{(PF)^2} \tag{1}$$

where k is a constant which depends on effective distance and antenna beamwidth, and S/I is measured in the highest baseband channel. Since (1) and the shape of the measured curves for large deviations agree with the results obtained both experimentally[2] and theoretically[3] for the intermodulation produced by one delayed echo, it appears reasonable to extend these results to the troposcatter case. From Bennett, *et al.*,[3] for peak deviations much smaller or larger than the reciprocal of the echo time T,

$$S/I = \frac{k_1}{(PFT^2)^2} \qquad PT \ll 1 \tag{2a}$$

$$S/I = k_2(P/F)^3 \qquad PT \gg 1. \tag{2b}$$

Combining these gives

$$S/I = \frac{[1 + k_3T^4P^5]}{F}. \tag{3}$$

Using the measurements[1] to evaluate the constants gives

$$S/I = \frac{5700}{(PF)^2}(T_0/T)^4[1 + (1/F)(T/T_0)^4(P/9.4)^5]. \tag{4}$$

P and F are in Mc/sec, and $T_0 = 8\times10^{-8}$ secs. The minimum value is obtained when the second factor in square brackets in (4) is equal to 5/3. The equivalent time delay T is

$$T = (d_a/c)(d/R)_e^2\frac{(\alpha/\theta)^2}{0.92 + 4.1(\alpha/\theta)^2} \tag{5a}$$

$$= \frac{(d_a/c)\alpha^2}{0.92 + 4.1\,\alpha^2/(d/R)_e^2}. \tag{5b}$$

where

d_a = actual distance between transmitter and receiver,
c = velocity of light,
α = antenna beamwidth in radians (equal transmitting and receiving beamwidths),

$$(d/R)_e = \frac{\text{effective distance}}{\text{effective earth's radius}},$$

$\theta = (d/R)_e$ = nominal scatter angle, the angle between 1) a plane tangent to the earth and passing through the center of the transmitting antenna, and 2) a similar plane for the receiving antenna.

Eq. (5), adapted from Prosin,[4] takes into account the antenna beamwidth as well as the effective scattering volume. The above equations are for slot noise tests, pre-emphasis, no diversity, equal antenna beamwidths, and for a path which is approximately symmetrical (effective scattering volume not radically separated from the midpoint of the path). It should be noted that with pre-emphasis the lowest signal-to-intermodulation ratio is not necessarily obtained at the maximum baseband frequency.[3] In calculating T_0 for the Florida-Cuba path the effective distance was taken to be the actual distance, the effective earth's radius was taken as 2.4 times the actual radius,[5] and the antenna beamwidth was taken to be 2.55×10^{-2} radians.

Two paths in another part of the world have the following parameters:

TABLE I

Actual distance d_a	440 miles	194 miles
Effective distance d_e	393 miles	124 miles
Effective earth's radius R_e	5300 miles	5300 (assumed)
Antenna beamwidth α	1.25×10^{-2}	3.8×10^{-2}
Highest baseband frequency F	0.3 Mc	0.3 Mc

The measured and calculated values of the signal-to-intermodulation ratio are shown in Fig. 1. The discrepancy of 6 db between measurements and calculations for path A may indicate that (5) requires modification. However, more measurements on long paths are required to confirm this.

One method of reducing the intermodulation for $PT \ll 1$ is to shift the multiplex frequency band so that it occupies less than an octave in frequency.[2] (Thus, for a 0–300 kc system, shift the band to 300–600 kc.) The S/I ratio should be improved by a large factor.

For $\alpha/(d/R)_e \gg 1$, (5b) and (4) show that the signal-to-intermodulation ratio varies as the eighth power of $(d/R)_e$ so that the ratio will decrease by large factors during the winter. On the other hand, for $\alpha/(d/R)_e \ll 1$, the signal-to-intermodulation ratio does not depend on $(d/R)_e$. Experimental data are not available to verify these conclusions.

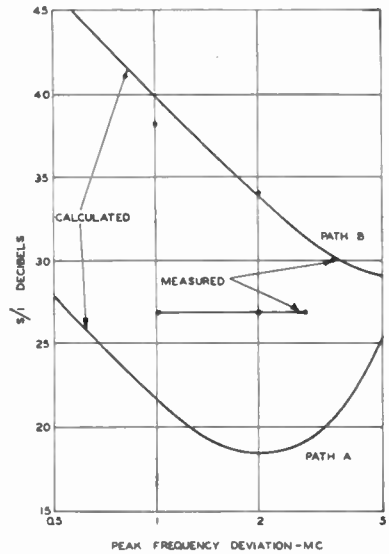


Fig. 1— Signal-to-intermodulation ratio vs frequency deviation. Parameters for the paths are given in Table I.

The theoretical formula given by Prosin[6] has the same form as (2a) for $PT \ll 1$, but the numerical coefficient is 160 times larger than the coefficient in (4). For $PT \gg 1$ the exponential factor in Prosin's formula becomes important, but only for very high deviations (over 100 Mc for the Florida-Cuba path).

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Stimulated Emission from Holmium Activated Silicate Glass*

A program of studying the optical properties of very pure silicate glasses activated with rare earth ions has been in progress for over a year now at this laboratory. One of the purposes of this work is to investigate stimulated emission processes in these glasses. The preparation of the glasses, the fabrication of the etalons and the experimental measurement techniques have been reported previously.[1] The present communication describes some preliminary investigation recently made on holmium activated silicate glass.

* Received March 26, 1962. This work was sponsored in part by the Federal Electric Corp., a subsidiary of the International Telephone and Telegraph Corp., under Contract AF 19(626)-21.

[1] C. E. Clutts, R. N. Kennedy, and J. M. Trecker, "Results of bandwidth tests on the 185 mile Florida-Cuba tropospheric scatter system," IRE TRANS. ON COMMUNICATIONS SYSTEMS, vol. CS-9, pp. 434–439; December, 1961.

[2] W. J. Albershein and J. P. Schafer, "Echo distortion in the FM transmission of frequency-division multiplex," PROC. IRE, vol. 40, pp. 316–328; March, 1952.

[3] W. R. Bennett, H. E. Curtis, and S. O. Rice, "Interchannel interference in FM and PM systems under noise loading conditions," *Bell Sys. Tech. J.*, vol. 34, pp. 601–636; May, 1955.

[4] A. V. Prosin, "Calculation of cross noise power in long distance scatter communications systems," *Radio Engrg. and Electronic Phys.*, vol. 6, pp. 9–18 eq. (46); January, 1961.

[5] R. E. Gray, "Tropospheric scatter propagation and meteorological conditions in the Caribbean," IRE TRANS. ON ANTENNAS AND PROPAGATION, vol. AP-9, pp. 492–496; September, 1961.

[6] Prosin, *op. cit.*, eq. (33).

* Received August 2, 1962.

[1] H. W. Etzel, H. W. Gandy, and R. J. Ginther, "Stimulated Emission of Infrared Radiation from Ytterbium Activated Silicate Glass," NRL Progr. Repts., pp. 27–28; February, 1962.

Stimulated emission from single crystal etalons of $CaWO_4$ activated with trivalent holmium at liquid nitrogen temperatures has been reported by Johnson, *et al.*[2] To this date, this is the only matrix in which stimulated emission from holmium has been reported and in $CaWO_4$, the emission occurs at a wavelength of 2.046 μ.

Stimulated emission of radiation has been observed in a holmium activated $LiMgAlSiO_3$ glass at liquid nitrogen temperature. This glass is of the same base composition as the glass reported previously,[1] but contains approximately one cationic mole per cent of holmium. An oscilloscopic trace of the stimulated emission from the holmium activated glass etalon as detected by a cooled InSb photovoltaic detector is shown in Fig. 1. In this experiment, a germanium filter was placed between detector and laser rod. From a knowledge of the filter's spectral transmission, it is known that the stimulated emission here occurs at wavelengths greater than 1.95 μ. It should be noted that strong well-defined bursts of stimulated emission commence about 350 μsec after the peak of the excitation pulse; this time delay is not surprising since the spontaneous lifetime, of this glass at liquid nitrogen temperature has been measured to be 700 ± 200 μsec. The shape of these spikes indicate that the glass matrix is not of high optical quality; this observation is confirmed by direct visual examination of the etalon. It is estimated that for the trace in Fig. 1, the etalon is being excited approximately 5 to 10 per cent over threshold which is about 3600 joules using a rather inefficient helical flash tube excitation method. With better optical coupling of a standard type, it is expected that the threshold would be at least an order of magnitude less.

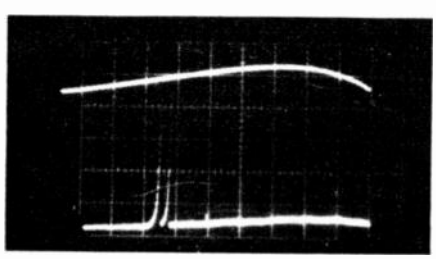


Fig. 1—Stimulated emission from holmium in silicate glass at a temperature of 78°K; excitation trace at top, laser beam trace at bottom. Writing speed is 100 μsec/cm; I_nS_5 detector with germanium filters in place.

Room temperature optical absorption measurements indicate strong absorption maxima at 3615A and 4490A with absorption coefficients of 6 and 8.5 cm^{-1}, respectively; the 4490A absorption band half width is about 200A. It is believed that the principal optical pumping spectral region is from 4400 to 4600A from the similarity of the glass absorption spectrum with that of the $CaWO_4$. From the nature of the absorption spectrum and the method of preparation, it is believed that the holmium in the glass is trivalent. Following Johnson,[2] it is supposed that the stimulated emission results from the $^5I_7$ to $^5I_8$ transitions in the $4\ f^{10}$ electronic configuration of the ion.

It should be noted that holmium is only the third ion which has been reported to exhibit laser action in a glass of any type. The first ion reported was trivalent neodymium in barium crown glass fibers[3] and the second was ytterbium in $LiMgAlSiO_3$ base glass rods of the same composition used in this work.[1] Work is now proceeding in order to improve our understanding of the stimulated emission of holmium in this glass.

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[2] L. F. Johnson, G. D. Boyd, and K. Nassau, "Optical maser characteristics of Ho^{+3} in $CaWO_4$," Proc. IRE, vol. 50, pp. 87–88; January, 1962.

[3] E. Snitzer, "Optical maser action of Nd^{+3} in a Barium crown glass," *Phys. Rev. Lett.*, vol. 7, pp. 444–446; December, 1961.

Simultaneous Laser Action of Neodymium and Ytterbium Ions in Silicate Glass*

Stimulated emission of radiation has been reported for three different ions in glasses. The first reported was neodymium in barium crown glass fibers,[1] the second was ytterbium in $LiMgAlSiO_3$ glass rods,[2] and the third was holmium in $LiMgAlSiO_3$ glass rods.[3] Laser action from neodymium had previously been reported for several matrices, the first being single crystal $CaWO_4$.[4] Trivalent neodymium represents a particularly favorable ion for laser action because the terminal energy level is essentially unpopulated at room temperature, and population inversion of the emitting energy level in this ion is relatively easy. The stimulated emission in the $4\ f^{13}$ electronic configuration of trivalent ytterbium in silicate glass has been ascribed to transitions between the $^2F_{5/2}$ and $^2F_{7/2}$ multiplets.[2] The present work describes the investigations of a glass in which both neodymium and ytterbium are incorporated.

Stimulated emission has been observed simultaneously from two different ions contained in the same $LiMgAlSiO_3$ glass etalon while being operated at liquid nitrogen temperature. This base glass was activated with one cationic mole per cent neodymium and one cationic mole per cent ytterbium and from their optical absorption spectra both are believed to be predominantly in the trivalent state. The absorption spectra for this glass appears to be a mere superposition of the spectra of the two ions when they are separately incorporated in this base glass composition; this result suggests that there is no strong interaction between the two incorporated rare earth ion species. It is possible, however, that an interaction between the two different ions may not be detected most sensitively by optical absorption.

An oscilloscopic record of the stimulated emission from the doubly activated glass etalon is shown in Fig. 1. Stimulated emission commences about 220 μsec after initiation of the flash excitation and occurs near the peak of the excitation pulse. This result suggests that the lifetime for the species causing the initial laser action is on the same order as the rise time of the exciting pulse. The spontaneous lifetime for singly activated Nd^{+3} in this glass is about 350 μsec and this ion is believed responsible for the initial stimulated emission observed. Subsequent experiments with a monochromator placed between laser and detector and set at the Nd^{+3} stimulated emission wavelength (1.06 μ) confirm this conclusion. In Fig. 1, heavy spiking occurs for Nd and lasts for almost 1200 μsec because the etalon is being excited approximately 350 per cent over the Nd threshold in this glass.

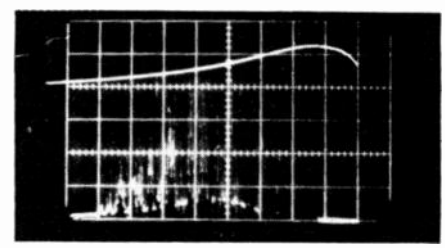


Fig. 1—Stimulated emission from neodymium and ytterbium in silicate glass at 78°K; excitation trace at top, laser trace at bottom. Writing speed is 200 μsec/cm. Sharp cut filter in place passing radiation of wavelengths greater than 1.0 μ.

In Fig. 1, an envelope commences to rise about 600 μsec after lamp flash initiation, and lasts for some 1000 μsec; the envelope peaks about 700 μsec after the maximum of the excitation pulse. This delayed emission is about what is observed in the singly activated ytterbium glass; there the lifetime for spontaneous emission is about 1450 μsec. Monochromatic investigation of this component of the emission in Fig. 1 indicates that it takes place at a wavelength of 1.015 μ and hence is due to ytterbium. The type of trace observed for the ytterbium component in this glass is the envelope as shown with spikes superimposed thereon; this emission pattern is characteristic of a fair optical quality glass etalon which is being excited just over threshold.[5] The trace in Fig. 1 is that obtained when the etalon is excited approximately 10 per cent over the ytterbium threshold. During the ytterbium (stimulated emission) period, the stimulated emission from the neodymium still persists and may be seen in Fig. 1.

Another effect observed in the present work is an apparent "fatigue effect" in the laser output from neodymium. It was found that when the etalon was excited at lamp

* Received August 2, 1962.

[1] E. Snitzer, "Optical maser action of Nd in a barium crown glass," *Phys. Rev. Lett.*, vol. 7, pp. 444–446; December, 1961.

[2] H. W. Gandy, H. W. Etzel, and R. J. Ginther, "Stimulated Emission of Infrared Radiation from Ytterbium Activated Silicate Glass," NRL Progr. Repts., pp. 27–28; February, 1962.

[3] H. W. Gandy and R. J. Ginther, this issue, pp. 2113–2114.

[4] L. F. Johnson and K. Nassau, "Infrared fluorescence and stimulated emission of Nd +3 in $CaWO_4$," Proc. IRE, vol. 49, pp. 1704–1706; November, 1961.

[5] The interpretation of this type of oscilloscope trace and its cause will be the subject of a future report.

input energies several times the neodymium threshold, the neodymium stimulated emission decreased and finally disappeared at this temperature, whereas the ytterbium emission did not. A study of this effect will be reported in the near future.

It is believed that this is the first reported work in which two different ion species have simultaneously exhibited laser action in the same host material.

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Direct Measurement of Optical Cavity Q*

The direct measurement of the bandwidth of an optical cavity resonator, in this case that of our unexcited Mark II gas maser, was undertaken as a part of a larger program involving coupled masers. The Mark II optical cavity is dimensionally identical to that of the Mark I maser furnishing the frequency-swept probe signal, both cavities being one meter long and being bounded by nominally 99 per cent reflecting plane parallel mirrors. Preliminary experiments showed the longitudinal mode frequency spacings between the two cavities to differ by no more than 50 kc out of the basic 150 Mc spacing. This gave assurance that the expected bandwidth measurements (0.3–1.2 Mc) would not be greatly perturbed by a vernier-type line broadening due to oscillation in more than one longitudinal mode by the probe maser. The same order of error arose through microphonically-induced jitter of the probe signal as well as the responding cavity.

Light waves being unguided, it was necessary in this experiment to physically align Mark I and II to extremely close angular tolerances. That is, the angular acceptance cone of Mark II as a receiving device is no doubt closely related to its beamwidth as a maser oscillator, in the order of 30 sec. Optimum transmission through Mark II, however, would probably occur only for some central portion of the cone, say 5 sec, and therefore the problem was to find and fix the optical axes of the two masers to within this angle.

An end-to-end configuration of the two masers was used for this purpose with the ultimate alignment being done by either using the mechanical hysteresis of the support table or by using the principle of a high compliance spring working against a low compliance one to obtain the fine motions required. The alignment indicator was taken to be the freedom of the detected Q-curve-like signals from off-axis modes.

Gross frequency alignment of the longitudinal modes between masers was accomplished by utilizing controllable direct current energization of a set of magnetostrictive tuning coils, one of which was wound on each of the four main Invar frame rods[1] of the Mark I maser. The tuning rate can be given roughly as 150 Mc/600 ampere-turns/coil. The oscillating maser was also swept over about a two megacycle frequency interval at 60 cps through the application of additional ac current to these same coils.

Frequency calibration of the Mark I maser was done by observing the photomixed beat-note on a spectrum analyzer between the fixed and oscillating Mark II laser and Mark I, when the latter was tuned either in a point-by-point manner with direct current or by its 60∼sweep. Fig. 1 illustrates the usual saddle-shaped power spectrum obtained with 60∼sweep, where the definition has been somewhat impaired by microphonic jitter. We found that the chosen sweep width was a good compromise between an upper limit determined by the detector response time and a lower limit set by mechanical and thermal stability considerations.

With adequate spatial and frequency alignment between the oscillating and receiving masers, the frequency modulated probe signal passed through Mark II and traced out its transmission characteristics via an Ektron detector and an oscilloscope. Fig. 2 affords an example of the best appearing Q curves. From this and pictures like it we were able to determine that the cold cavity bandwidth under optimum conditions was about 0.3 Mc. More often the bandwidth was measured to be about 0.35–0.4 Mc. The passive Q, incidentally, is very nearly one billion.

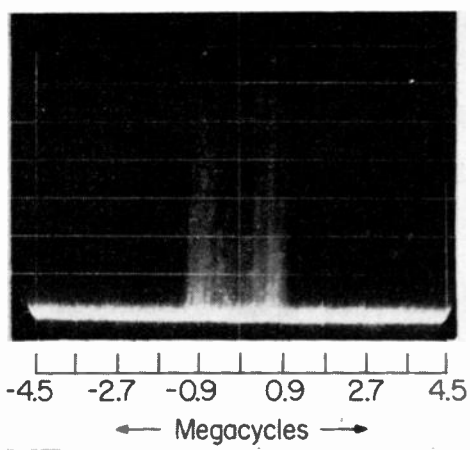


Fig. 1—Power spectrum of frequency-swept maser.

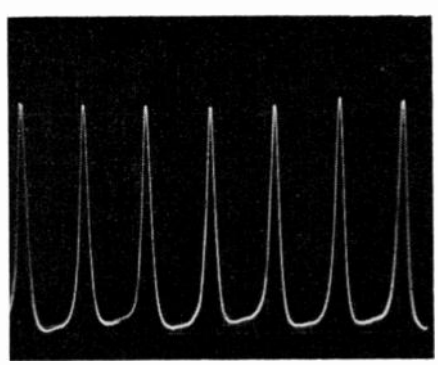


Fig. 2—Q-curves representative of the best results.

If we neglect diffraction losses, the mirror reflectance coefficient calculated from the above bandwidth value turns out to be 99.4 per cent. One has in this technique an elegant method of determining mirror reflectance coefficients in the difficult >99 per cent region.

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Resonant Amplification in a Gas Maser*

The optical gain and the manner in which the gain varies with input signal level were measured on our Mark II gas maser operated at selected excitation levels up to the edge of oscillation point. In this mode of usage the Mark II amplifier could be expected to exhibit, along with gain, the special features of a resonant device, namely an extremely narrow amplifying bandwidth, and in this case, extremely narrow spatial acceptance angles.

A few general remarks are in order concerning the possible varieties of gas laser amplifiers. The central problem to be overcome in such devices is the very low specific gain γ in the equation, $P_{out} = P_{in}e^{\gamma z}$. For ordinary discharge tube diameters and pressures (of He and Ne), γ might be 0.05/m. To achieve a gain of 10, say, one would then need to traverse a gas column 46 m long. In the absence of breakthroughs towards higher γ values, all gas laser amplifiers with significant gains must, in one way or another, prove this order of column length.

The dictates of practicality suggest then that the varieties of gas laser amplifiers will be one-to-one with the ways in which one can fold an optical path.

Non-resonant gas maser amplifiers are perfectly feasible and in fact are the subject of current research. Here, however, we will only consider a feedback or resonant form, whose geometry takes advantage of the fact that there is no restriction that the same gas path cannot be traversed more than once. In return for this spatial economy, nature exacts a toll in the stringent relation between the length of the optical cavity W and frequency for any sort of transmission at all through the maser. The form of such a gas maser amplifier would be simply that of a gas maser oscillator excited below the start-oscillation level.

The on-resonance transmission through an amplifier like this is given by (1) below.

$$T_r = \frac{T^2 A_{s.p.}}{(1 - A_{s.p.}R)^2} \qquad (1)$$

* Received August 1, 1962.

[1] C. F. Luck, R. A. Paananen, and H. Statz, "Design of a helium-neon gaseous optical maser," Proc. IRE (*Correspondence*), vol. 49, pp. 1954–1955; December, 1961.

* Received August 1, 1962.

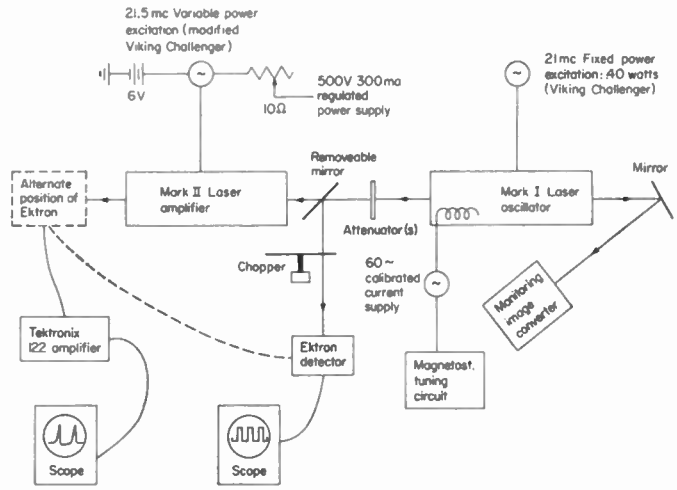


Fig. 1.

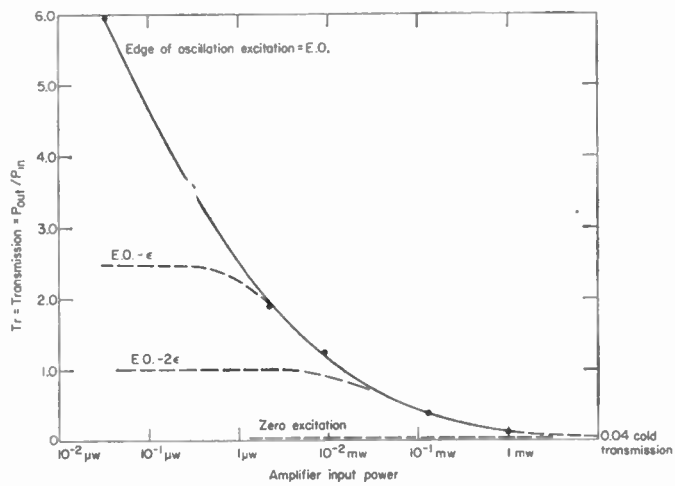


Fig. 2.

where the values in a specific case might be

$$T = 2 \cdot 10^{-3}$$

(power transmission coefficient of one mirror)

$$R = 0.994$$

(power reflection coefficient)

$$\therefore A = 4 \cdot 10^{-3}$$

$(A+R+T=1)$

$$W \cong 1 \text{ m}$$

(max. excitable length about 4/5 m)

$$A_{s.p.} = 1.00 - 1.04$$

(single pass power gain, dependent upon excitation level)

$$T_r = 10$$

(for example).

Then,

$$(1 - A_{s.p.}R) = T\sqrt{\frac{A_{s.p.}}{T_r}} \cong 0.6 \cdot 10^{-3} < 0.1 \text{ per cent.}$$

That is, the $A_{s.p.}$ value must be set and maintained to within less than 0.1 per cent of the start oscillation value to achieve a gain of 10 with the chosen mirror coefficients.

Some details of the inter-maser space and frequency alignment procedures necessary to perform coupled maser experiments are given in a companion letter (Direct Measurement of Optical Cavity Q). Fig. 1 illustrates the particular instrumentation found suitable by us for measurements of optical gain and gain saturation.

Most of the essential results of this work are reproduced as the curves of Fig. 2. The solid line represents the exact edge-of-oscillation gain data for about a $3 \cdot 10^4$ power input range. Achieving gains greater than 6 was not feasible due to the instability of the regeneration calculated earlier. The dotted curves are indicative of the gain behavior for excitations not at the E.O. point, culminating, of course, in the flat cold transmission, here 4 per cent.

A complete explanation of the gain saturation behavior in Fig. 2 has not yet been derived. Eq. (2) below gives, without any theoretical justification, a reasonable fit to the E.O. curve.

$$P_{out} = 0.04P_{in} + 0.085\sqrt{P_{in}} \quad (P_{in} \text{ in mw}). \quad (2)$$

Much more satisfactory gain figures are predicted by (1) if mirrors designed for amplifier, rather than oscillator, service are utilized. Gains of several hundred seem to be possible for R values close to 0.96. Work in this direction is continuing.

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Junction Transistor with Reverse-Biased Input*

It was shown earlier[1] that a junction transistor can be operated as a voltage amplifier with its input (base-emitter junction) reverse biased. Although, this results in some loss in voltage gain, still, the gain remains sufficiently above unity. With the transistor type CK721 (Raytheon), it was possible to have a voltage gain of 20 to 30 db. The reverse-biased operation gives the special advantages of high input impedance and reduced feedback.

With present-day transistors the signal handling capacity of the reverse-biased input for faithful amplification is rather low (about 100 mv), compared to the forward-biased case. But with some modifications in the transistor construction[1] the situation is expected to improve.

With forward biasing, the output (collector) current (I_c) is found to increase approximately linearly with input (base) current (I_b), rather than the input voltage (V_b). [Considering a grounded emitter configuration.] This justifies the use of current gain factor $\alpha_s(=\partial I_c/\partial I_b)$ as the relevant transfer function.[2] Because, under such operation α_s, rather than transconductance $g_s=(\partial I_c/\partial V_b)$, remains constant over a wide range. But the situation is completely different with reverse-biased operation. Here, the input (base) reverse current is negligibly small (a few microamperes) and does not show a linear relation with the output current (I_c). Under such an operation the output current I_c is found to vary linearly with input voltage V_b (Fig. 4 in Chatterjee[1]), rather than input current I_b. As such, the use of a transconductance $g_s(=\partial I_c/\partial V_b)$ seems more appropriate in this case. With reverse-bias operation, the current amplification factor α_s loses its significance and rather the transconductance (g_s) is to be used for calculating the amplifier gain. With the transistor type CK721 (Raytheon) and with a collector voltage of 20 v (negative), the transconductance was found to be about 230 μmhos. This value of transconductance compares favorably well with those in voltage amplifier vacuum triodes and reasonable voltage gain can be obtained. It may be

* Received, March 2, 1962.

[1] B. Chatterjee, "A transistor voltage amplifier with reverse-biased input," *J. Inst. Telecom. Eng.* (India), vol. 7, pp. 249–254; November, 1961.

[2] "Cathode Ray," *Wireless World*, vol. 67, pp. 490; September, 1961.

mentioned here that with the input reverse biased, a transistor can be safely operated with a higher collector voltage than specified for forward-biased operation.

It may not be irrelevant to mention that a transistor with its input reverse biased is primarily useful as a voltage amplifier—especially where a high input impedance is desirable. Because of their low output current, they are not capable of delivering the maximum available power. Power amplifiers may preferably be used with a forward bias—as in high power vacuum tube amplifiers, where the grids are often driven positive to extract the maximum available power.

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Measurement of Bulk-Lifetime by Microwave Electrodeless Technique with Steady Illumination*

Jacobs, *et al.*,[1] have suggested a method for the measurement of bulk-lifetime of carriers in semiconductors in which the sample in the form of a rod is inserted in a waveguide with its axis perpendicular to the broad face of the guide. The sample is illuminated at one end by a steady source of light. Change in absorption of microwave on illumination for different values of the length d between the illuminated end-face and the point of entry within the guide is measured and the diffusion length is determined from the slope of the curve giving the logarithms of the change in absorption as a function of d. As the authors point out the method is fruitful provided 1) the distribution of excess carriers along the axial direction of the sample is exponential, 2) change in absorbed power is linearly related to the integrated excess carrier density inside the guide, and 3) $\sigma \ll \omega\epsilon$. It is further stated that the complication arising out of surface recombination may be avoided with this method. In actual practice assumption 2) may be satisfied, if, as stated by the authors, the level of illumination is low. Assumption 3) may be fulfilled through careful choice of material and ω. The validity of assumption 1) and the effect of surface recombination are, however, matters of more detailed discussion. In the present communication this aspect of the problem is discussed.

A sample that can be treated as one-dimensional is considered and the incident radiation is assumed to be monochromatic. Further, the sample is assumed to possess a thin surface layer where recombination occurs profusely. A part of the incident radiation flux is absorbed in the thin layer in a manner characteristic of the layer itself and the remainder penetrates inside the sample following a law characteristic of the bulk-absorption properties of the material. The steady-state continuity equation may, therefore, be written as

$$\frac{\partial^2 p}{\partial x^2} - \frac{p}{L^2} = -\frac{\alpha}{D}\eta G_1' e^{-\alpha x} = -\frac{\alpha}{D} G_1 e^{-\alpha x} \quad (1)$$

where p, L, D, α, η have the usual meaning. The quanta of light penetrating into the volume is G_1'.

The excess carrier density p should satisfy the boundary conditions

$$-Sp\big|_{x=0} + G_2 = -D\frac{\partial p}{\partial x}\bigg|_{x=0}; \quad p \to 0,\ x \to \alpha$$

where S=recombination velocity and G_2= carriers generated on the surface. On solving (1) the total number of carriers in the region of the sample inside the waveguide is found to be

$$p' = \frac{G_1\tau}{(\alpha^2L^2 - 1)}\left[\frac{D\alpha + S}{\frac{D}{L} + S}\alpha L(1 - e^{-b/L})e^{-d/L} - (1 - e^{-\alpha b})e^{-\alpha d}\right] + \frac{G_2 L}{\frac{D}{L} + S}(1 - e^{-b/L})e^{-d/L} \quad (2)$$

where b=smaller dimension of the waveguide.

It would be convenient to consider separately three special cases.

Case I

$$\frac{1}{L} \ll \alpha, \quad i.e., \quad \alpha L \ll 1.$$

For this case the term containing $e^{-\alpha d}$ may be neglected and the slope of the ln p' vs d curve would give L.

Case II

$$\frac{1}{L} \gg \alpha, \quad i.e., \quad \alpha L \ll 1.$$

Under this condition the first and the third term may be neglected compared to the second. The final slope of *ln* p' vs d curve, in this case, gives the absorption constant α.

From the available experimental data[2] it would appear that G_e with visible irradiation would favor Case I, whereas G_e with infrared radiation would favor Case II.

Case III

$$\frac{1}{L} \doteq \alpha, \quad i.e., \quad \alpha L \doteq 1.$$

A linear relation may be obtained in this case only for high values of d. The final slope of the *ln* p' vs d curve might give either α or $(1/L)$, whichever is smaller. However, for large values of S and for $\alpha L \doteq 1$, the plot of *ln* p' as a function of d would have zero slope for values of d comparable to L. Further, in some cases, the curve might have a maximum, in which its position and the maximum value of normalized p' may be used to obtain α and L using (2). Alternatively, if the maximum is not realized, α and L may be determined by comparing the values of p' for $d = L$ and $d = 2L$.

It is to be noted that the curves of Jacobs, *et al.*,[1] show in one case a zero slope and in the other a maximum. They ascribe this deviation from the expected linear plot to experimental errors. The above discussion offers an alternative explanation if it is assumed that αL is of the order of unity in the experimental condition. One may then determine α and L as discussed above. Assuming that the final slope gives $1/L$, the values of α obtained in both the cases are found to be 4. On the other hand, if it is assumed that the final slope gives α, the value of lifetime comes out to be about 1 msec, which agrees reasonably well with those given by the photo-conductive decay method. In both the cases the product αL is found to be closely equal to unity. It must, however, be remembered that the experimental data were probably obtained with a composite beam of light. The α value as found from the above analysis should therefore be regarded as an effective value.

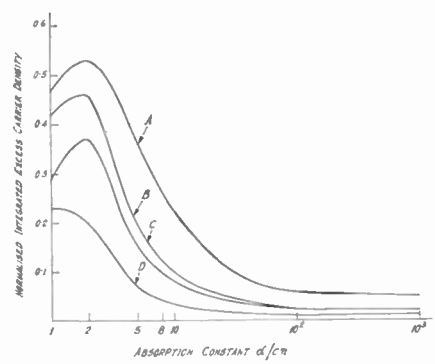


Fig. 1—Integrated excess carrier density for different values of the absorption constant. A—L =0.2 cm, $d = L$, B—L =0.2 cm, $d = 2L$, C—L =0.5 cm, $d = L$, D—L =0.5 cm, $d = 2L$.

The rather low values of α as found above is also corroborated by Fig. 1, which gives plots of normalized p' as a function of α for $d = L$ and $d = 2L$ with $S = 2000$; $b = 1$ cm, $L = 0.5$ and 0.2 cm. It is found that p' attains a maximum for values of α in the range 1–10. The available data on G_e show that α values in the range 1–10 are obtained with infrared radiation of wavelength 1–2 μ[2]. It may be noted that for an ordinary tungsten lamp, radiation of this range of wavelength would constitute about 30 per cent of that in the visible range for which α values are much higher.[3] If such a lamp is used as the source of irradiation then one can easily show, with the help of Fig. 1, that the analysis of experimental data would yield a low value of α and a value of αL roughly equal to unity.

In the light of the above discussions one may conclude that the method of Jacobs, *et al.*,[1] gives L correctly irrespective of the value of S if $\alpha L \gg 1$. If, however, $\alpha L \doteq 1$, the data may be used to obtain α and L if S is

* Received February 5, 1962.

[1] H. Jacobs, A. P. Ramsa and F. A. Brand, "Further consideration of bulk lifetime measurement with a microwave electrodeless technique," PROC. IRE, vol. 48, pp. 229–233; February, 1960.

[2] M. L. Schultz and G. A. Morton, "Photoconduction in germanium and silicon," PROC. IRE, vol. 43, pp. 1819–1828; December, 1955.

[3] V. K. Zworykin and E. G. Ramberg, "Photoelectricity and its Application," John Wiley and Sons, Inc., New York, N. Y., p. 18; 1950.

large. On the other hand, if $\alpha L \ll 1$ the method does not give L, but α.

The authors would like to express their sincere thanks to Prof. J. N. Bhar for his kind interest in the work.

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Measurements of Probability Densities of Small Ensembles of Periodic Waveforms*

We wish to demonstrate with measured data some interesting properties of finite ensembles which have not to our knowledge been explicitly illustrated in the literature. With but very few exceptions[1,2] the statistics of either large ensembles (where the central limit theorem is applicable) or of single periodic waveforms are discussed.

Presented here is a set of measurements that were obtained for the summation of two and three periodic waveforms with various energy ratios.[3] These measurements have been made with advanced instrumentation techniques[4] which yield a high degree of accuracy.

Fig. 1 presents experimental measurements of the probability density for the sum of two independent sine waves as the energy ratio is varied. Curve (a) is a single sine wave and is represented by an energy ratio of 0:1. Curve (b) illustrates the large effect on the probability density of a small perturbation in the form of an additional waveform of much lower energy. In the equal energy case (d) the characteristic form is observed as reported by Slack.[1]

Fig. 2 presents density curves for the summation of three independent cosine waves where the energy of one wave is greater than or equal to the others. This illustrates an interesting concept in probability theory. The rate of convergence to a Gaussian type of distribution is influenced by the energy ratio. The convergence is most rapid when the energies are equal.[5]

Fig. 3 indicates the density function resulting when a square wave and an independent sine wave are combined (equal energy). Four modes dominate the function. These are at the maxima and minima of the resulting waveform. But yet when a third independent wave of equal energy with the first two is added, the strong modal structure disappears. Fig. 4 indicates this effect when an equal-energy triangular wave is added. Comparison with the probability density of the summation of three equal cosine waves suggests that the convergence to Gaussian may be more rapid when the summation includes nonidentical waveforms.

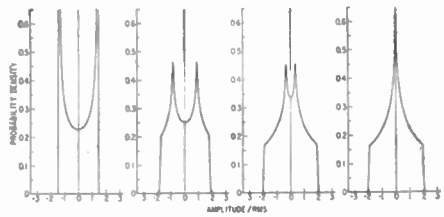


Fig. 1—Probability densities representing the sum of two independent cosine waves for various energy ratios. (a) 0:1. (b) $\frac{1}{2}$:1. (c) $\frac{1}{\sqrt{2}}$:1. (d) 1:1.

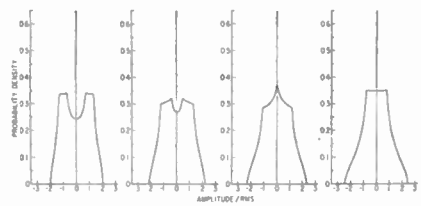


Fig. 2—Probability densities representing the sum of three independent cosine waves for various energy ratios. (a) $\frac{1}{\sqrt{2}}:\frac{1}{\sqrt{2}}$:1. (b) $\frac{1}{2}:\frac{1}{2}$:1. (c) $\frac{1}{2}:\frac{1}{2}$:1. (d) 1:1:1.

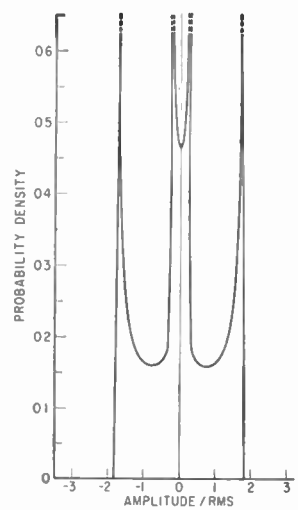


Fig. 3—Probability density of the sum of independent square and sine waves for equal energy ratios.

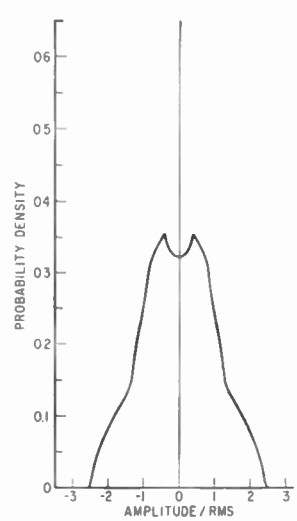


Fig. 4—Probability density of the sum of independent square, cosine, and triangular waves for equal energy.

The data described herein serves to indicate the wide variety of probability density functions that exist between the realms of single periodic waveforms and the summation of very large numbers of these waveforms or Gaussian distributions.

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Temperature Sensitivity of Twoport*

The use of a noise figure in describing the noisiness of a twoport becomes less significant in low-noise devices which are operated at temperatures much below room temperature. For this reason the "effective input noise temperature" was introduced.[1]

In a recent communication,[2] Fisher defines the "excess noise temperature" T_r of a twoport by

$$T_r = \frac{\Delta P_{eoi}}{kA_e\Delta_f}, \tag{1}$$

where A_e is the exchangeable power gain and ΔP_{eoi} is the exchangeable power output due only to the noise sources internal to the twoport. According to the definition of the "effective input noise temperature," it is the temperature of an input termination which, when it is connected to the noise-free twoport, would result in the same noise power as that of the actual twoport connected to a noise-free input termination. Hence the effective input noise temperature T_e is given by

$$T_e = \frac{\Delta P_{eoi}}{kA_e\Delta_f} \tag{2}$$

yielding the identical result as (1). Thus

$$T_r = T_e. \tag{3}$$

Similarly the "temperature sensitivity" T_s of a twoport with a particular termination having an "equivalent noise temperature" T_{eq} can be expressed, using the input noise temperature, by

$$T_{et} = T_e + T_{eq} \cong T_s \tag{4}$$

where T_{et} is a total equivalent input noise temperature. Writing T_{eq} in terms of noise

* Received February 7, 1962; revised manuscript received, February 23, 1962.

[1] M. Slack, "The probability distribution of sinusoidal oscillations combined in random phase," *J. IEE*, vol. 93, pt. 3, pp. 76–86; March, 1945.

[2] T. Sheridan, "Time-Variable Dynamics of Human Operator Systems," AFCRC-TN-60-169; 1960.

[3] H. L. Fox and E. A. Starr, "The statistics of finite ensembles," *J. ASA*, vol. 33, p. 1679; November, 1961.

[4] The measurements were made using a B+K Instruments Probability Density Analyzer Model 160.

[5] This is the "weak" or "equal-component" central limit theorem vs Liapounov's theorem; see M. Loeve, "Probability Theory," D. Van Nostrand Co., Inc., New York, N. Y.; 1955.

* Received February 23, 1962.

[1] "IRE standards on methods of measuring noise in linear twoports, 1959," Proc. IRE, vol. 48, pp. 60–68; January, 1960.

[2] S. T. Fisher, "A completely consistent definition of temperature sensitivity including negative conductance devices," Proc. IRE (*Correspondence*), vol. 50, p. 204; February, 1962.

ratio n, which may be negative, we obtain

$$T_{eq} = nT, \tag{5}$$

where T is the room temperature in degrees K. Hence (4) has the same significance as the "effective noise figure" which includes the effect of the noisiness of the input termination with its noise ratio other than unity.

For a cascaded twoport we have

$$T_{e12} = \frac{kT_{e1}A_{e1}A_{e2}\Delta_f + kT_{e2}A_{e2}\Delta_f}{kA_{e1}A_{e2}\Delta_f}$$

$$= T_{e1} + \frac{T_{e2}}{A_{e1}}. \tag{6}$$

This is in complete agreement with Fisher's (9).

As to the heterodyne receiver, an "effective input noise temperature is defined for each pair of corresponding frequencies" according to Note 1 of the IRE Standards on Noise[1] which again is in complete agreement with the advantages pointed out by Fisher.

It seems, therefore, that the "temperature sensitivity" is a redundant term.

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Author's Comments[3]

I would like to state to Mr. Shimada and to IRE readers in general my purposes in submitting my original communication. They were:

1) To show that noise temperature and temperature sensitivity (or system noise temperature) can be uniquely defined without regard to whether source conductances (or resistances) are positive or negative;
2) To show, in particular, that a negative noise temperature is meaningful;
3) To point out the advantages of using the temperature-sensitivity concept rather than the noise-figure concept.

In many cases the symbol T_r, defined in my communication, and T_e, defined by the Standards,[1] may be used interchangeably. However, T_r, as I defined it, has some very important properties which are not expressly set forth in the Standards definition of T_e. These are as follows:

1) T_r is defined in terms of exchangeable power and exchangeable gain. The concepts of the noise temperature of the input termination and of noise-free networks are not required in the definition of T_r, although they may be helpful in understanding the physical significance of T_r.
2) T_r (also T_{eq}, the noise temperature of the input termination) is negative when, and only when, the source conductance (input termination) is negative.
3) All terms of the "cascade formula" have the same sign, which is the same as the sign of the source conductance of the *first* network in the cascade. This is independent of the signs of the output conductances of any of the networks in the cascade.
4) The excess noise temperature T_r is defined in terms of noise originating in the network itself. It does *not* include noise from the source at image frequencies in the case of a heterodyne receiver. In contrast to this, Note 3 of Section 5 of the Standards states that T_e is related to noise figure by

$$T_e = 290(F - 1).$$

If T_e is to include only noise originating in the network itself, then this relationship is valid *only* for the twoport (single-input-channel) case. It cannot be extended to the heterodyne case in which there are inputs at image frequencies.

Although T_e can be shown to have properties 1)–3) of the preceding paragraph, it being assumed that the noise temperature of the input termination has been properly defined for the case of negative conductance, it is my opinion that any future Standards article ought to state specifically the significance of the noise temperature of the source and the equivalent input noise temperature of the twoport when the source conductance is negative.

For the single-input-channel case it is true that

$$T_e = T_r = 290(F - 1),$$

where F is the noise figure. However, if the equation $T_e = 290(F-1)$ is used for a multiple-input-channel receiver, then T_e includes noise from the source at image frequencies. This follows from the definition of F. On the other hand T_r is defined to represent noise originating only in the network itself whether the latter is a twoport or a multiple-input-channel receiver.

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[3] Received March 15, 1962.

Self-Consistent Electron Trajectories in a Magnetron*

The electron trajectories obtained numerically by Tibbs and Wright[1] are generally accepted[2] as being typical of an oscillating magnetron. However, Walker[3] has shown that these trajectories are not self-consistent.

We programed the problem of Tibbs and Wright for an IBM 704 computer. Like Tibbs and Wright, we used Hartree's method of self-consistent fields. Tibbs and Wright had performed only one cycle of the computation. We carried the computation through 36 cycles. Even so, our trajectories did not converge toward a self-consistent result.

When we dropped the requirement of space-charge-limited cathode emission and changed the values of the independent variables for a better correspondence with experimental data, we quickly obtained a self-consistent solution. A composite plot of the self-consistent trajectories is given in Fig. 1. They were computed for the CV 76 (modern designation, CV 1481). This is an S-band, 500-kw magnetron. It has 8 vanes and operates in the π mode. The calculations were made for the following conditions:

Frequency: 3000 Mc
Magnetic field: 0.23 webers/m^2
Peak anode current: 30 a
Peak anode voltage: 26.25 kv
Amplitude of traveling voltage wave on anode: 12 kv

The anode voltage and current were measured values and the RF voltage corresponded to maximum efficiency for a zero-space-charge calculation.

The computed efficiency came out to be 50 per cent, a value which agrees reasonably well with the measured value of 57 per cent. The calculated back-bombardment power was 6 per cent of the RF output power. About half the current emitted from the cathode went to the anode; the other half returned to the cathode.

Fig. 2 shows a plot of the space-charge density (averaged over the angular coor-

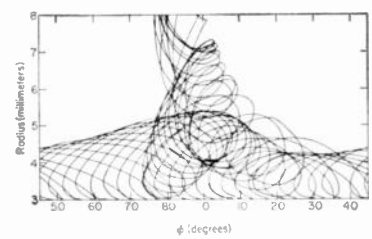


Fig. 1—Self consistent trajectories.

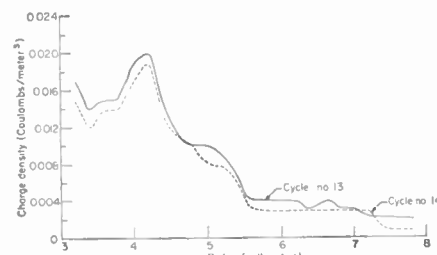


Fig. 2—Space-charge density (averaged over the angular coordinate).

* Received February 23, 1962. This work was supported by BuShips under contract no. NObsr-77592.

[1] S. R. Tibbs and F. I. Wright, "A Self-Consistent Field and the Space Charge Configuration for the CV76," CVD Report Mag. 41; March 23, 1945.

[2] See for example: G. B. Collins, Ed., "Microwave Magnetrons," McGraw-Hill Book Co., Inc., New York, N. Y., p. 270; 1948.
R. Latham, A. H. King and L. Rushforth, "The Magnetron," Chapman and Hall, London, England, p. 62; 1952.
S. I. Bychkov, "Magnetron Transmitters," Naval Publishing House, Defense Ministry of the USSR, Moscow, p. 37; 1955.

[3] G. B. Collins, Ed., "Microwave Magnetrons," McGraw-Hill Book Co., Inc., New York, N. Y., pp. 274–282; 1948.

dinate ϕ) vs the anode radius. The charge distribution in which the trajectories of Fig. 1 were calculated is compared with the charge distribution derived from these trajectories. The figure shows the self-consistency of the charge distribution.

Figs. 1 and 2 show that the "hub" region of the space charge extends beyond the synchronous radius (3.4 mm). It goes out approximately to the Hull cutoff radius (4.4 mm). Within the hub, the space charge has more the character of a double-stream distribution (with maxima at the cathode and Hull radii) than of a Brillouin single-stream distribution.

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Comments on "Analysis of AGC Loops"*

Certain discrepancies in my earlier correspondence[1] have been brought to my attention. Namely, although (9) is a solution of the differential equation (8), it is not a solution of (4), from which (8) was derived. The conclusion has been reached that (9) is not the complete solution of (8); in fact, it is suspected that (9) is a singular solution.

Efforts have been made to find a complete solution, so far without success. It should not be surprising to learn that the solution is not possible in explicit form. Communications from other interested persons would be welcome.

Despite the fact that (9) is incomplete (or singular), it is a good approximation to the error voltage, particularly if loop gain is large. This may be seen by manipulating (4) to obtain

$$E = -\frac{20}{KA}\log\frac{x}{x_0 + \dfrac{E}{G_0}}, \qquad (4a)$$

which approaches (9) for E/G_0 sufficiently small.

The material following (9) is correct, not being dependent upon the solution of the differential equation.

The author is indebted to D. Rothman of Electronic Specialty Co. for bringing the discrepancy and (4a) to his attention.

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* Received February 27, 1962.
[1] Floyd M. Gardner, "Analysis of AGC loops," PROC. IRE (*Correspondence*), vol. 50, p. 97; January, 1962.

The Maximum Power Output of the Tunnel-Diode Oscillator*

The amplitude and power output of a tunnel-diode oscillator may be discussed on the base of approximate third order polynomial characteristics, derived by van der Pol,[1]

$$I - I_0 = a(U - U_0) + b(U - U_0)^3 \qquad (1)$$

Although this formal approach in many respects truly describes the oscillation, certain properties of the oscillator, for instance the discontinuous decrease and the abrupt vanishing of the oscillation by the change of the load or the supply voltage, remain unclarified. In a previous article[2] the characteristics of a tunnel-diode were derived by quantum mechanical reasoning, as

$$I = a(U - U_m)^2 \tanh\frac{1}{2}\frac{U}{U_T} \qquad (2)$$

which approximates the real characteristics more accurately than former expressions. When treating oscillations, with regard to (2), the following approximations are appropriate:

$$I = \begin{cases} g_0 U & U < U_M \\ \dfrac{\gamma}{2}(U - U_m)^2 + I_m & U > U_M \end{cases} \qquad (3)$$

The oscillation is determined by the conductance averaged over one cycle.[1] If the full cycle runs in the interval approximated by quadratic function (Fig. 1) then the average conductance is

$$G_{av} = \frac{1}{2\pi}\int_0^{2\pi} G(\omega t)d(\omega t) = \gamma(U_d - U_m). \qquad (4)$$

This conductance compensates for the damping conductance G_L of the oscillator circuit in a stationary state, that is

$$G_L + G_{av} = 0. \qquad (5)$$

In such a case the amplitude of the oscillation increases up to the limit of the quadratic interval ($U_1 = U_d - U_M$). It is obvious from (4) and (5) that by increasing the forward voltage at a given G_L load, the oscillation starts at the value U_M and stops at the voltage $U_d = U_M - (G_L/\gamma)$, when the amplitude of the oscillation is

$$U_1 = U_d - U_M = U_d - U_m + \Delta U = \Delta U\left(1 + \frac{G_L}{\gamma\Delta U}\right). \qquad (6)$$

Let us now consider the commencement of the oscillation when the operating point is at the start of the quadratic section (at U_M). In this case the averaged conductance of the tunnel-diode is

$$G_{av} = \frac{1}{2\pi}\int_0^{2\pi} G(\omega t)d(\omega t) = \frac{\gamma}{2}[U_1 + (U_d - U_m) + \Delta U]. \qquad (7)$$

From this the amplitude of the oscillation at this operating point is

$$U_1 = \frac{\pi}{2}\Delta U\left(1 - \frac{g_0}{\gamma\Delta U} - 2\frac{G_L}{\gamma\Delta U}\right). \qquad (8)$$

The calculated and measured characteristics of the oscillation are plotted in Fig. 2.

The maximum power output of the tunnel-diode oscillator is obtainable near the ceasing of the oscillation, *i.e.*,

$$P_{max} = \frac{1}{2}U_1^2 G_L = \frac{1}{2}\Delta U^2\left(1 - \frac{G_L}{\gamma\Delta U}\right)^2 G_L. \qquad (9)$$

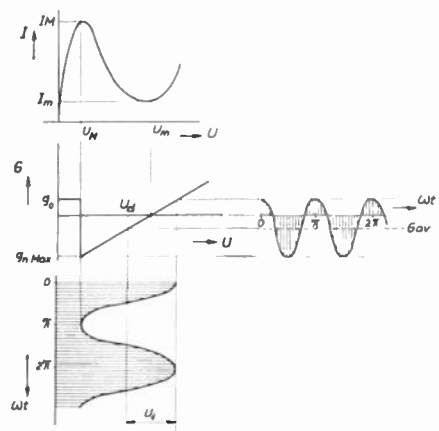
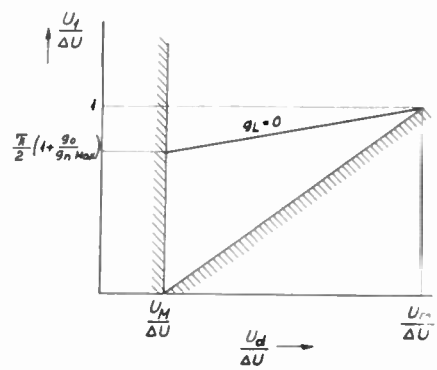


Fig. 1—The average conductance of the tunnel-diode. The full cycle runs in the quadratic interval.



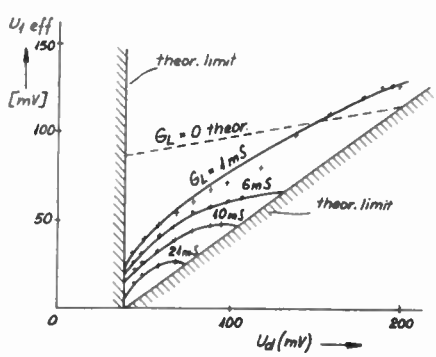


Fig. 2—Oscillation characteristics of the tunnel-diode oscillator. (The amplitude of the oscillation vs operating point dc voltage.) (a) Theoretical limits. (b) Measured characteristics. (Parameter: load conductance.)

* Received February 26, 1962; revised manuscript received March 12, 1962. The work reported in this paper was performed at the Central Measurement Research Laboratory, Budapest, Hungary.
[1] K. Tarnay, "The Tunnel-Diode," Ph.D. dissertation, Technical University, Budapest, Hungary; 1961.
[2] K. Tarnay, "Approximation of the tunnel-diode characteristics," PROC. IRE (*Correspondence*), vol. 50, pp. 202–203; February, 1962.

Assuming that $\gamma\Delta U = g_{n\,max}$ and referring to (3), ΔI is

$$\Delta I = I_M - I_m = g_{n\,max}\cdot\frac{U_M + U_M}{2}\cdot \quad (10)$$

A combination of these equations yields the maximum power output

$$P_{max} = \Delta U \Delta I \frac{U_m - U_M}{U_m + U_M}\left(1 - \frac{G_L}{g_{n\,max}}\right)^2 \cdot \frac{G_L}{g_{n\,max}} \quad (9a)$$

The optimum power output of the oscillator may be obtained by calculating the extremum of (9a) in the function of the load conductance

$$P_{opt} = \frac{4}{27}\,\Delta U \Delta I\,\frac{U_m - U_M}{U_m + U_M}\cdot \quad (11)$$

The power output is optimal at $G_L = -\frac{1}{3}g_{n\,max}$ and at $U_d = U_m - \frac{1}{3}\Delta U$ voltage. Assuming $U_M = 0.2\ U_m$ as usual, with customary tunnel-diodes, the optimum power output is

$$P_{opt} = \frac{8}{81}\,\Delta U \Delta I. \quad (11a)$$

This is only the 52.2 per cent of the value calculated by the formal third order approximation ($3/16\Delta U\Delta I$). When the load conductance is not the optimum, the maximum output power is

$$P_{max} = P_{opt}\left[\frac{27}{4}\left(1 - \frac{G_L}{g_{n\,max}}\right)^2\frac{G_L}{g_{n\,max}}\right] \quad (9b)$$

The power output vs the load conductance is shown in Fig. 3.

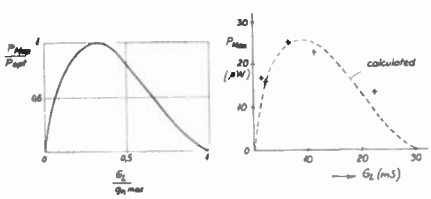


Fig. 3—Maximum output power of the tunnel-diode oscillator. (a) Theoretical. (b) Measured.

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Noise Measurements on Tunnel Diodes*

Since the initial studies of Yajima and Esaki[1] in 1958, relatively few papers reporting basic noise measurements in tunnel diodes have been published. Montgomery and Lee[2] reported on noise measurements at 5 Mc, intended to split the total tunnel current in its forward and reverse components. J. J. Tieman reported measurements taken by R. L. Watters of G.E. Research Laboratories.[3]

The present note will report noise spectra between 100 kc and 30 Mc on several diodes, for currents in the low-voltage positive-conductance region, the excess-current region and the high-voltage positive-conductance region. The biasing circuit involved regular RC coupled circuits at the lower frequencies and tuned LC circuits above 0.55 Mc. No step-up transformers were employed to avoid possible errors. The noise of the input 6922 cascode circuit was subtracted. The noise was calibrated with a standard noise diode. In all cases the noise spectra measured in the lower positive conductance region, taken from 100 kc to 30 Mc, were completely flat. Some data for a ZJ56A diode are listed in Table I.

If the noise corresponds to full shot noise[4] of the individual tunneling currents—the Zener current and the Esaki current—one obtains from Fermi Dirac statistics the following simple result,[5] relating the equivalent saturated diode current I_{eq} to the tunnel diode current I_d:

$$I_{eq} = I_d \coth (eV/2kT). \quad (1)$$

The calculated magnitudes according to (1) are also included in Table I. The correspondence is quite good and does indicate full shot noise, *i.e.*, the tunneling of individual carriers is independent.

In Fig. 1 we show some noise spectra in the excess current and second positive-conductance region for the same diode. It is clear that the shot noise level is only reached for frequencies over 10 Mc. The large amount of LF noise is apparently associated with the excess current and with the normal *p-n* junction current in these regions. The current dependence is complex. The shot noise levels as a function of bias for this diode are also summarized in Fig. 2.

TABLE I
Experimental Data for I_{eq} in the Lower Positive Region of ZJ56A Tunnel Diode

I_d = 0.9 ma		I_d = 0.8 ma		I_d = 0.7 ma		I_d = 0.6 ma		I_d = 0.5 ma	
Freq. Mc	I_{eq} ma	Freq. Mc	I_{eq} ma	Freq. Mc	I_{eq} ma	Freq. Mc	I_{eq} ma	Freq. Mc	I_{eq} ma
0.55	1.27	0.55	1.37	0.55	1.42	0.55	1.51	0.55	1.53
1	1.31	1	1.32	1	1.44	1	1.50	1	1.53
2	1.26	2	1.33	2	1.39	2	1.49	2	1.49
4	1.31	4	1.28	4	1.40	4	1.40	4	—
8	1.30	8	1.33	8	1.41	8	1.48	8	—
15	1.26	15	1.32	15	—	15	—	15	—
30	1.26	30	—	30	—	30	—	30	—
$\overline{I_{eq}} = 1.28 \pm 0.02$		$\overline{I_{eq}} = 1.33 \pm 0.015$		$\overline{I_{eq}} = 1.41 \pm 0.02$		$\overline{I_{eq}} = 1.51 \pm 0.02$		$\overline{I_{eq}} = 1.52 \pm 0.02$	
Calculated: $I_{eq} = 1.27$		$I_{eq} = 1.35$		$I_{eq} = 1.46$		$I_{eq} = 1.55$		$I_{eq} = 1.62$	

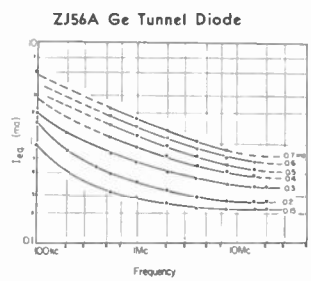


Fig. 1—Noise vs frequency for constant diode current in the upper positive region of the characteristic.

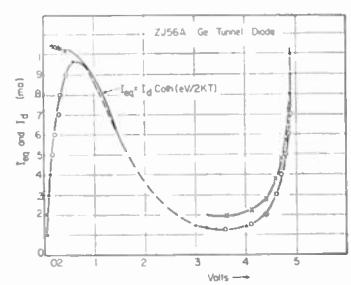


Fig. 2—Diode current and shot noise vs voltage for ZJ56A.

Somewhat different results were obtained for another type GE diode, *viz.*, IN2940. The noise for low bias is white full shot noise as before. Also for very large bias, I_{eq} approaches I_d, *i.e.*, the *p-n* junction current behaves as expected. However, the shot noise in the valley region of the characteristic exceeds full shot noise by one order of magnitude. It is not likely that this must be attributed to the occurrence of two opposite tunneling currents. Apparently, the excess current, presumably composed of electrons which tunnel via states in the forbidden gap to the valence band, is quite noisy and cannot be attributed to independent random events. Further investigation of the noise in this region as well as in the negative resistance region is in progress.

The authors would like to thank Dr. van der Ziel for his stimulating interest.

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[1] T. Yajima and L. Esaki, "Excess noise in narrow *p-n* junctions," *J. Phys. Soc. Japan*, vol. 13, pp. 1281–1287; November, 1958.

[2] C. A. Lee and H. C. Montgomery, "Determination of forward and reverse tunneling currents in Esaki diodes by shot noise measurement," *Bull. Am. Phys. Soc.*, vol. 5, p. 160; March, 1960, and private communication.

[3] J. J. Tieman, "Shot noise in tunnel diode amplifiers," Proc. IRE, vol. 48, pp. 1418–1423; August, 1960.

[4] We shall refer to "shot noise" as the noise observed at sufficiently high frequencies when the spectrum has leveled off to a constant value; we shall refer to full shot noise when Schottky's formula for the tunneling currents of each separate polarity is satisfied.

[5] R. A. Pucel, "The equivalent noise current of Esaki diodes," Proc. IRE, vol. 49, pp. 1080–1081; June, 1961. This expression can also be obtained, writing I_{eq} in the form of the overlap integral, which transforms into $A(Ev - Ec)^2[1 - \exp(-eV/kT)]\ Y$ where Y is a numerical integral defined by Bates. (See C. W. Bates, "Tunneling currents in Esaki diodes," *Phys. Rev.*, vol. 121, pp. 1070–1071; February, 1961.)

Ray Theory in the Ionosphere*

Ray theory remains useful as long as the ray path prescribes the path of the flow of electromagnetic energy. Assume a linearly stratified ionosphere whose electron density varies with height. Let there be no magnetic field and let collisions be neglected. At a point P in this medium let the electron density be N_e. For a frequency ν the refractive index μ at P is

$$\mu = (1 - N_e e^2/\pi m \nu^2)^{1/2}. \quad (1)$$

where e is the charge of the electron and m is its mass. It is synonymous to (1) that μ at P is isotropic so that $\mu = \mu_\Psi$ at P for $0 \leq \Psi \leq \pi$. The group velocity v_g of an electromagnetic wave in this medium is defined as[1]

$$v_g = c\mu \quad (2)$$

for $\mu < 1$(1). The phase velocity $v_{ph} = c/\mu$ such that $v_{ph} \cdot v_g = c^2$, where c is the velocity of light attributed to free space. In this idealized ionosphere a ray path is determined by means of Snell's law,

$$\mu \sin \Psi = \sin i. \quad (3)$$

The product $\mu \sin \Psi$ is a constant along the ray path whose angle of incidence at the linear interface between free space and the ionized medium is i.

With these prerequisites an elementary approach can be used to derive: the equivalent normal-incidence frequency, the theorems for the equivalence of path, and of virtual height.[2]

For $\Psi = \pi/2$ (3) the isotropy of μ (1) yields

$$(1 - N_e e^2/\pi m \nu^2)^{1/2} = \sin i \quad (4)$$

and

$$\nu \cos i = (N_e e^2/\pi m)^{1/2}. \quad (5)$$

For a normal-incidence frequency ν' which is reflected at $P(\mu = 0)$ it follows from (1) for $\nu \rightarrow \nu'$

$$\nu' = (N_e e^2/\pi m)^{1/2}, \quad (6)$$

and from (5), (6)

$$\nu \cos i = \nu' = \nu_1 \cos i_1 = \nu_2 \cos i_2 = \cdots = \nu_n \cos i_n, \quad (7)$$

where ν' is the equivalent normal-incidence frequency. Although N_e was assumed to vary with height such that $dN_e/dh \neq 0$, the isotropy of μ (1) is a valid assumption since (7) agrees with existing knowledge. Let us now derive the "equivalence of path" theorem by an equally simple reasoning.

If μ of (2) is substituted into (3) one obtains

$$v_g \sin \Psi = c \sin i = v_x, \quad (8)$$

where v_x is the velocity component of the flow of electromagnetic energy in the x direction. Since $v_x = c \sin i$ is a constant the same horizontal speed of the energy flow is obtained if the wave is unrefracted, *i.e.*, if it continues to travel along the direction of incidence with the velocity of light in free space. With the assumptions made initially (8) is true along the entire ray path. Thus, both the wave traveling with v_g along the true path and the wave traveling with c along the virtual path, reach any particular value of x within the same time interval. In regards to total travel time true and virtual path are therefore equivalent. This theorem of the equivalence of paths can therefore be understood without the use of integrals. A similar elementary approach yields the theorem for the "equivalence of virtual heights."

In order to derive the theorem for the equivalence of virtual heights the rays belonging to (ν_1, i_1), (ν_2, i_2), (ν_3, i_3), etc., which satisfy (7), are usually assumed to be returned from the same true height. Since the validity of Snell's law (3) is independent of the height variation of N_e it is possible to use any $N_e(h)$ distribution without restricting the conclusion. The simplest case is a homogeneous ionosphere $[N_e(h) = \text{constant}]$ which is linearly bounded against free space. Instead of asking for the same "reflection" height let us determine the relationship between both virtual heights y_{v_1}, y_{v_2} corresponding to the rays (ν_1, i_1) (ν_2, i_2) as they reach the same true height h. In Fig. 1 the depicted nomenclature of a ray path can be indexed and serve a double purpose. It follows by definition

$$v_{g_1} \tau_1 \cos \Psi_1 = v_{g_2} \tau_2 \cos \Psi_2 = h, \quad (9)$$

where $v_{g_1} = c\mu_1$, $v_{g_2} = c\mu_2$ and τ_1, τ_2 are the time intervals needed by (ν_1, i_1) (ν_2, i_2) to reach the true height, h. From (1)–(3), and (9) one obtains

$$\frac{\tau_1}{\tau_2} = \frac{(\cos^2 i_2 - N_e e^2/\pi m \nu_2^2)^{1/2}}{(\cos^2 i_1 - N_e e^2/\pi m \nu_1^2)^{1/2}}. \quad (10)$$

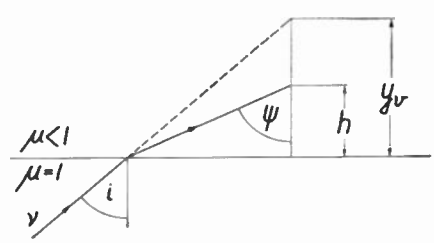


Fig. 1.

From the previous path-equivalence theorem follows:

$$y_{v_1} = c\tau_1 \cos i_1, \quad y_{v_2} = c\tau_2 \cos i_2. \quad (11)$$

From (10), (11)

$$\frac{y_{v_1}}{y_{v_2}} = \frac{\tau_1 \cos i_1}{\tau_2 \cos i_2} = \frac{(1 - N_e e^2/\pi m \nu_2^2 \cos^2 i_2)^{1/2}}{(1 - N_e e^2/\pi m \nu_1^2 \cos^2 i_1)^{1/2}}. \quad (12)$$

Because of (7) $y_{v_1} = y_{v_2}$ and the "equivalent virtual height" theorem is confirmed.

The inequality[3]

$$\frac{v_{ph}}{2\pi\mu\nu} \frac{d\mu}{dh} \ll 1 \quad (13)$$

determines whether or not a ray path does represent the path of energy flow. Neither in the above derivations, nor elsewhere[2] was (13) explicitly considered. Moreover, in the derivation of the "equivalent virtual height" theorem a discontinuity was assumed. These circumstances suggest that the theorems are a corollary of Snell's law. Their truth is contingent upon the truth of the latter. The physical meaning of the ray paths is subsequently decided by using (13).

For a circular stratification (3) is replaced by Bouguer's law

$$(\mu_1 r_1 \sin \Psi_1 = \mu_n r_n \sin \Psi_n = \text{constant})$$

and the above theorems lose their exact meaning. For frequencies for which electron collisions cannot be neglected, simple ray theory no longer applies. Allowing for the magnetic field, which makes the ionosphere an anisotropic medium, further restricts the usefulness of these theorems.

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* Received, March 6, 1962; revised manuscript received, March 19, 1962.

[1] H. R. Mimno, "The physics of the ionosphere," *Rev. Mod. Phys.*, vol. 9, pp. 1–43; January, 1937.

[2] S. K. Mitra, "The Upper Atmosphere," The Asiatic Society, Calcutta, India, pp. 237–240; 1952.

[3] D. E. Kerr, "Propagation of Short Radio Waves," M.I.T. Rad. Lab. Ser., McGraw-Hill Book Co., Inc., New York, N. Y., vol. 13, pp. 53–58; 1951.

On "The Noise Figure of Negative Conductance Amplifiers"*

In a recent communication Brown[1] proved that a feedback amplifier used as a negative conductance amplifier can have an optimum noise figure much lower than the optimum noise figure of the same amplifier without feedback. It is the aim of this correspondence to show that this result is due to an error in the calculation.

The error occurs first in Brown's (8), where the factor $A^2\beta^2$ in front of E_{n2}^2 has been omitted. As a consequence Brown's (11)–(13) are in error. We shall now replace Brown's (8), (11), (12), and (13) by the corrected (8a), (11a), (12a) and (13a).

The noise figure for direct coupling between the source and the negative conductance circuit is

$$F' = 1 + \frac{A^2\beta^2 E_{n2}^2}{I^2 Z_1^2 (1 - A\beta)^2} + \frac{E_{n1}^2}{N^2 I^2 Z_1^2}. \quad (8a)$$

If the source is coupled through an ideal transformer of turns ratio N, I must be replaced by NI and Y_s by $N^2 Y_s$, so that (10) may be written

$$N^2 Y_s \approx -\frac{1}{Z_1(1 - A\beta)},$$

or

$$A\beta \approx \left(1 + \frac{1}{N^2 Y_s Z_1}\right).$$

Substituting into (8a), one obtains

$$F' = 1 + \frac{E_{n2}^2 (N^2 Y_s + 1/Z)^2}{N^2 I^2} + \frac{E_{n1}^2}{N^2 I^2 Z_1^2}. \quad (11a)$$

* Received February 5, 1962.

[1] F. W. Brown, "The noise figure of negative-conductance amplifiers," PROC. IRE (*Correspondence*), vol. 49, pp. 520–521; February, 1961.

Considered as a function of N, F' is optimized for

$$N^2 = \frac{\sqrt{E_{n2}^2 + E_{n1}^2}}{E_{n2} Y_s Z_1} \tag{12a}$$

and the optimized value is

$$F'_{\text{opt}} = 1 + \frac{2Y_s}{I^2 Z_1} \cdot \left[E_{n2}^2 + E_{n2}\sqrt{E_{n2}^2 + E_{n1}^2}\right] \tag{13a}$$

which is exactly equal to the optimum noise figure without feedback. [See Brown's (3).]

The conclusions are therefore that the two amplifiers have the *same* noise figure and that Brown's analysis cannot explain Ishii's results on negative resistance klystron amplifiers.[2]

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[2] K. Ishii, "Noise figures of reflex klystron amplifiers," IRE TRANS. ON MICROWAVE THEORY AND TECHNIQUES, vol. MTT-8, pp. 291–294; May, 1960.

Analysis and Synthesis of a Class of Nonlinear Systems*

The procedure outlined in this work is a numerical technique which can be used for the purpose of analysis and synthesis of a very broad class of nonlinear systems where the operation of the system can be represented by a linear differential equation over any arbitrarily small time interval. The work is divided into two parts:

1) Formulation of the piece-wise linear analysis,
2) Use of the results of (1) for the purpose of systematical synthesis.

The formulation of the piece-wise linear analysis and a simple example is given below. From an application point of view it compares with the Z-form technique suggested by Boxer and Thaler[1] although it is simpler to apply (there is no need for a table of transforms such as Z-form tables, and the result is more accurate for a number of problems). The shortcoming of the technique suggested here is the necessity of obtaining the roots of the characteristic equation corresponding to the linearized equation.

Let a system be described by the equation

$$f(y, \dot{y}, \ddot{y}) = 0 \tag{1}$$

and assume the solution exists and is

$$y = y(t). \tag{2}$$

The function $f(y, \dot{y}, \ddot{y})$ can readily be approximated by a linear differential equation over the period $nT \le t \le (n+1)T$ where T is small. For this purpose the necessary initial conditions are $y(nT)$ and $\dot{y}(nT)$, *i.e.*, the final values of the preceding period $(n-1)T < t < nT$. In terms of these initial conditions $Y_n(s)$, the Laplace transform of $y(t)$ can be obtained and is valid only during the interval $(nT < t < n+1)T$. Consequently, since

$$y(nT + t) = \frac{1}{2\pi i}\int_{c-j\infty}^{c+j\infty} Y_n(s) e^{st} ds \tag{3}$$

where $0 \le t \le T$,

$$y[(n+1)T] = \frac{1}{2\pi i}\int_{C-j\infty}^{C+j\infty} Y_n(s) e^{sT} ds = \sum \text{Res. of } Y_n(s) e^{sT}. \tag{4}$$

The initial conditions $y(nT)$ and $\dot{y}(nT)$ are used to determine the $Y_n(s)$. In other words $y(n+1)T$ is obtained from $f(y, \dot{y}, \ddot{y}) = 0$ and initial conditions. The steps to be followed are summarized in the following:

1) The quantity T is chosen such that $f(y, \dot{y}, \ddot{y}) = 0$ can be approximated by a linear equation over any interval with a duration T seconds.
2) The Laplace transform of $y(t)$ is obtained in terms of $y(nT)$ and $\dot{y}(nT)$ which is $Y[s, y, (nT), \dot{y}(nT)]$, valid only when $nT \le t \le (n+1)T$.
3) The $y(n+1)T$ is obtained which is the sum of residues of
$$Y[s, y(nT), \dot{y}(nT)]e^{st}.$$
4) The actual given initial conditions are used to obtain $y(T)$ and then systematically $y(nT)$ will be obtained up to any desired $n = N$.

Example

$$\dot{y} + y^2 = 1. \tag{5}$$

Let $y = y(nT) =$ constant, over the interval $nT < t < (n+1)T$ and take $T = 0.1$ sec. Therefore

$$\dot{y} + y(nT)y = 1 \tag{6}$$

$$Y_n(s) = \frac{1 + sy(nT)}{s[s + y(nT)]}. \tag{7}$$

Hence

$$y[(n+1)T] = \sum \text{Res. of } [Y_n(s)e^{0.1s}] = \frac{1 - [1 - y^2(nT)]e^{-.1y(nt)}}{y(nT)}. \tag{8}$$

The original nonlinear equation happens to have an exact solution

$$y = \tanh(t). \tag{9}$$

The following table is a comparison between the solution obtained through the suggested technique and the exact solution.

t	$y(t)$	$y(t)$ exact
0	10	0
0.1	0.1000	0.09967
0.2	0.1941	0.1970
0.3	0.2856	0.2910
0.4	0.3816	0.3799

The second phase of the work is the synthesis of nonlinear systems by using the above technique. The procedure is given as follows:

1) Adopt a specific value for T. It is noted that the computational error involved in this case is not additive; consequently T can be made as small as required to obtain the nonlinear function as accurately as desired.
2) From the knowledge of $y(nT)$ and $y(n+1)T$ the value for the nonlinear element is obtained in a manner similar to the analysis procedure described above.

Example

Suppose the synthesis of nonlinear first-order equation of the form

$$\dot{y} + f(y)y = u(t) \tag{10}$$

is desired. The desired response to a unit step may be given by any single-valued curve. The problem is then to determine $f(y)$ as a function of y. The unknown function $f(y)$ is obtained in an approximate form in a manner that it assumes a constant value over the interval $T \le t \le (n+1)T$. T can be chosen arbitrarily small, based on the desired degree of approximation.

From the results of the previous section

$$y(n+1)T = \frac{1}{f(y)}\{1 - [1 - y(nT)f(y)]e^{-Tf(y)}\}. \tag{11}$$

Successively substituting the values for $y(nT)$ and $y(n+1)T$, values for $f(y)$ can be obtained as a function of y from (11).

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Inductive Probability as a Criterion for Pattern Recognition*

The above communication[1] raises the question of how to apply Carnap's concept of inductive probabilities in predicting, from a finite sequence of events, a probability pattern for the next events. This problem is partly solved by splitting up the sample sequence into sections of equal length and predicting from their frequency, in terms of the wellknown Laplace formula, the probability of the future occurrence of such sections. In closing, the question is raised of how to proceed if the sample sequence is not composed of a complete period, *i.e.*, if the chosen length of the sections is not an aliquot part of the length of the sample sequence.

On first assessment, this problem appears to me to be quite readily solvable by drop-

* Received January 15, 1962.
[1] R. Boxer, and S. Thaler, "A simplified method of solving linear and nonlinear systems" PROC. IRE, vol. 44, pp. 89–101; January, 1956.

* Received March 5, 1962.
[1] B. Harris, PROC. IRE (*Correspondence*), vol. 49, pp. 1951–1952; December, 1961.

ping as many symbols at the beginning of the sequence as is necessary to obtain a complete period. This amounts, it is true, to the sacrifice of available information, but this shortcoming is inherent in the very nature of the approach described.

To make maximum possible use of the information contained in the sample sequence, the statistical analysis should, in my opinion, be extended to cover *all* pairs, triples, etc. Above all, the conditional probabilities should also be analyzed. To make this clear, I would refer to the example cited.

The sample sequence is

$$A\ B\ A\ B\ A\ B\ A\ B.$$

A statistical analysis of the individual symbols at first yields the quantities

$$n(A) = n(B) = 4;$$

hence

$$p(A) = p(B) = \frac{4+1}{8+2} = \frac{1}{2}.$$

We now proceed with calculating the conditional probability for the occurrence of the two symbols if symbol B is the condition (which amounts to the obvious assumption that the process will produce a Markov chain). The statistical analysis of the sequence yields

$$n(BA) = 3, \qquad n(BB) = 0;$$

hence

$$p(B \mid A) = \frac{3+1}{3+2} = \frac{4}{5},$$

analogous is $p(B|B) = \frac{1}{3}$. (In designating the conditional probability, the representation chosen here deviates from the conventional form in that the condition has been placed *ahead* of the vertical stroke and the symbol in question behind this stroke.)

The same values are obtained when investigating the triples (ABA) and (ABB). The quadruples yield

$$n(BABA) = 2,$$

$$n(BABB) = 0,$$

$$p(BAB \mid A) = \frac{2+1}{2+2} = \frac{3}{4}.$$

Thus, greatest significance in this case has the analysis of the pairs and triples, on the basis of which it can be predicted with a probability of $\frac{4}{5}$ that the next symbol will be an A.

In order to be able to predict, in addition to the next symbol, also the symbol following the next (which is different from predicting only the symbol following the next), we start with the statistical analysis of all pairs. We arrive at $m=4$ possibilities. We find that

$$n(AB) = 4,\ n(BA) = 3,\ n(AA) = n(BB) = 0$$

and

$$p(AB) = \frac{4+1}{7+4} = \frac{5}{11}$$

$$p(BA) = \frac{3+1}{7+4} = \frac{4}{11}$$

$$p(AA) = p(BB) = \frac{1}{11}.$$

Furthermore, for the conditional probabilities there results

$$n(BAB) = 3,$$

$$n(BAA) = n(BBA) = n(BBB) = 07$$

and hence

$$p(B \mid AB) = \frac{3+1}{3+4} = \frac{4}{7}$$

(the other three probabilities are each given by $\frac{1}{7}$).

When taking the last two symbols as condition, we obtain the same value; when taking more, the probability will again decline.

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Author's reply[2]

This is not a criticism of Dr. Berger's comment, but rather a request for further extensions of his ideas.

The purpose of applying the concept of inductive probability to a finite sequence of events was not to determine a probability pattern for the next events (although this was the immediate result), but rather to assign measures to the various possible patterns that could be present. The patterns considered have three characteristics: basic repetition period, sequence of characters of which it is composed, and absolute location within the given sequence. The main difference between Dr. Berger's approach and mine is the attempt to maintain information as to the absolute location of the periodicity. Extending the technique to cover all pairs, triples, etc., gives a measure which is, in effect, an average over all possible absolute locations. How does one avoid this and yet not sacrifice available information?

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[2] Received March 28, 1962.

True RMS Measurement with Tunnel Diodes*

Neu's communication on a tunnel-diode frequency-doubling circuit[1] aroused considerable interest here, since I had previously done some work in the same area, by describing a similar circuit.[2]

I wish to stress that the same principle was also applied for measuring the true rms value of ac voltages. Namely, if the operating point of the tunnel diode is the peak point of the characteristic (where it is of quadratic character) and an ac voltage $U.\cos \omega t$ is superimposed on the dc component, Fig. 1, then the current of the tunnel diode has an ac component $\frac{1}{2}KU^2 \cdot \cos 2\omega t$, and the dc component changes with $\frac{1}{2}KU^2$. The latter is proportional to the rms value. Fig. 2 shows experimental results. The circuit measured the true rms values as was proven by a measurement using two independent ac voltages.

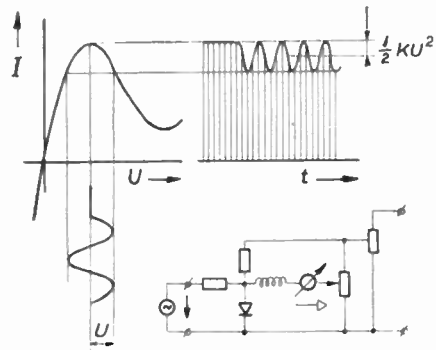


Fig. 1.

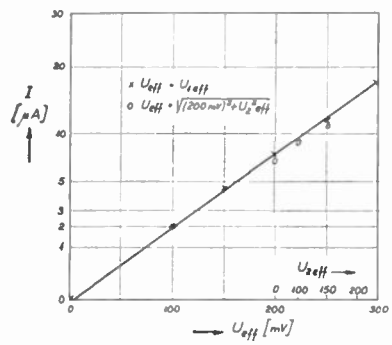


Fig. 2.

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* Received March 7, 1962.

[1] F. D. Neu, "A Tunnel-diode wide-band frequency doubling circuit," PROC. IRE (*Correspondence*), vol. 49, pp. 1963–1964; December, 1961.

[2] K. Tarnay, "The Tunnel Diode," Doctoral dissertation, Technical University, Budapest, Hungary, pp. 139–141 and 151–155; May, 1961.

A Table of Solutions of Riccati's Equations*

During 1961 many workers [1–8] wrote on Riccati's nonlinear differential equations (NDE's) which have a history of well over two hundred years. Perhaps one of the many reasons why Riccati's NDE's are so important is that they replace all homogeneous

* Received March 8, 1962. A part of this note was presented at the meeting of the American Mathematical Society, Atlantic City, N. J., April 19, 1962 as "Riccati's Equation with Loosely Interrelated Coefficients."

second-order ordinary linear differential equations (LDE's) by simple changes of variables. These equations describe many physical systems, and matters of current interest such as tapered lines, $2n$-port circuits, RF waves in the ionosphere and time-varying networks. Nevertheless, the exact solution of a generalized Riccati's NDE having three independent arbitrary variable coefficients is still unknown.

Under these prevailing circumstances, a handy table of exact solutions for certain special though physically significant cases may be useful. This correspondence provides a table of solutions of Riccati's NDE's with interrelated variable coeficients, and gives a brief summary of the present state of the art.

McLachlan [9] shows in his book that an exact particular solution, $y=1/x$, satisfies

$$y'(x) + xg(x)y(x) + y^2(x) = g(x), \quad (1)$$

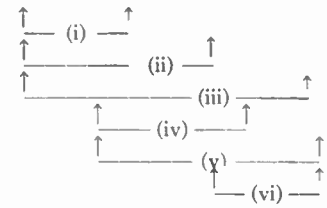
where $y(x)$ is the unknown, the prime is used for a derivative with respect to x, which is the only independent variable such as distance or time, and $g(x)$ is any known function of x. This solution is obvious if (1) is regrouped as,

$$(y' + y^2) + (xgy - g) = 0. \quad (2)$$

Suppose $f(x)$, another arbitrary known function of x, is employed instead of $g(x)$ in the right-hand side of (1), then $y=f/(xg)$ results from the second bracket group, while $y=1/x$ comes from the first bracket group. To eliminate this apparent discrepancy, the requirement of $f(x)=g(x)$ is introduced as a natural answer.

Following the same line of thought, all possible combinations of regrouping a generalized Riccati's NDE are applied. Six (as ${}_4C_2=6$) ways to bracket such an equation are indicated by linking lines in (3),

$$y'(x) + P(x)y(x) + Q(x)y^2(x) = R(x). \quad (3)$$



Unfortunately these six particular exact solutions tabulated below must accompany their corresponding necessary interrelationship among three general coefficients, $P(x)$, $Q(x)$ and $R(x)$.

Tabulated exact particular solutions of $y'(x) + P(x)y(x) + Q(x)y^2(x) = R(x)$

	A particular solution	Required interrelationship
(i)	$ae^{-\int P dx}$	$R = a^2Qe^{-2\int P dx}$
(ii)	$\dfrac{1}{\int Q dx + b}$	$R = \dfrac{P}{\int Q dx + b}$
(iii)	$\int R dx + c$	$Q = -\dfrac{P}{\int R dx + c}$
(iv)	$-\dfrac{P}{Q}$	$R = -\left(\dfrac{P}{Q}\right)'$
(v)	$\dfrac{R}{P}$	$Q = \left(\dfrac{P}{R}\right)'$
(vi)	$\sqrt{\dfrac{R}{Q}}$	$P = -\left(\ln\sqrt{\dfrac{R}{Q}}\right)'$
(vii)	$\dfrac{e^{-\int(P-2Q)dx} - \int Pe^{-\int(P-2Q)dx}dx - a}{e^{-\int(P-2Q)dx} + \int Pe^{-\int(P-2Q)dx}dx + a}$	$R = Q - P$
(viii)	$\dfrac{\int Pe^{-\int(P+2Q)dx}dx + b - e^{-\int(P+2Q)dx}}{\int Pe^{-\int(P+2Q)dx}dx + b + e^{-\int(P+2Q)dx}}$	$R = Q + P$

Note: a, b, and c are arbitrary constants.

There are actually three independent table entries in (i) through (vi) because, for example, (iii) and (iv) furnish the identical general solution of (3). Rajagopal [8] reports (i) originating from a totally different approach. Similarly (iv) and (v) were previously obtained by the writer using linearizing transforms, $y=1/(fg)$ and $y=fg$ respectively [3]. However, it appears difficult to identify the linearizing transform for a generalized Riccati's NDE for (vi). It was in the course of this searching process that a new and practical application was found.

A new transform of the dependent variable is

$$y(x) = \frac{f(x) - g(x)}{f(x) + g(x)}. \quad (4)$$

When (4) is substituted in (3), another Riccati's NDE is derived,

$$2h' - 2(Q + R)h + (P + Q - R)h^2 = (P + R - Q), \quad (5)$$

where

$$h(x) = f(x)/g(x).$$

Eq. (5) is readily solved exactly if proper interrelationships among variable coefficients are realized. When, for instance,

$$R(x) = Q(x) - P(x), \quad (6)$$

(5) is reduced to a solvable Bernoulli's NDE. Alternatively, if

$$R(x) = Q(x) + P(x), \quad (7)$$

(5) becomes a first-order LDE giving an exact solution. Since linearizing transforms of the dependent variable in a generalized Riccati's NDE exhibit dualism [3], it is easy to see that (6) and (7) can be obtained by using,

$$y(x) = \frac{f(x) + g(x)}{f(x) - g(x)}. \quad (8)$$

There are several ways to linearize Riccati's NDE [3], [10]. Two simple ways are to use transforms, $y=u'/(Qu)$ and $y=(u'-Pu)/(Qu)$, where $u(x)$ is any function of x. These transforms with (6) or (7) linearize (3) respectively,

$$u'' + \left(P - \frac{Q'}{Q}\right)u' - (Q \mp P)Qu = 0 \quad (9)$$

and

$$u'' - \left(P + \frac{Q'}{Q}\right)u' - \left[\left(\frac{P}{Q}\right)' + Q \mp P\right]Qu = 0 \quad (10)$$

where $P(x)$ and $Q(x)$ are two independent, known, arbitrary functions of x. The degree of contribution of (6) and (7) may be more or less proportional to the frequency of appearance of solved equations (9) and (10) in practical physical systems.

These "relaxed" interrelationships, (6) and (7), compared with more complicated other interrelationships such as, $R=-(P/Q)'$, $Q=(P/R)'$, and $P=-(\ln\sqrt{R/Q})'$, are new to the best knowledge of the writer

in the literature of Riccati's NDE's. These eight table entries represent a great number of specialized cases, for example, McLachlan's illustration [9], (1) is a special case of (v). Many existing text books carry exercise problems of Riccati's NDE's together with specified variable coefficients, such as $P(x) = \exp(x)$, $Q(x) = \sin x$, and $R(x) = \ln x$.

It is felt, however, that the foregoing represents a set of more useful relationships for the solutions of problems in the rapidly growing nonlinear systems.

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[9] N. W. McLachlan, "Ordinary Non-Linear Differential Equations in Engineering and Physical Sciences," Clarendon Press, Oxford, England, 2nd ed., p. 13; 1956.
[10] I. Sugai, "Riccati's nonlinear differential equation," *Am. Math. Monthly*, vol. 67, pp. 134–139; February, 1960.

On a New Security Problem*

STATEMENT OF THE PROBLEM

If we are successful in the near future in contacting an alien culture some light years distant, the following facts, which are immediately apparent, should be considered.

1) We will be participating in such communication at the earliest date that it would be technologically feasible for us to do so, *i.e.*, we would be beginners.
2) It would be positing an absurd coincidence to suggest that our interstellar correspondent would be in the same situation. In other words, arguing on probabilistic grounds, it is almost certain that the alien culture would be some centuries, millennia, or eons ahead of us in the development of physical science.
3) It follows therefore that any military contact between the communicants would find us in a position of complete helplessness.

Some weak points of this argument have already received some consideration, as follows.

4) Dr. E. M. Purcell, of Harvard University, has recently (at Green Bank, W. Va.) argued in painstaking fashion that interstellar logistics would require the expenditure of unthinkable amounts of energy over great periods of time and would therefore be impractical.
5) One may speculate that advanced cultures will have reached the point where their ability to create material wealth will have saturated their demand for it. Many social scientists have expressed the opinion that human society will lose its pugnacious characteristics when we have reached such an elevated level of technology. If this is true of human cultures, it may also be a universal law. In that case, we would have nothing to fear from contact with any culture that is advanced enough to solve the problems of interstellar logistics.

TENTATIVE CONCLUSION

While arguments 4) and 5) may find many adherents, I would like to emphasize that 3) can be affirmed with almost complete certainty, whereas 4) and 5) are highly speculative, and do not have the fail-safe characteristic. It follows, therefore, that

6) As a matter of practical security, immediate steps should be taken to ensure that no irresponsible attempts will be made to effect two-way interstellar communication. No broadcasts should be made. If informative signals are received from an intelligent stellar source, they should not be answered, but every effort should be made to decode them and to study their implications.

DISCUSSION OF POSSIBLE OBJECTIONS

I think that many scientists will be inclined to re-assert Dr. Purcell's argument with great emphasis, insisting that he has demonstrated very fundamental limits on the possibilities of interstellar travel, in respect to both the energy cost and the extreme length of the trip.

Anticipating such a conflict of opinion, I would like some consideration given to the following points.

7) The duration of the trip is objectionable to us because it represents, even under the most favorable assumptions, a substantial fraction of our life span. This is a highly anthropocentric point of view. It does not follow that an alien would take the same viewpoint. Some but not all of the reasons why he might see the matter differently are
   a) He may have a greatly longer life span.
   b) He may have cultural attitudes which do not lead to impatience under trip conditions. For example the military expansion of his culture might be undertaken not to expand his economy but by reason of instinct—a sort of generalized procreative drive. (It has even been suggested that this is true of human beings.)
   c) If his culture is able to supply normal material and social comforts during the trip then it is hard to see why any objection would arise.
8) Regarding the energy cost, Dr. Purcell has assumed total conversion of matter to energy, presumed to be the "ultimate" energy source, and found that one proposed trip would require some 400,000 tons of matter and antimatter as fuel. I would like him to give his reasons for supposing that a wealthy alien would consider this an expensive trip. How many alien bank accounts has Dr. Purcell audited?
9) The arguments that a proposed action is impossible may take two different forms. A mathematical impossibility is absolute; thus it will always be true that the rules of Euclidean geometry do not permit the exact trisection of a general angle with straight edge and compass alone. But Dr. Purcell's argument is not of this type.

He assumes that he can describe the ultimate energy source though he cannot yet make use of it. While this notion has a definite aura of plausibility, it is far from rigorous. The history of science has shown that this general type of argument is a very hazardous one to use because the advance of technology has a way of rendering such arguments obsolete without directly proving them false. While we might indulge in a little science fiction to turn up possible reasons why Dr. Purcell's argument may eventually fail, I would prefer a more rigorous approach.

In recent decades, mathematicians have been working on a "theory of confirmation" in order to answer questions of this type: to exactly what degree can a theory be said to be confirmed by known facts? We suggest therefore that a study of Dr. Purcell's argument be made along these lines.

FINAL CONCLUSION

If such a study should assign a low-confidence figure to the unreasonable-energy-cost argument (I, for one, think that it would), this would greatly reinforce the request for rigorous security precautions.

It seems to me merely a matter of common sense that we should be very careful about our first contact with a nonhuman culture, especially since it appears so likely that they would be much superior to us in experience, scientific knowledge, and weaponry. We may have only one chance to display our common sense in this matter.

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* Received February 6, 1962.

The Effects of Doppler Dispersion on Matched Filter Performance*

We have noted with interest the recent correspondence of Ramp and Wingrove,[1] and Cook and Heiss,[2] commenting on the effects of Doppler dispersion on "linear FM-pulse compression." Since linear FM-pulse compression falls within the broader domain of matched filtering, the effects of Doppler dispersion can be treated with greater generality by modifying previous expressions for the "matched filter uncertainty function,"[3] in which Doppler has been approximated by a simple frequency translation.

The Doppler effect results from a continuous compression or expansion of the path length traversed by a signal during transmission. Mathematically this is exactly represented by a change in time scale and change in amplitude (to conserve energy) as follows:

$$f(t) \rightarrow \sqrt{\alpha}\, f(\alpha t). \tag{1}$$

It is shown first that the time-bandwidth product of a signal is not modified by Doppler. Following Gabor,[4] the normalized mean and mean-square pulse widths of a Doppler shifted signal are defined by the moments of $\sqrt{\alpha}f(\alpha t)$:

$$\overline{t_\alpha^n} = \frac{1}{E}\int_{-\infty}^{\infty} t^n \left|\sqrt{\alpha}\, f(\alpha t)\right|^2 dt = \overline{t^n}/\alpha^n \tag{2}$$

where E is the signal energy. The effective duration of a Doppler modified signal Δt_α in terms of the unshifted ($\alpha = 1$) signal duration Δt is found from Gabor's defining relation and our (2) to be

$$(\Delta t_\alpha)^2 = \overline{t_\alpha^2} - (\overline{t_\alpha})^2 = \frac{1}{\alpha^2}\left[\overline{t^2} - (\overline{t})^2\right]$$

$$= (\Delta t)^2/\alpha^2. \tag{3}$$

Similarly, the mean and mean-square bandwidths of $\sqrt{\alpha}f(\alpha t)$ are related by Gabor to the moments of $G(w/\alpha)/\sqrt{\alpha}$, which is defined as twice the positive spectrum of $\sqrt{\alpha}f(\alpha t)$, as follows:

$$\overline{w_\alpha^n} = \frac{1}{2E}\int_0^{\infty} w^n \left|\frac{1}{\sqrt{\alpha}}\, G\left(\frac{w}{\alpha}\right)\right|^2 \frac{dw}{2\pi}$$

$$= \alpha^n \overline{w^n}. \tag{4}$$

In a similar way, the effective bandwidth of a Doppler shifted signal Δw_α in terms of the unshifted ($\alpha = 1$) signal bandwidth Δw is found from Gabor's defining relation and our (4) to be

$$(\Delta w_\alpha)^2 = \alpha^2 (\Delta w)^2. \tag{5}$$

The time-bandwidth product of a Doppler shifted signal in terms of the unmodified waveform may be found from (3) and (5) to be:

$$\Delta t_\alpha \Delta w_\alpha = \left(\frac{\Delta t}{\alpha}\right)(\alpha \Delta w) = \Delta t \cdot \Delta w. \tag{6}$$

Since the time-bandwidth product of a signal is not altered by Doppler, its dimension in signal space is unchanged.

An approximation to include the effects of Doppler dispersion will now be developed. A narrow-band Doppler shifted signal may be written in polar complex notation as

$$\sqrt{\alpha}\, f(\alpha t) = \sqrt{\alpha}\ \mathrm{Re}\left\{a(\alpha t) e^{j[w_0 \alpha t + \phi(\alpha t)]}\right\}, \tag{7}$$

where $a(t)$ is the envelope function, $\phi(t)$ the phase modulation, and w_0 the carrier frequency. Since α is always very close to unity, it is usually replaced by $(1+\delta)$, where δ is very closely approximated by

$$\delta \approx \pm 2v/c. \tag{8}$$

In (8), v is the rate of change of effective transmission length and c represents the velocity of propagation in the transmission medium.

If $a(t)$ and $\phi(t)$ are well-behaved functions, $a(t+\delta t)$ and $\phi(t+\delta t)$ may be expanded in a Taylor series about $(\delta t = 0)$:

$$a(t + \delta t) = a(t) + \dot{a}(t)\delta t + O[(\delta t)^2] \tag{9}$$

$$\phi(t + \delta t) = \phi(t) + \dot{\phi}(t)\delta t + O[(\delta t)^2]. \tag{10}$$

Substituting (9) and (10) into (7) and ignoring terms in $(\delta t)^2$ or higher, results in

$$\sqrt{\alpha}\, f(\alpha t) = \sqrt{\alpha}\ \mathrm{Re}\left\{\left[a(t) + \dot{a}(t)\delta t\right] \cdot e^{j[w_0 t + w_0 \delta t + \phi(t) + \dot{\phi}(t)\delta t]}\right\}. \tag{11}$$

The term $\exp\{jw_0\delta t\}$ represents the earlier Doppler approximation, for which $\delta w_0 = w_d$. Assuming $\dot{\phi}(t)\delta t$ is small, the exponential containing this term may be replaced by the first two terms of its series expansion. Neglecting terms in $(\delta t)^2$, (11) become

$$\sqrt{\alpha}\, f(\alpha t) = \sqrt{\alpha}\ \mathrm{Re}\left\{\left[a(t) e^{j(w_0 t + \phi(t) + w_d t)}\right] \cdot \left[1 + \frac{\dot{a}(t)}{a(t)}\delta t + j\dot{\phi}(t)\delta t\right]\right\}. \tag{12}$$

Note that the first bracketed term in (12) approximates Doppler by a linear frequency translation, while the remainder contains first-order correction terms reflecting the effects of frequency dispersion.

Applying the modified Doppler approximation to Siebert's[3] "uncertainty function" yields

$$\psi_m(\tau, w_d) = \frac{\left|\sqrt{\alpha}\int_{-\infty}^{\infty} u(t)u^*(t+\tau)e^{jw_d t}\left\{1 + \frac{\dot{a}(t)}{a(t)}\delta t + j\dot{\phi}(t)\delta t\right\}dt\right|}{\int_{-\infty}^{\infty} |u(t)|^2 dt}, \tag{13}$$

where

$$\int_{-\infty}^{\infty} |u(t)|^2 dt = 2E, \tag{14}$$

and

$$u(t) = a(t)e^{j\phi(t)}. \tag{15}$$

The absolute magnitude of the first term in the integral is the (conventional) "uncertainty function" $\psi(\tau, w_d)$ in which the effect of Doppler has been approximated by a linear translation.

An upper bound on the difference between (13) and $\psi(\tau, w_d)$ can be found by neglecting $\sqrt{\alpha}$, which is approximately unity, and rewriting (13) as an inequality:

$$\psi_m(\tau, w_d) \leq \psi(\tau, w_d) + \frac{|\delta| \int_{-\infty}^{\infty} [|t|\, a(t+\tau)][\dot{a}^2(t) + a^2(t)\dot{\phi}^2(t)]^{1/2} dt}{2E}. \tag{16}$$

Hölder's Inequality states that

$$\int |f \cdot g| \leq \left\{\int |f|^p\right\}^{1/p} \cdot \left\{\int |g|^{p'}\right\}^{1/p'}, \tag{17}$$

where f and g are members of class L^p and $L^{p'}$ respectively. Since the two bracketed expressions in the second term of (16) will usually be members of Class L^2 for pulsed signals, (16) becomes

$$\psi_m(\tau, w_d) \leq \psi(\tau, w_d) + |\delta| \left\{\frac{\int_{-\infty}^{\infty} [|t|\, a(t+\tau)]^2 dt}{2E}\right\}^{1/2} \cdot \left\{\frac{\int_{-\infty}^{\infty} [\dot{a}^2(t) + a^2(t)\dot{\phi}^2(t)]dt}{2E}\right\}^{1/2}. \tag{18}$$

The second bracketed integral in (18) has been interpreted in the literature[5] as the mean-square frequency about w_0, which is equal to the effective bandwidth Δw when $a(t)$ is symmetrical and $\dot{\phi}(t)$ is zero or has odd symmetry. This relation follows from a substitution of the Fourier transform of (15) into (4). Defining $a(t)$ from $-T/2$ to $T/2$, an upper bound on the integral in the first bracket is found to be $T^2/4$. Hence (18) may be rewritten as

$$\psi_m(\tau, w_d) \leq \psi(\tau, w_d) + |\delta|\, \tau \Delta w/2$$

$$< \psi(\tau, w_d) + |\delta|\, \frac{k}{2}(\Delta t \cdot \Delta w), \tag{19}$$

* Received March 9, 1962.

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[2] C. E. Cook and W. H. Heiss, "Linear FM-pulse compression Doppler distortion effects," to be published in PROC. IRE (*Correspondence*).

[3] W. McC. Siebert, "A radar detection philosophy," IRE TRANS. ON INFORMATION THEORY, vol. IT-2, pp. 204–221; September, 1956.

[4] D. Gabor, "Theory of communication," *J. IEE*, vol. 93, pt. III, pp. 429–457; November, 1946.

[5] P. Bello, "Joint estimation of delay, Doppler and Doppler rate," IRE TRANS. ON INFORMATION THEORY, vol. IT-6, pp. 330–341; June, 1960.

where k relates the effective pulse duration Δt to the actual pulse duration T. Note that the difference term in (19) is proportional to the time-bandwidth product, envelope shape factor k, and δ which is related to velocity.

For a pulse signal with a rectangular envelope, k is approximately 3.5. Assuming relative velocities in the order of 17,000 miles per hour, $|\delta|$ is approximately 0.5×10^{-4}. For this case, the conventional and modified uncertainty expressions will differ by less than 10 per cent for all values of (τ, w_d), if

$$\Delta t \cdot \Delta w < 1100. \tag{20}$$

For a specific value of w_d, the "uncertainty function" represents the output waveform from a matched filter. Hence the above result provides a bound on the reduction in matched-filter peak signal output when the conventional Doppler approximation is used. It is clear that for very large time-bandwidth products and large Doppler velocities, a better Doppler approximation, such as (13), should be employed.

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Mechanical Quadratures to Synthesize Nonuniformly-Spaced Antenna Arrays*

A new method of synthesizing nonuniformly-spaced antenna arrays has been developed and several examples using different field patterns have been used to verify this synthesis technique. In the examples chosen, broadside directive radiation patterns with constant phase variation have been used to synthesize nonuniformly-spaced symmetric arrays.

For a finite aperture of over-all width $2a$, the radiation pattern can be written as

$$F(u) = \int_{-a}^{a} g(x)\cos ux\,dx, \tag{1}$$

where $u = \pi \sin\theta$, θ is the angle measured from the aperture normal and x is the distance from the origin to any point along the aperture measured in half wavelengths. The constant involved has been suppressed, as the field pattern is to be normalized in its final form.

The integral involved in (1) can be evaluated by mechanical quadratures in a form which can be identified with the formulation of an array of discrete elements located within the interval of integration. The specific form of the mechanical quadrature[1] used is

$$\int_a^b f(x)dx = \sum_{j=1}^{n} H_j f(a_j), \tag{2}$$

where the values of a_j and H_j can be evaluated uniquely. As indicated in (2), a suitably weighted sum of n values of $f(a_j)$ can evaluate the integral exactly if $f(x)$ can be written as a polynomial of degree less than or equal to $(2n-1)$. Kopal[1] has shown that the n values of a_j are real, distinct and within the interval (a, b). In addition all values of H_j are real and positive.

Identifying $f(x)$ in (2) with the integrand in (1), $f(x)=g(x)\cos(ux)$, and remembering that u is a constant for this integral, then $F(u)$ can be written as

$$F(u) = \sum_{j=1}^{n} H_j g(a_j)\cos(ua_j), \tag{3}$$

which can be readily identified as the formulation for an array of $2n$ discrete elements with spacings governed by the n values of a_j.

A more general approach to the array synthesis procedure involves the use of weight functions. In this form the mechanical quadrature[1] becomes

$$\int_a^b w(x)f(x)dx = \sum_{j=1}^{n} H_j f(a_j). \tag{4}$$

The n values of a_j are the roots of a polynomial orthogonal with respect to the weight function $w(x)$ on the interval (a, b). The n values of H_j can be determined by solving the set of linear equations generated by letting $f(x)=x^k$ for $k=0, 1, 2, \cdots, 2n-1$. In this form it is more efficient to identify the weight function as the aperture excitation $g(x)$. The only requirement on the weight function is that it does not vanish within the interval of integration and that it is integrable. Using this notation the Fourier integral in (1) becomes

$$F(u) = \int_{-1}^{1} g(y)\cos uy\,dy = \sum_{j=1}^{n} H_j \cos(ua_j), \tag{5}$$

where the transformation $ay=x$ has normalized the limits of integration. In this form the element positions are given by the n values of a_j and the current excitation coefficients are given by the n values of H_j.

One of the radiation patterns suggested by Unz[2] is the function

$$F(u) = e^{-(au/\pi)^2}\cos\left(\frac{2b}{\pi}u\right). \tag{6}$$

An advantage of this pattern is the independent control of beamwidth and sidelobe level. The constants a and b are determined from the design parameters for the beamwidth and sidelobe level and the relationships may be found in the original paper.

From (6) and the Fourier transform, $g(x)$ can be determined[3] and is given by

$$g(x) = K_1 e^{-K_2 x^2}\cosh(K_3 x), \tag{7}$$

where K_1, K_2, and K_3 are constants which have been evaluated. Using (1) and (7), $F(u)$ in (6) can be written as

$$F(u) = \frac{K_1}{\sqrt{K_2}}\int_{-\infty}^{\infty} e^{-y^2}\cosh\left(\frac{K_3}{\sqrt{K_2}}y\right)\cdot\cos\left(\frac{u}{\sqrt{K_2}}y\right)dy \tag{8}$$

by a suitable change of variable. Identifying e^{-y^2} as the weight function $w(y)$ in (4), the field pattern $F(u)$ can be expressed by

$$F(u) = K\sum_{j=1}^{n} H_j \cosh\left(\frac{K_3}{\sqrt{K_2}}a_j\right)\cdot\cos\left(\frac{u}{\sqrt{K_2}}a_j\right). \tag{9}$$

Kopal[1] has shown that the n values of a_j in this case, $w(y)=e^{-y^2}$, are the roots of the classic Hermite polynomial of degree n. The values of a_j and H_j are tabulated in Kopal.[1] Several examples using different beamwidths and sidelobe levels have been calculated and they give very good results, which will be published at a later date.

Another pattern used with excellent results is the approximation to the Dolph-Tchebycheff distribution given by Taylor.[4] In this case the orthogonal polynomials were generated by approximating the aperture excitation by a polynomial $w(x)$ and using this entire function as the weight function in the integral in (4). An alternative approach which has been used for the generation of the aperture distribution $g(x)$ employs the envelope of the standard tabulated[5] Dolph-Tchebycheff current distributions for uniformly-spaced arrays.

Programs have been written and used for several exponential and Dolph-Tchebycheff radiation patterns on an IBM 650 digital computer at the University of Kansas. In each case the results were within the prescribed design limits. Detailed analysis of the above method and more complete numerical results will be published at a later date.

In reference to an abstract of a technical paper[6] which was presented at the 1962 IRE International Convention, it appears that some work along similar lines is being done independently by Dr. Y. T. Lo at the University of Illinois.

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* Received March 15, 1962; revised manuscript received, March 22, 1962.

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Analysis of Nonlinear Sampled-Data Systems by Frequency Response Method*

Sampled-data control systems containing nonlinear element have been analyzed by both Tou with z-series method[1] and Kinnen with multirate sampling method.[2] Now in this correspondence another method by frequency response is introduced. Fig. 1 shows an error-sampled nonlinear system in which the nonlinear element is in series with $G(s)$. If the nonlinear element is temporarily left out, we have for the linear system

$$C^*(s) = \frac{G^*(s)}{1 + GH^*(s)} R^*(s). \tag{1}$$

Then for a complex input $r(t) = e^{j\omega t}$, the output of the system in terms of z transforms becomes

$$C(z) = z \frac{G(z)}{(z - e^{j\omega T})[1 + GH(z)]} \tag{2}$$

where T is the sampling period. Now if $G(z)/(z-e^{j\omega T})[1+GH(z)]$ can be expressed as

$$= c_0/(z - z_0) + c_1/(z - z_1) + c_2/(z - z_2) + \cdots + c_n/(z - z_n) \tag{3}$$

then with the aid of L'Hospital's Rule we can prove for stable systems where $z_r's < 1$ that at sampling instants

$$c(t)_{t\to\infty} = \frac{G^*(j\omega)}{1 + GH^*(j\omega)} e^{j\omega t}. \tag{4}$$

Now put the nonlinear block back to the system and let its output be represented by

$$y^*(e) = N_l e^*(t) + f^*(e) \tag{5}$$

where $e^*(t)$ is the input and N_l is a linear transfer function being either a constant or a complex quantity. Thus the nonlinear system can be replaced by an equivalent linear system (Fig. 2) in which its total output $c(t)$ is divided into two parts: $c_1(t)$, the system output when $f(e)=0$ and $\delta(t)$, the correction term due to $f(e)$. Thus according to (4)

$$c_1(t)_{t\to\infty} = \frac{N_l G^*(j)}{1 + N_l G^*(j)} \cos(t + \gamma_0) \tag{6}$$

when $r(t)=\cos t$ and $H(s)=1$. Suppose $e_1(t)$ represents the error when $f(e)=0$, then we have for $r(t) = \cos t$ (using subscript c)

$$e_1(t) = r(t) - c_1(t) = a_0 \cdot \cos(t + \theta_0) \tag{7}$$

and

$$e(t) = e_1(t) - \Delta_c(t). \tag{8}$$

Thus for a particular nonlinear element whose $f(e) = -e^3$, we have an approximate expression through (8)

$$f(e) = e_1^3 - 3e_1^2\Delta_c(t). \tag{9}$$

Now considering only the fundamental and third-harmonic terms of $\Delta_c(t)$ we assume

$$\Delta_c(t) = b_1 \cos(t + \theta_1) + b_3 \cos(3t + \theta_3). \tag{10}$$

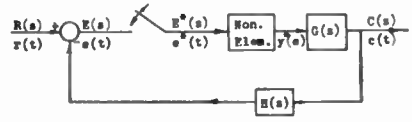


Fig. 1—A basic nonlinear sampled-data control system.

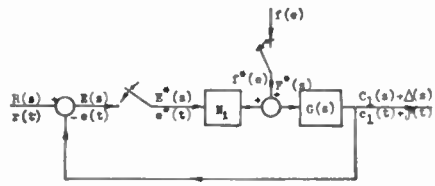


Fig. 2—The equivalent linear sampled-data control system.

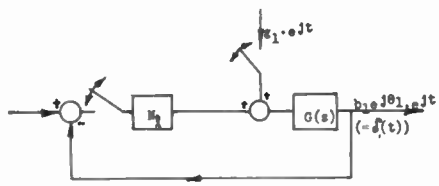


Fig. 3—The system considering $g_1 \cdot e^{jt}$ only.

Then substituting (7) and (10) into (9) gives

$$f_c(e) = -a_0^3 \cdot \cos^3(t + \theta_0) + 3a_0^2 \cdot \cos^2(t + \theta_0)\,\Delta_c(t) \tag{11}$$

$$= f_{1c}(t) + f_{3c}(t), \tag{12}$$

where $f_{1c}(t)$ and $f_{3c}(t)$ contain terms associated with $\cos t$ and $\cos 3t$ respectively. Now consider an input $r(t)=\sin t$ and let

$$\Delta_s(t) = b_1 \sin(t + \theta_1) + b_3 \sin(t + \theta_3), \tag{13}$$

similarly we obtain

$$f_s(e) = f_{1s}(t) + f_{3s}(t). \tag{14}$$

Then for a complex input $r(t) = e^{jt}$ we have

$$f(e) = f_c(e) + jf_s(e) \tag{15}$$

$$= g_1(b_1e^{j\theta_1}, b_3e^{j\theta_3})e^{jt} + g_3(b_1e^{j\theta_1}, b_3e^{j\theta_3})e^{j3t}. \tag{15a}$$

So according to (4) we finally obtain (Fig. 3)

$$b_1e^{j\theta_1} \cdot e^{jt} = \frac{G^*(j)}{1 + N_l G^*(j)} (g_1 \cdot e^{jt}). \tag{16}$$

Now write

$$b_1e^{j\theta_1} = R_1 + jI_1$$
$$b_3e^{j\theta_3} = R_3 + jI_3. \tag{17}$$

Then by equating respectively the real and imaginary terms on both sides of (16) we have

$$k_1R_1 + k_2R_3 + k_3I_1 + k_4I_3 = k_5$$
$$n_1R_1 + n_2R_3 + n_3I_1 + n_4I_3 = n_5. \tag{18}$$

where all k_n's and n_n's are constants. Similarly, when g_3 is considered, another set of equations can be drawn for the third-harmonic components. Thus b_1, b_3, θ_1 and θ_3 can be solved and both $\delta(t)$ and $c(t)$ can be found. The $c(t)$ thus obtained is valid only at sampling instants.

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* Received March 12, 1962; revised manuscript received March 26, 1962.

[1] J. Tou, "Analysis of sampled-data systems containing nonlinear element," PROC. IRE (*Correspondence*), vol. 46, p. 915; May, 1958.

[2] E. Kinnen, "An Analysis of Nonlinear Sampled-Data Control Systems," Ph.D. dissertation, Purdue University, Lafayette, Ind.; 1958.

The Microwave Diagnosis of a Column of Ionized Gas*

We shall present a method of determining the electron distribution in a dilute column of ionized gas. The term "dilute" means here that the electron density is small enough that each electron behaves as an independent scatterer and that the frequency ν of collisions between electrons and neutral particles is much smaller than the frequency ω of the incident wave. These conditions ensure that the gas column may, to a good approximation, be considered to be uniformly irradiated by the incident wave, which we take to be plane, transverse-magnetic and incident normal to the column's axis.

The signal scattered from the column is assumed to be received by a receiver situated at a distance R from the axis and at an angle α from the direction of the incident wave vector k. The electron density N is independent of the coordinate z along the axis and of the azimuth about the axis. Thus, $N = N(r)$, where r is the radial distance from the axis. The column has a radius $a \ll R$ and is assumed to be sufficiently long that the scattered signal at the receiver is virtually that which is scattered by an infinitely long column.

Under the above conditions, Eshleman[1] has shown that the back-scattered electric field ($\alpha = \pi$) is approximately

$$E_z^s|_\pi = A\,|E_z^i|\,\exp\left[i\left(kR - \frac{\pi}{4} - \omega t\right)\right] \cdot \left(\frac{2}{\pi kR}\right)^{1/2} \int_0^a N(r) J_0(2kr)\,dr, \tag{1}$$

where $A = -i\mu_0 e^2\pi/2m_e$, $|E_z^i|$ is the absolute value of the incident field and J_0 is the zeroth-order Bessel function; μ_0, e, and m_e are the permeability of space and the charge and mass of an electron, respectively. We define the scattering coefficient ρ_α by the relation

$$E_z^s|_\alpha = |E_z^i| \left(\frac{2}{\pi kR}\right)^{1/2} \rho_\alpha; \tag{2}$$

therefore, the back-scattering coefficient is

$$\rho_\pi(k) = A \int_0^a N(r) J_0(2kr)\,r\,dr. \tag{3}$$

It is easy to extend (1) to the forward scattering case,[2] $\alpha = 0$. The result is

$$E_z^s|_0 = A\,|E_z^i|\,\exp\left[i\left(kR - \frac{\pi}{4} - \omega t\right)\right] \cdot \left(\frac{2}{\pi kR}\right)^{1/2} q/2\pi, \tag{4}$$

where q is the number of electrons per unit length.

$$2\pi \int_0^a N r\,dr.$$

Thus

$$\rho_0 = Aq/2\pi. \tag{5}$$

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[1] V. R. Eshleman, "The Mechanism of Radio Reflections from Meteoric Ionization," Electronics Res. Lab., Stanford University, Calif., Tech. Rept. No. 49; July, 1952.

[2] J. R. Barthel, "The Microwave Diagnosis of a Column of Ionized Gas," General Dynamics/Convair, San Diego, Calif., Rept. No. ZPh-096; May, 1961.

In the narrow column (or long wave) limit, $ka \ll 1$, we find $\rho_0 = \rho_\pi$.

If we determine $\rho_\pi(k)$ experimentally, we may find $N(r)$ from (3) by applying the Fourier-Bessel integral,[3] which states

$$f(z) = \int_0^\infty J_m(kz)kdk \cdot \int_0^\infty f(\xi)J_m(k\xi)\xi d\xi \quad (m > -\tfrac{1}{2}). \quad (6)$$

By using (3) in (6), we find

$$N(r) = (4/A)\int_0^\infty J_0(2kr)\rho_\pi(k)kdk. \quad (7)$$

Eq. (7) holds only in a dilute column. The values of k at which this model is inapplicable must be sufficiently limited that the $\rho(k)$ at those values make a negligible contribution to the integral in (7).

At any point in the column, the dielectric constant

$$K = 1 - \omega_p^2/\omega^2 = 1 - Ne^2/m_e\epsilon_0\omega^2$$

becomes negative when $\omega < \omega_p(r)$. In a region where $K<0$, the incident wave attenuates, contrary to our assumption. Let δ be the distance over which the wave attenuates by $1/e$ (skin depth). If $\omega_p(\max)/c = k'$, δ is real for $k<k'$. As $k \to 0$,

$$\delta \to \delta_{\min} = c/\omega_p(\max) = 1/k'. \quad (8)$$

Now suppose $\delta_{\min} \gg a$, or $ka' \ll 1$; then attenuation can be neglected even at very low frequencies. This condition implies that

$$m_e\epsilon_0c^2/e^2 \gg N_{\max}a^2 \quad (>q/\pi). \quad (9)$$

Since $m_e\epsilon_0c^2/e^2 = 10^{14}/\pi$ electrons per meter, the condition $q \ll 10^{14}/m$ is a necessary (but not sufficient) condition for $\delta_{\min} \gg a$. $\delta_{\min} \gg a$ is sufficient for the validity of (7).

We have used k to mean vacuum wave number ω/c. In the plasma the wave number is

$$\bar{k} = \frac{\omega}{c}\sqrt{1 - \frac{\omega_p^2}{\omega^2}}.$$

An additional requirement for the validity of (7) is, therefore, that the contribution to the integral is negligible for values of k for which $(k-\bar{k})a$ is of the order unity or greater. If $\omega \geq \omega_p$ (max),

$$(k - \bar{k})a = k'a\left(\frac{\omega}{\omega_p} - \sqrt{\frac{\omega^2}{\omega_p^2} - 1}\right) \leq k'2.$$

Therefore if $k'a \ll 1$, significant phase errors are confined to the $\omega < \omega_p$, which does not contribute significantly to (7) if condition (9) is met. If $k'a$ is of the order unity or greater, we must in general take account of the variation of $\bar{k}$ in the gas column, although for certain electron distributions the contribution to (7) of the k region is which there are phase effects may be small.

It is readily seen that collisions have no effect if $\nu \ll k'c$.

As an independent check on the applicability of the dilute column model in a given case, we may use the forward scattered signal (if available) in (4) to determine q. If our model is valid, (4) should give identical values of q for each $k>k'$.

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[3] P. Morse and H. Feshback, "Methods of Theoretical Physics," McGraw-Hill Book Co. Inc., New York, N. Y., p. 766; 1953.

Causality*

Causality as invoked by Deutsch is axiomatic in nature.[1] Then there are several rather fundamental limitations that may be imposed on this monobloc computer:

1) There is no known logical system, whose consistency is proved, that can be used as a basis for the computer.[2]
2) The mandatory uncertainties in the computer's knowledge of the monobloc's structure must be such that no future event is thereby affected. Heisenberg's Principle of Quantum Indeterminancy may be invoked during the time that the computer is acquiring data about the formation of the monobloc. The simultaneous invocation of a monobloc and a supercomputer would be pressing a *deus ex machina* too far.

One should more properly infer that, given our local solar system, one might predict the ineluctable evolution of life. In any event, predictability does not necessarily alter the operative value of such concepts as free will and human creativity. If there is no physically realizable means of executing such predictability, these actions and events remain unpredictable, even to ourselves.

This conversely means that the boundary level of realizable predictability is worthy of investigation and is capable of some quantification.

Although some men seem to be intelligent, it seems too early to ascribe this quality to man. It appears that social activity has a generic propensity to sink to the lowest common denominator. On the other hand it appears incredible that it is impossible to raise the level of this lowest common denominator to at least a survival level.

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* Received March 14, 1962.
[1] S. Deutch, "Causality," Proc. IRE (*Correspondence*), vol. 50, p. 222; February, 1962.
[2] E. Nagel and J. R. Newman, "The World of Mathematics," Simon and Schuster, Inc., New York, N. Y., vol. 3, pt. XI-4; 1956. This is a discussion of K. Goedel's proof from "On Formally Undecidable Propositions of Principia Mathematica and Related Systems."

Comments on "Symmetrical RC Distributed Networks"*

Hellstrom[1] described a class of RC distributed lines where the specific series resistance $r(x)$, was varied along the line, and the ratio $r(x)/c(x)$ was kept constant. It is interesting to note that this class is essentially identical with the uniform line case if only the measuring scale dx is chosen correctly. This becomes evident if one performs the following scale transformation

$$du = \frac{r(x)}{r_0}dx = f(x)dx \quad (1)$$

in the differential equation of the symmetrical network

$$\frac{d^2V}{dx^2} - \frac{f'}{f}\frac{dV}{dx} - j\omega r_0c_0f^2V = 0. \quad (2)$$

The result is recognized as the differential equation governing the uniform line:

$$\frac{d^2V}{du^2} - j\omega r_0c_0V = 0. \quad (3)$$

Consequently the chain parameters of the symmetrical line can be directly obtained from those of the uniform line by the same transformation (1).

On the other hand, the case where the product $r(x)c(x)$ remains constant leads to possible realization of network functions appreciably different from those obtainable with uniform lines. This class of RC distributed lines is governed by

$$\frac{d^2V}{dx^2} - \frac{f'}{f}\frac{dV}{dx} - j\omega r_0c_0V = 0. \quad (4)$$

Unlike (2) only special cases of $f(x)$ will permit the analytical solution of (4). The exponential line is among the most interesting cases of this class. Considering a distributed RC ladder where the series resistance varies exponentially with the distance

$$r(x) = r_0e^{hx} \quad (5)$$

and the shunt capacitance correspondingly,

$$c(x) = c_0e^{-hx}. \quad (6)$$

Eq. (4) then becomes

$$\frac{d^2V}{dx^2} - h\frac{dV}{dx} - j\omega r_0c_0V = 0. \quad (7)$$

Assuming a solution in the form of $\exp(\gamma x)$, the possible values for γ are found,

$$\gamma_{1:2} = \frac{h}{2}\left(1 \pm \sqrt{1 + 4j\frac{\omega c_0r_0}{h^2}}\right) \quad (8)$$

and after defining

$$\gamma_1 = -\alpha_1 - j\beta_1; \qquad \gamma_2 = \alpha_2 + j\beta_2;$$
$$\theta = \tan^{-1}\frac{4\omega r_0c_0}{h^2},$$

one finds that

$$\beta_1 = \beta_2 = \frac{h}{2}\left(\frac{1}{\sqrt{1 - \tan^2\theta/2}} - 1\right) = \beta$$
$$\alpha_1 = \frac{h}{2}\left(\frac{1}{\sqrt{1 - \tan^2\theta/2}} - 1\right)$$
$$\alpha_2 = \frac{h}{2}\left(\frac{1}{\sqrt{1 - \tan^2\theta/2}} + 1\right). \quad (9)$$

γ_1 and γ_2 are the voltage propagation constants in the positive and negative directions of x. It is seen from (9) that the attenuation constant in the positive direction (α_1) is many times smaller than the attenuation constant in the negative direction (α_2), while the phase constants β_1 and β_2 are equal.

The elimination of θ from (9) results in the following relations between α, β and h:

$$\beta = \sqrt{\alpha_1(\alpha_1 + h)} = \sqrt{\alpha_2(\alpha_2 - h)}. \quad (10)$$

* Received March 19, 1962.
[1] M. J. Hellstrom, Proc. IRE (*Correspondence*), vol. 50, pp. 97–98; January, 1962.

Eq. (10) is geometrically represented by the circle of Fig. 1 which is very useful for the design of exponential RC lines. It gives a clear picture of the phase shift that one obtains within the practical limitations of line loss and physical size. For prescribed values of α_1 and h one can find the values of β and α_2 as indicated on the circle. From the uniform line ($h=0$) one always gets $\alpha=\beta$, while from the exponential line a larger phase shift can be obtained for the same value of attenuaton allowed. This effect is greatest for high values of h/α_1.

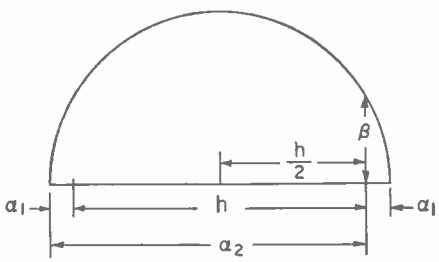


Fig. 1—Geometrical representation of the relations between α_1, α_2, β and h.

If the following approximation is made:

$$e^{-\alpha_2 z} \ll e^{+\alpha_1 z},$$

then the complete solution can be put in a form of a simple chain matrix,

$$\begin{bmatrix} V(0) \\ I(0) \end{bmatrix} = \frac{e^{-\gamma_1 z}}{\gamma_2 - \gamma_1} \begin{bmatrix} \gamma_2 & r_0 \\ j\omega c_0 e^{hz} & -\gamma_1 e^{hz} \end{bmatrix} \begin{bmatrix} V(x) \\ I(x) \end{bmatrix}$$

$$= \frac{\sqrt{\cos\theta}\cdot e^{\alpha_1 z + j(\beta z - \theta/2)}}{h} \cdot \begin{bmatrix} \gamma_2 & r_0 \\ \jmath\omega c(x) \dfrac{r(x)}{r_0}(\alpha_1 + j\beta) & \end{bmatrix} \cdot \begin{bmatrix} V(x) \\ I(x) \end{bmatrix}.$$

It is also important to mention that lines of constant RC product can be easily built in the form of resistive thin films of variable width.

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Author's Reply[2]

The above comments afford the opportunity to add a point to my note[1] which presents the requirements on the resistance and capacitance density functions such that the network will be symmetrical *at all lengths*. It is possible to construct networks, not in this class, which are symmetrical at a specific network length, but not at all lengths.

Note that although the performance of the symmetrical network is identical to that of the uniform network (and it is in this sense that the proportional network defines a group of equivalent networks) the network itself is *not identical* since its r and c functions may have unlimited forms, subject to the proportionality requirement.

The transformation suggested by Pearl,

[2] Received, March 26, 1962.

or its equivalent integral form

$$u = \int_0^x f(x)dx,$$

is a special case of one given by Liouville,[3] *i.e.*,

$$z = \int_0^x \sqrt{rc}\, dx.$$

Its usefulness in changing a general nonuniform transmission-line problem (or distributed RC network) into one with constant propagation factor (or RC product) has been noted recently.[4] This same transformation can be used to assist in defining groups of equivalent networks, *i.e.*, networks with the same terminal performance.[5] The proportional network constitutes such a group with the uniform network being a single member.

The case, RC = constant, is not a "Class" of networks in the sense that it does not prescribe anything about network performance. Mr. Pearl's differential equation, (4) is no closer to solution due to the choice RC = constant than the original network equation because the differential equation of any network can always be put into form (4).

The exact exponential-network open-circuit impedance parameters and a discussion of the propagation factor, incidentally, have both been published.[6,7]

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[3] E. Kamke, "Differentialgleichungen, Lösungsmethoden und Lösungen," Edwards Bros., Inc., Ann Arbor, Mich., p. 261; 1945.
[4] I. Jacobs, "A generalization of the exponential transmission line," Proc. IRE, vol. 47, pp. 97–98; January, 1959.
[5] M. G. Hellstrom, "Equivalent distributed RC networks or transmission lines," IRE Trans. on Circuit Theory Vol.-CT-9, pp. 247–251; September, 1962.
[6] B. L. H. Wilson and R. B. Wilson, "Shaping of distributed RC networks," Proc. IRE, vol. 49, pp. 1330–1331; August, 1961.
[7] W. A. Edson, "Tapered distributed RC lines for phase shift oscillators," Proc. IRE, vol. 49, pp. 1021–1024; June, 1961.

Capacitance Calculations for Several Simple Two-Dimensional Geometries*

A number of situations arise in solid-state device research in which capacitance estimates are required for two-dimensional configurations where edge effects are not negligible, and for which confinement of the displacement to the dielectric medium because of its high permittivity prohibits the use of handbook formulas.[1] Typical of these are the complementary configurations of Fig. 1. In the figure, the constant potential segments of the boundaries are indicated by solid lines, while the dotted lines designate segments on which the normal derivative of the potential vanishes.

Fig. 1(a) represents the right half of an infinite sheet of dielectric of thickness b, having a completely electroded lower surface ($\phi=0$), and bearing an infinitely long strip electrode ($\phi=V$) of width $2a$ on the upper surface. The boundary conditions $\phi_n=0$ result from considerations of symmetry at the left, and of confinement of the displacement to the dielectric sheet at the upper right.

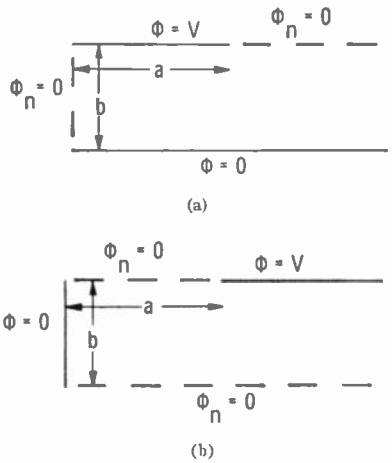


Fig. 1—Boundary conditions.

By placing a z-plane origin at the lower left corner of the figure, one can employ the Schwarz-Christoffel transformation

$$\tanh\frac{\pi z}{2b} = \mathrm{sn}\,(w, k)$$

to map the dielectric region into a K by K' rectangular region in the w plane, where K and K' are complete elliptic integrals of the first kind for moduli $k=\tanh \pi a/2b$ and $k'=(1-k^2)^{1/2}$, respectively.[2] The sides of the w-plane rectangle correspond to the four boundary segments of the z-plane figure, from which it follows that the geometric (dimensionless) capacitance c_λ for Fig. 1(a) is K/K'. Because of the complementary nature of the geometries of Fig. 1(a) and (b), the geometric capacitance of the latter is K'/K. These, of course, must be multiplied by the permittivity of the dielectric in question in order to obtain the desired capacitance per unit length in usable units:

$$C_\lambda = \epsilon c_\lambda = \mathrm{K}\epsilon_0 c_\lambda,$$

where K is the dielectric constant.

Fig. 2 summarizes the above results, as well as those for other configurations, which are solvable by a similar, although somewhat more involved transformation.

Fortunately, in the application of these formulas some rather good approximations are available. These are based on the fact that, for $k\to1$, $K'\to\pi/2$ and $K\to\ln(4/k')$. Thus, in the case of $c_{\lambda 1}$ for $a\gg b$,

$$c_{\lambda 1} \approx (a + 0.44b)/b$$

* Received March 19, 1962.
[1] See, for example, "American Institute of Physics Handbook," McGraw-Hill Book Co., Inc., New York, N. Y., sect. 5b; 1957.
[2] A helpful monograph on this subject is F. Bowman, "Introduction to Elliptic Functions with Applications," John Wiley and Sons, Inc., New York, N. Y.; 1953.

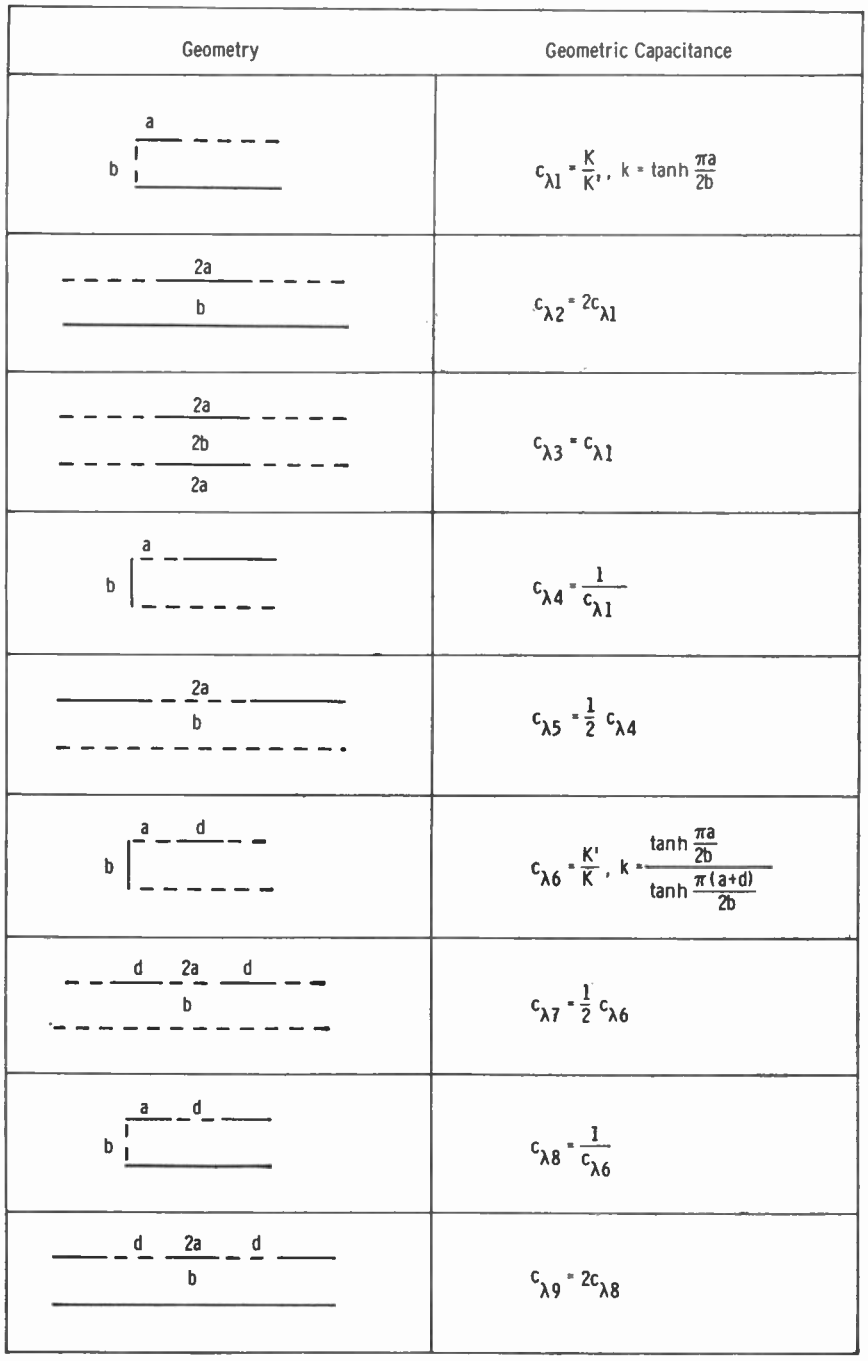


Fig. 2—Capacitances of various geometries.

where 0.44 is $(2 \ln 2)/\pi$. This result is applicable to all of the cases, $c_{\lambda 2}, \cdots, c_{\lambda 5}$, dependent on $c_{\lambda 1}$. Furthermore, although the approximation is derived for $a \gg b$, it is accurate within less than one per cent for $a = b$.

Similarly, $c_{\lambda 6}$ for $a \gg b$ becomes

$$c_{\lambda 6} \approx b/(a + 0.44 + \Delta)$$

$$\Delta = \frac{1}{\pi} \ln \left(\frac{1 + D}{2D} \right)$$

$$D = \tanh \frac{\pi d}{2b}.$$

This approximation was tested for $a = b$, $d = 0.1b$ and found to be accurate to within less than 1 per cent. The term Δ has the value 0.44 for $d = 0.091b$, and a value 10 times smaller for $d = 0.65b$. For this reason, it is practical in most cases for which $d \geq b$ to use the simpler expression for $c_{\lambda 4}$ $(d = \infty)$, rather than $c_{\lambda 6}$.

The derivation of the results herein assumes homogeneity and isotropy throughout a dielectric sheet of infinite extent. In most practical applications, uniformity is satisfied and neglect of the requirement on infinite extent has been found to be of little consequence, probably because significant edge-effect fields extend on the order of only one sheet thickness past the electrode edges (as indicated by the approximation formulas). Discrepancies between calculated and measured values of capacitance have been noted, however, for very high dielectric-constant ferroelectric ceramics with narrow electrodes. Such materials are known to be anisotropic and to possess surface layers having permittivities quite different from those of the bulk. Even in these cases, calculations of capacitance made by the methods described and utilizing the bulk dielectric constant have been found useful for estimating purposes.

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Synchronization Error*

That there will be some error when a receiver synchronizes with an incoming noisy signal is intuitively obvious. The magnitude of the error in the case of ideal sync signal and receiver processing is indicated by a simple application of information theory. A great deal of insight into the nature of the problem can be obtained from the analysis.

Let the time period over which the receiver expects to receive the synchronizing signal (whose waveform is exactly known) be L, and let the total energy in the signal be U. If the bandwidth of the signal is B, and the noise is white with density N_0, the noise power received with the signal is BN_0. Shannon's theorem on the capacity of a channel states that the maximum amount of information which can be received is

$$I = LB \ln (1 + U/LBN_0) \text{ natural bits.}$$

It is apparent that I increases with B. If B is very large, the argument of the logarithm is nearly unity, and a good approximation for fixed U/N_0 is

$$\ln (1 + U/LBN_0) \doteq U/LBN_0, \quad B \gg U/LN_0.$$

In the limit

$$I = U/N_0 \text{ natural bits.}$$

This is the maximum amount of information which can be received in time L when the signal has energy U and the noise density is N_0. Furthermore, according to Shannon's results, if the proper signal and reception technique is used, this information is perfectly reliable, *i.e.*, there is zero probability of error.

Now let the information contained in the signal be an indication of the time at which the signal is received. Exactly $n = \exp(U/N_0)$ states can be specified by the information. If the interval L is divided into n equal subintervals,[1] each message state can be assigned to correspond to reception of the signal within one of the subintervals. Thus synchronization occurs at some time within the subinterval corresponding to the signal. If the mid-point of the subinterval is used, the maximum synchronization error e for the

* Received March 12, 1962.

[1] The problem is only academic if n is not an integer.

ideal system is then one-half the width of the subinterval.

$$e = (L/2) \exp(-U/N_0).$$

This result is remarkable in the same way that Shannon's result is remarkable: There is no probability distribution associated with the error; an absolutely firm bound on the ideal error has been achieved. It is also inadequate in the same sense that Shannon's result is. It does not indicate the practical manner by which this error rate can be achieved.

As a corollary to the result note that the same final error occurs whether two synchronizing signals, each with energy U, are sent sequentially with the second, reducing the error left by the first, or whether one signal with energy $2U$ is sent.

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On Visual Perception and Retinal Motions*

The experimental evidence concerning the basis of visual perception presents a confusing turmoil of data to the designer of pattern recognition systems. The critical question seems to be: with what capacities should the system be "born" and how much of its capability is either required or, better, left to be "learned"? There seems some demonstration of predetermined or innate ability to perceive form, texture, and depth—at least to some nontrivial extent—in human, cat, and goat infants.[1,2] Other evidence suggests an initial period of nervous system "plasticitiy" subsequent to which a "hardening" sets in and precludes development of important properties if there has been sensory deprivation in this initial time of growth; if this is true for man, as it seems to be true for monkey and cat, it provides some small reinforcement of the notion that there is a good deal more specification in the sensory system (visual) than is suggested by such statistical principles as are embodied in the Perceptron. At least it reveals that later perception and/or learning still arises in neural networks with largely fixed, nonstatistical properties and organization. While this would not rule out the importance or desirability of initially "freely-connected" (Rosenblatt) circuitry as a possible context for processes of self-organization, it does raise the question of the theoretical need of statistical procedures in perception or learning if determinate networks have the same abilities.

The importance of retinal motions[1-4] in form recognition has been underestimated. Some of these motions are oscillatory, do not yield to voluntary control, and are not consciously perceived save when the *effects* of their having been stabilized with respect to stimulus objects in experimental situations reveal their presence to subject or experimenter.

Let us assume for a moment an array of receptors containing the representation of an object-pattern as a distribution of receptor excitation intensities, and stipulate that there is no relative motion of any sort between this retinal system and the external stimulus. It is a simple task to project only the contours in the retinal image onto another array A_1 of neurons. A well-known difficulty is involved here in having the system recognize figures which have not been somehow position "normalized," that is, brought into some standardizing region of A_1. So long as the discrimination process is motion independent and depends only on the excitation of special, more-or-less unique sets of neurons, this normalization procedure, in one form or another, is mandatory; it is implicit even in the procedure of repeatedly presenting the object in different positions in the visual field together with some suitable type of response coercion which we may speak of as "learning." The particulars here are inconsequential; rather it is important that a) the system must rely on the classification of the whole pattern presented as a unit entity and not, for example, on the "straightness" or "curviness," etc., of the various regions in the pattern of contours, and b), if a) is true, the spatial structure (geometry) of the stimulus is lost as it must be recoded into a new geometry and energy distribution for internal system representation. By considering only these last two items, a) and b), it may be seen that systems embodying such characteristics can claim little similarity either to structural or functional processes in the visual sensory system of man.

We can begin to demonstrate our assertion most simply and convincingly by contemplating the consequences of item b) above; all we need do is require such a system to "draw," for example, its interpretation of the circle it has been able to perceive (discriminate from all other geometrical entities). Unless it can perform a quite difficult inverse transform, the system's output will be its internal representation of a circle which, to us (man), will be a meaningless assortment of spatial patterns and energy amplitudes. Since man can reproduce his perceptions with rather good accuracy, it seems a safe assumption that no such recoding as has been described occurs in man's sensory system, unless one can postulate with any confidence this unwieldy inverse transform. As for item a), the forms customarily perceived by man are not the same perceived when the retinal oscillations are stabilized[3,4] relative to the stimulus, *i.e.*, there is no perception as we "know" it without these retinal motions under control of the autonomic nervous system.

One can speculate that there is some provision in the visual system for mapping the image on A_1 onto another array A_2 in such a way that the image geometry is preserved to a degree depending on the complexity of the pattern; and that, in general, in a world largely without "simple" patterns, this mapping fails to preserve certain portions of the image on A_1. Therefore, without some motion of the retinal system, there would be some regions of the image that seem to "fade" out in passage from A_1 to A_2. The phenomenon of "fading" and partitioning has been observed experimentally[3,4] with the feeling that those parts of the image that are preserved after retinal system-object image stabilization, are "meaningful" units of perception. Be this as it may, it would appear that the only principle compatible with the preservation of the images of the external world after passage through certain neural network processes and normalization to eliminate variations with size and position, must depend upon the discrimination of such local properties as straightness and curviness. Cognition of pattern then becomes a matter of the conditioning of the association between certain nervous system activity connected with the control of retinal motions (which may seem currently to be a substratum of "noise" in neurophysiological studies) and the geometry-preserved patterns of activity representing stimulus objects.

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Demodulation Effect of an Envelope Detector at Low Signal-to-Noise Ratios*

An envelope detector suffers an effective loss of percentage modulation when the signal-to-noise ratio (SNR) approaches unity. The mechanism responsible is discussed, and the degradation of the percentage modulation is computed for different SNR. A method for partially correcting this condition is suggested.

If we think of an envelope detector as a synchronous detector (which in fact it is), then it is readily seen that noise can interfere with the detection process at low SNR. The synchronous feature comes about from the fact that the radio-frequency carrier switches the diode off during negative half cycles and on during positive half cycles. Noise amplitude peaks comparable to the signal amplitude can interfere in two ways. The diode may be switched on during a negative-signal half cycle, or the noise peaks may switch the diode off during a positive-

* Received by the IRE, March 6, 1962.

[1] R. L. Frantz, "The origin of form perception," *Sci. Am.*, vol. 204, p. 66; May, 1961.

[2] E. J. Gibson, "The visual cliff," *Sci. Am.*, vol. 202, p. 64; April, 1960.

[3] R. Ditchburn and D. Fender, "The stabilized retinal image," *Optica Acta*, vol. 2, p. 128.

[4] R. Pritchard, W. Heron, and D. Hebb, "Visual perception approached by the method of stabilized images," *Can. J. Psychol.*, vol. 14, p. 6.

* Received March 13, 1962; revised manuscript received April 3, 1962.

signal half cycle. At large SNR there is little danger of this occurring, but at low ratios (of the order of magnitude of one to one) it is apparent that the normal detection process will be interfered with. The over-all effect is an apparent reduction in the percentage of modulation of the amplitude-modulated radio-frequency signal.

The reduction in the percentage of modulation may be calculated for different SNR where the noise is Gaussian in a narrow band. We are primarily interested in SNR in the neighborhood of one.

Bennett[1] has shown that Gaussian noise in a narrow band has a Rayleigh distribution, and for a noise power of σ^2 watts in the band combined with a sinusoidal carrier signal of amplitude A, the amplitude distribution is given as

$$p(v) = (v/\sigma^2) \cdot \{\exp - [(v^2 + A^2)/2\sigma^2]\} I_0(Av/\sigma^2) \quad \text{for } v \geq 0$$
$$p(v) = 0 \quad \text{for } v < 0. \quad (1)$$

The signal or carrier is unmodulated, and we will determine the output of the detector for this case first. We will then consider modulation as simply varying the amplitude of the carrier, which is actually the case. By considering the amplitude as positive only (or envelope), we essentially introduce the rectification action of a linear detector. We can account for the negative swings by considering a full-wave linear detector. The average value out of the detector is then seen to be

$$\bar{v} = \int_0^\infty v p(v) dv \quad (2)$$

$$\bar{v} = (1/\sigma^2)[\exp - (A^2/2\sigma^2)] \cdot \int_0^\infty v^2 [\exp - (v^2/2\sigma^2)] I_0(Av/\sigma^2) dv. \quad (3)$$

The integral may be found in Foster and Campbell,[2] and after some manipulation is seen to yield

$$\bar{v} = (\sqrt{2\pi}\,\sigma/2)[\exp - (S/N)] \cdot {}_1F_1(3/2; 1; S/N) \quad (4)$$

where

$$S/N = A^2/2\sigma^2$$

the unmodulated SNR. The function ${}_1F_1$ is Kummer's confluent hypergeometric series, and is plotted for these indices in Jahnke and Emde[3] for the range of argument of interest.

This plot is duplicated in Fig. 1 together with the exponential term and the product of both the exponential term and hypergeometric term. Over most of the range of interest, the product of these two is a straight line and this approximation will be made. Eq. (4) then may be written as

$$\bar{v} = \sqrt{2\pi}\,(\sigma/2)[0.4(S/N) + 1]. \quad (5)$$

If we consider the modulation as a slow variation of the carrier, then we need only compute the output as given by (5) for the maximum and minimum of the carrier (assuming sinusoidal modulation). We may then write

$$A^2(1+m)^2/2\sigma^2 = S_{\max}/N = S(1+m)^2/N \quad (6a)$$

and

$$A^2(1-m)^2/2\sigma^2 = S_{\min}/N = S(1-m)^2/N. \quad (6b)$$

The dc output for the extremes of carrier output is then

$$\bar{v}_{\max} = \sqrt{2\pi}\,(\sigma/2)[0.4(S/N)(1+m)^2 + 1] \quad (7a)$$
$$\bar{v}_{\min} = \sqrt{2\pi}\,(\sigma/2)[0.4(S/N)(1-m)^2 + 1]. \quad (7b)$$

If we now define

$$m_{\text{out}} = (\bar{v}_{\max} - \bar{v}_{\min})/(\bar{v}_{\max} + \bar{v}_{\min}) \quad (8)$$

consistent with the concept of modulation index we find upon substituting (7) into (8)

$$m_{\text{out}} = 2m(S/N)/[(S/N)(1 + m^2) + 2.5]. \quad (9)$$

A plot of this final result is shown in Fig. 2 for SNR up to 3. Because of the straight-line approximation made, a slight error (less than 10 per cent) may be encountered at large modulation indices near 1. Using a different approach Ragazzini[4] has demonstrated the demodulation effect.

If we present the output of the envelope detector on an oscilloscope and slowly increase the noise level into the detector, we observe that the demodulated output begins indeed to shrink in size as the noise becomes comparable to the signal. The output SNR, therefore, deteriorates at low input SNR for two reasons: the output noise increases, and the output signal decreases. Over 3-db improvement could be realized if carrier phase information would be available to cut off the diode during the half cycle when it is not conducting. In addition to eliminating the noise for half the carrier cycle, the demodulation effect of the above is reduced. Whether or not the addition of a phase-locked loop is justified to supply the carrier phase information will probably be contingent upon other factors.

ACKNOWLEDGMENT

Thanks are due Dr. E. Pflumm for his valuable assistance in reducing the mathematics.

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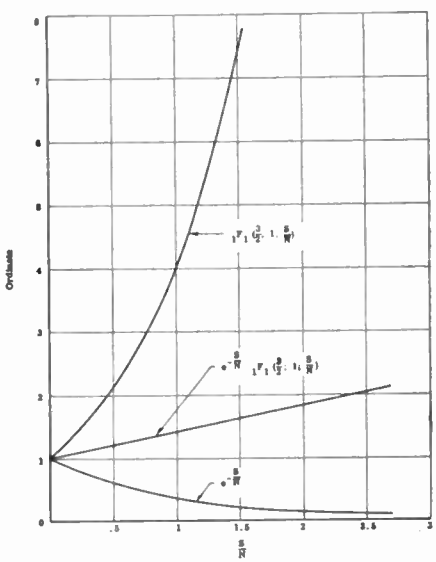


Fig. 1.

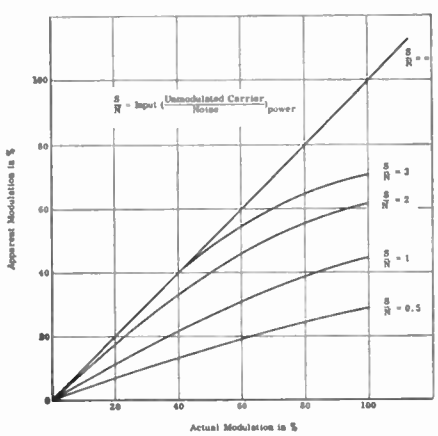


Fig. 2.

[1] W. R. Bennett, "Methods of solving noise problems," PROC. IRE, vol. 44, pp. 609–638; May, 1956.
[2] R. M. Foster and G. A. Campbell, "Fourier Integrals for Practical Applications," D. Van Nostrand Co., Inc., New York, N. Y.; 1948.
[3] E. Jahnke and F. Emde, "Tables of Functions," Dover Publications, New York, N. Y., p. 278; 1945.
[4] J. R. Ragazzini, "The effects of fluctuation voltage on the linear detector," PROC. IRE, vol. 30, pp. 277–288; June, 1942.

Cerenkov Radiation as Leaky Surface Waves*

An argument based on physical considerations [1] points out that Cerenkov radiation can be considered as the equivalent of reflection of an electromagnetic wave approaching a dielectric vacuum interface from infinity in a semi-infinite dielectric. The reflection of this incident wave results in leaky surface waves on the dielectric side of the boundary, and satisfies the Cerenkov condition, *i.e.*, the angle θ between the normal to the wave front and the boundary is $\cos\theta = (\beta\sqrt{\mu_r\epsilon_r})^{-1}$. In the absence of the incident wave, a beam of charged particles moving along the interface on the vacuum side supplies the equivalent fields which, under proper conditions, correspond to the leaky surface waves of the structure. That these waves are indeed the natural modes of the system can be shown as follows:

Consider a semi-infinite lossless dielectric μ, ϵ, with a cylindrical hole of radius r_0. In the hole $\mu = \mu_0$, $\epsilon = \epsilon_0$. Assuming propagation of the form $\exp[i(\omega t - kz)]$, simultaneous solution of the two Maxwell curl equations can be reduced to the Helmholtz scalar equation

$$(\nabla_t^2 + h^2)\begin{bmatrix} E_z \\ H_z \end{bmatrix} = 0 \quad (1)$$

* Received March 12, 1962. This work was supported by Rome Air Development Center under contract number AF30(602)2151.

where h^2 is the eigenvalue of ∇_t^2—the Laplacian operating on transverse coordinates.

This equation has to be satisfied in the hole (subscript 0) and in the dielectric (subscript 1). It is well known that all the field components can be obtained from the electric- and magnetic-field components in the direction of propagation [2].

On the vacuum side of the boundary

$$h_0^2 = k_0^2 - k^2 \tag{2}$$

where k = propagation constant, and $k_0^2 = \omega^2\mu_0\epsilon_0$.

In the dielectric

$$h_1^2 = k_1^2 - k^2 \tag{3}$$

with $k_1^2 = \omega^2\mu_1\epsilon_1$.

The simultaneous solution of (2) and (3) yields

$$\frac{\omega^2}{c^2} = \frac{h_1^2 - h_0^2}{\mu_r\epsilon_r - 1} \tag{4}$$

$$k^2 = \frac{h_1^2 - \mu_r\epsilon_r h_0^2}{\mu_r\epsilon_r - 1} \tag{5}$$

for the frequency and propagation constant dependence on the eigenvalues. Here $c^2\mu_0\epsilon_0 = 1$, $\mu_r\mu_0 = \mu_1$, and $\epsilon_r\epsilon_0 = \epsilon_1$. Thus far, the equations are quite general, and the only restriction is that a coordinate system be used in which the Helmholtz equation for the z-component is separable [3]. We can easily satisfy this restriction by using a circular coordinate system, which is most convenient for the geometry under consideration.

Several points are clear: 1) since the selected structure is lossless, the propagation constant cannot be complex [4]; 2) the radian frequency (ω) and the phase velocity must be real for propagation to take place, implying that $k^2 > 0$; 3) if $h_1^2 > 0$, the fields in the dielectric medium will be described by a Hankel function due to the fact that there is only one finite boundary condition at $r = r_0$, and the fact that Hankel functions have no finite zeros; 4) if $h_1^2 < 0$, the field will be described by a modified Bessel function of the second kind and will represent a trapped surface wave propagating along the interface [5], [8]. This also follows from well known solutions of (1) in a circular cylindrical coordinate system.

If $h_1^2 > 0$, a leaky surface wave exists in the dielectric medium with propagation of the form $\exp[i(\omega t - h_1 r - kz)]$ when the asymptotic representation for Hankel function is used. The phase velocity along the normal to the wave front is easily calculated [2]:

$$v_{ph} = \frac{\omega}{\sqrt{h_1^2 + k^2}} \tag{6}$$

at an angle to the z-axis

$$\theta = \cos^{-1}\left(\frac{k}{\sqrt{h_1^2 + k^2}}\right). \tag{7}$$

The substitution of (3) into (7) yields

$$\cos\theta = \frac{k}{k_1} = \frac{1}{\left(\frac{\omega}{k}\right)\frac{\sqrt{\mu_r\epsilon_r}}{c}} = \frac{1}{\beta\sqrt{\mu_r\epsilon_r}} \tag{8}$$

where $c\beta = (\omega/k) = v_z$. But (8) exactly represents the Cerenkov condition [6], requiring the wave front velocity along the interface (in the z direction) to be equal to the velocity of the particle moving along the interface; and requiring that this velocity, v_z, be greater than the phase velocity of the medium. Comparison of (6) and (8) shows this to be the case when $h_1^2 > 0$.

The limiting case occurs when $h_1^2 \to 0$, where transition from leaky to trapped surface waves takes place [5, 8]. This condition corresponds to incidence at the critical angle ϕ_c [7] at the interface in the limit as $\beta \to 1$. Since the angle of incidence ϕ is related to the Cerenkov angle by $\frac{1}{2}\pi$, *i.e.*, $\theta = \frac{1}{2}\pi - \phi$, then $\cos\theta = \sin\phi$, and in the limit as $\beta \to 1$, $\sin\phi \to \sin\phi_c = (\mu_r\epsilon_r)^{-1/2}$.

Thus, making use of the formalism usually invoked in solution of problems of propagation of electromagnetic surface waves, the results obtained corroborate the viewpoint presented by Brumbaugh [1]; namely, that Cerenkov radiation can be viewed as a result of the continuous reflection of incident waves from a vacuum-dielectric boundary, with the necessary excitation fields supplied by the motion of charged particles on the vacuum side of the boundary.

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Noise Figure for Negative Source Resistance*

When parametric amplifiers and Esaki diode amplifiers are used as the first stage in a cascade of twoport amplifiers, it sometimes happens that the stage following such an amplifier "looks" into a source impedance (the output impedance of the preceding stage) with a negative real part. For brevity, we shall call a twoport that exhibits an output impedance with a negative real part "negative resistance twoport."

The measurement or computation of the over-all noise figure of the cascade does not encounter conceptual difficulties because the source (usually antenna) impedance connected to the input of a receiving system has a positive real part. However, often one is interested in determining experimentally the noise figure of the first stage from measurements on the cascade, because all pertinent noise parameters of the first stage may be determined from the value of noise figure for various source impedances.[1] In the evaluation of such measurements one cannot have recourse to the cascading formula using the noise figure definition of the IRE. Indeed, the source impedance connected to the second stage has a negative real part and the IRE definition of the noise figure does not apply to this case. Yet it is convenient to use a cascading formula because it leads to the determination of the noise figure of the first stage in a systematic way. This is possible if the noise figure definition based on the exchangeable power concept is used that is consistent with the cascading formula.[2] One has for the over-all noise figure of two stages in cascade

$$F_e = F_{e1} + \frac{F_{e2} - 1}{G_{e1}}. \tag{1}$$

Here the F_e's are the noise figures of the first and second stages defined on the basis of exchangeable power,[2] and G_{e1} is the exchangeable gain of the first stage. In order to find F_{e1} from a measurement of F_e one must know G_{e1} and $(F_{e2} - 1)$ for the value of source impedance that is driving the second stage in the cascade. If this impedance has a negative real part, one must overcome the practical problem of determining $(F_{e2} - 1)$ experimentally for a negative source resistance, a rather difficult and inconvenient measurement to perform directly.

The purpose of this communication is to show how one may determine the second term in (1) from signal measurements on the first stage, and from noise measurements on the second stage by using only source admittances with a positive real part. For this purpose, we rewrite the second term in a form containing terms that are more conveniently measured.

$$\frac{F_{e2} - 1}{G_{e1}} = \frac{(F_{e2} - 1)G_{1\,\mathrm{out}}}{G_{e1}G_{1\,\mathrm{out}}}. \tag{2}$$

$G_{1\,\mathrm{out}}$ is the output conductance of the first stage which acts as the source conductance of the second stage. It is well-known[1] that the conventional IRE definition of the noise figure of a twoport can be expressed as a function of the source admittance $Y_s = G_s + jB_s$ for $G_s > 0$ as follows:

$$F = F_0 + \frac{R_n}{G_s}[(G_s - G_0)^2 + (B_s - B_0)^2]. \tag{3}$$

The quantities F_0, R_n, G_0 and B_0 are the four noise parameters that describe the noise of the twoport (now the second stage). The extended definition[2] of the noise figure F_e preserves the analytic form of F in terms of G_s

* Received March 29, 1962; revised manuscript received, April 9, 1962. This work was supported in part by the U. S. Army Signal Corps, the Air Force Office of Scientific Research, and the Office of Naval Research.

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and B_s for negative values of G_s. If one plots

$$H = (F_s - 1)G_s = (F_0 - 1)G_s + R_n[(G_s - G_0)^2 + (B_s - B_0)^2] \quad (4)$$

as a function of G_s, one finds it to be a parabola with a minimum

$$H_{\min} = G_0(F_0 - 1) - \frac{(F_0 - 1)^2}{4R_n} + R_n(B_s - B_0)^2 \quad (5)$$

at

$$G_s - G_0 = -\frac{F_0 - 1}{2R_n}. \quad (6)$$

It can be shown that $H_{\min} > 0$ in all cases. A typical plot of H vs G_s has the appearance of Fig. 1. (It should be noted, however, that $H_{\min}$ may occur for negative values of G_s.) A measurement of H for positive G_s may be extrapolated to negative values of G_s. One simple way of doing this is to plot

$$(H - H_{\min}) \text{ vs } x = \left(G_s - G_0 + \frac{F_0 - 1}{R_n}\right)^2. \quad (7)$$

This plot is a straight line, a fact that can serve for a check of the experimental accuracy of the measurements with positive values of G_s, as well as a simple means of extrapolating H to negative values of G_s. The measurement of $G_{1\,\text{out}}$ is straightforward. The experimentally obtained value of $G_{1\,\text{out}}$ is used as the source conductance on the previously obtained plot to read off the value of $H=(F_s-1)G_s$ for the second stage.

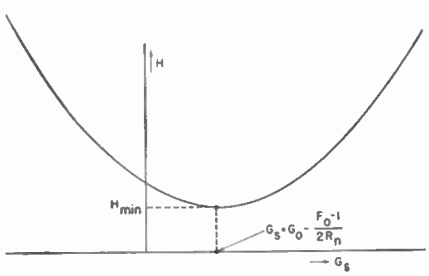


Fig. 1—Plot of H vs G_s.

The value of $G_{e1}G_{1\,\text{out}}$ may be obtained from a transducer gain G_T measurement on the first stage. If we use Norton's equivalent signal current source i_s and admittance Y_s of the source, and the corresponding signal quantities i_0, $Y_{1\,\text{out}}$ at the output port of the first stage connected to the source, we have for the transducer gain of the first stage

$$G_{T1} = \frac{\overline{i_0^2}}{i_s^2} \frac{4G_L G_s}{|Y_{1\,\text{out}} + Y_L|^2} \quad (8)$$

with $G_L = \text{Re}\,(Y_L)$, where Y_L is the admittance of the load connected to the output of the first stage. The exchangeable gain in turn is

$$G_{e1} = \frac{\overline{i_0^2}}{i_s^2} \frac{G_s}{G_{1\,\text{out}}}. \quad (9)$$

Therefore

$$G_{e1}G_{1\,\text{out}} = G_{T1} \frac{|Y_{1\,\text{out}} + Y_L|^2}{4G_L}. \quad (10)$$

Thus one may obtain $G_{e1}G_{1\,\text{out}}$ from measurements of the transducer gain of the first stage, of the load admittance Y_L and of the output admittance of the first stage. With these measurements one may determine $(F_{e2}-1)/G_{e1}$ as it appears in the cascading formula. $(F_{e1}-1)$ may thus be evaluated.

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Optimization of Pulse Transmission*

Gerst and Diamond[1] have considered the elimination of intersymbol interference in time-invariant linear systems by properly specifying the input waveform. In the transmission of digital data it is also of interest to design the signaling waveform to insure maximum energy at the output per unit of input energy. This communication is concerned with the design of signaling waveforms for linear systems which will not only insure a time-limited output signal but will also insure maximum energy transfer. If the system transfer function in Fig. 1 is

$$F(s) = \frac{\sum_{i=0}^{m} k_i s^i}{\sum_{i=0}^{n} h_i s^i}, \text{ with } m \le n \text{ and } h_n = 1 \quad (1)$$

then we let

$$e_0(t) = k_m \frac{d^m}{dt^m} f(t) + k_{m-1} \frac{d^{m-1}}{dt^{m-1}} f(t) + \cdots + k_0 f(t) \quad (2)$$

and

$$e_{in}(t) = \frac{d^n}{dt^n} f(t) + h_{n-1} \frac{d^{n-1}}{dt^{n-1}} f(t) + \cdots + h_0 f(t). \quad (3)$$



Fig. 1.

Without being too restrictive we consider the class of $n+1$ times differentiable output signal shapes to which the calculus of variations is easily applied. We then minimize

$$\int_0^T [e_0^2(t) - e_{in}^2(t)]dt = \int_0^T g(t)dt \quad (4)$$

subject to the constraint

$$\int_0^T e_{in}^2(t)dt = E_{in} \quad (5)$$

and require $f(0)=f(T)=0$. The solution may be obtained by solving the Euler-Lagrange equation.

For the simple case of a low-pass RC network the above steps are easily applied. Here

$$e_0(t) = \frac{1}{RC} f(t) \quad (6)$$

$$e_{in}(t) = \frac{d}{dt} f(t) + \frac{1}{RC} f(t) \quad (7)$$

and the Euler-Lagrange equation becomes

$$\frac{\partial g}{\partial f} - \frac{d}{dt}\left(\frac{\partial g}{\partial f'}\right) = 0 \quad (8)$$

from which we find

$$f(t) = \sqrt{\frac{2E/T}{1 + \frac{\pi^2 R^2 C^2}{T^2}}} \sin \frac{\pi t}{T}. \quad (9)$$

Hence,

$$e_{in}(t) = \frac{1}{RC}\sqrt{\frac{2E}{T}} \sin\left(\frac{\pi}{T}t + \tan^{-1}\frac{1}{RC}\frac{T}{\pi}\right) \quad 0 \le t \le T \quad (10)$$

and

$$e_0(t) = \frac{1}{RC}\sqrt{\frac{2E/T}{1 + \frac{\pi^2 R^2 C^2}{T^2}}} \sin \frac{\pi t}{T} \quad 0 \le t \le T. \quad (11)$$

Note that the signaling waveform is a gated sinusoid whose phase depends upon the baud length and the RC time constant.

It is interesting to examine the cost of eliminating intersymbol interference. Illustrated in Fig. 2 is the energy transfer for a rectangular signaling pulse of duration T, and the pulse given by (10). For example, when $T=RC$, the additional cost in energy is over 6 db; however, the time required to receive 90 per cent of the rectangular transmitted pulse is 3 times the duration of the optimum received pulse given by (11).

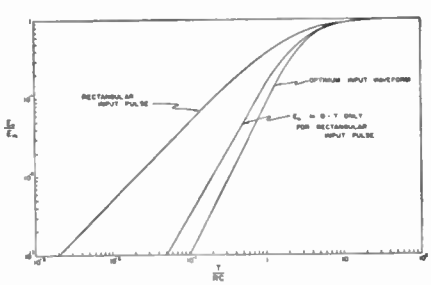


Fig. 2—Energy transfer ratio for the RC filter.

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* Received March 26, 1962.
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Frequency Folding with Up-Converters*

The purpose of this communication is to describe the results of an experiment in frequency folding in a noninverting parametric up-converter.

Various papers[1,2] have described the sensitivity of nonlinear resistance mixers driven by distributed local oscillators where the RF bandwidth is considerably greater than the IF bandwidth. Such devices are effective in folding a wide RF band into a narrow IF band; however, the conversion efficiency of a mixer with distributed local oscillators decreases approximately in the same ratio as the number of local oscillators added and the sensitivity of such systems is very poor. The degradation in conversion efficiency can be attributed to a decrease in the Fourier series coefficients of the dynamic conductance of the mixer crystal,[1] and to the generation of parasitic frequencies which cannot be reactively terminated and thus absorb power from the signal.[2]

Detrimental effects of this nature using nonlinear resistive mixers lead to speculation as to whether a distributively pumped nonlinear capacitance (*i.e.*, varactor) would yield similar results. Analysis of a double-pumped varactor[3] has shown that the Fourier series coefficients of the dynamic capacitance may be actually increased through double pumping. When a double-pumped varactor is used in a noninverting up-converter configuration where the input bandwidth is greater than the output bandwidth, parasitic frequencies are also generated; however, in this type of device the parasitics which cannot be reactively terminated absorb power from the pumps rather than the signal.

The small signal analysis of a double-pumped noninverting up-converter having frequency relations as shown in Fig. 1 has been performed[3] resulting in the Manley-Rowe energy relation as given in (1).

$$P_{f_5} = -\left(P_{f_1}\frac{f_5}{f_1} + P_{f_2}\frac{f_5}{f_2}\right). \qquad (1)$$

Eq. (1) shows that the double-pumped up-converter is essentially a device which converts energy from a bandwidth 2β into a bandwidth β with a conversion gain f_5/f_1 for signals at f_1 and f_5/f_2 for signals at f_2. The same generalization would hold true if the device were operated as a double-pumped down-converter except that conversion loss would result rather than conversion gain.

In order to verify the conclusions of the small signal analysis, an S- to X-band up-converter was implemented using a Microwave Associates MA 4298 varactor. The up-converter frequency relations selected are shown in Fig. 2. The up-converter had a calculated gain of 2.8 db based on an effective $f_{co}=70$ kMc and a filling ratio of 1/4. The measured gain was 2.5 db. The frequency relations are slightly different from the case analyzed and shown in Fig. 1 in that two outputs are obtained (f_5 and f_5') rather than folding the two input signals (f_1 and f_2) into one single output (f_5). The output spacing was deliberate such that the single- and double-pumped outputs might readily be displayed and compared on a spectrum analyzer.

Fig. 3 illustrates the results obtained in the single- and double-pumped cases. In Fig. 3(a), the device was operated as a standard single-pumped up-converter with the input $f_2=3060$ Mc and the output $f_5'=10{,}200$ Mc. In Fig. 3(b), the device was again single-pumped with the input $f_1=2840$ Mc and the output $f_5=10{,}195$ Mc. Fig. 3(c) shows the results of applying both signals and both pumps simultaneously and some slight degradation in output at $f_5=10{,}195$ Mc can be observed. The slight decrease was found to be due to improper termination of the parasitic frequency at $f_1+f_3=9980$ Mc where an output existed which was approximately 13 db below f_5.

Another parasitic output frequency was found to exist at $f_2+f_4=10{,}415$ Mc which was about 18 db below f_5' and apparently not large enough to produce any visible degradation in that output.

The results of the experiment verify the conclusions of the small signal analysis and (1). It would thus appear that distributive pumping of noninverting up-converters can, with proper design, provide a means of folding a wide signal bandwidth into a narrow output bandwidth without suffering any degradation in conversion efficiency.

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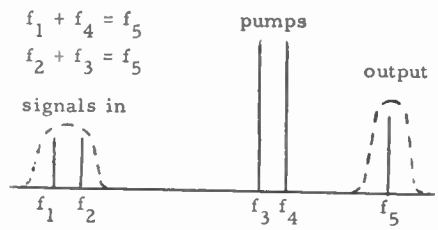


Fig. 1—Noninverting up-converter frequency relations.

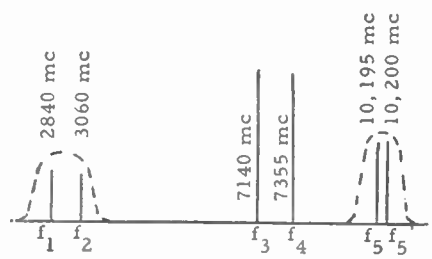


Fig. 2—Experimental up-converter frequency relations.

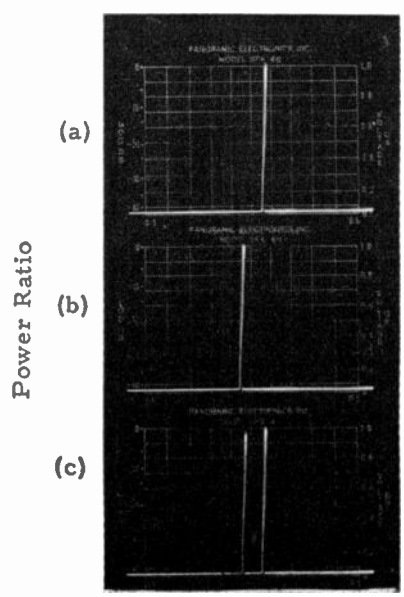


Fig. 3—Up-converter output. (a) and (b) Single-pumped. (c) Double-pumped.

* Received March 21, 1962; revised manuscript received, March 23, 1962.

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Antenna Gain-Loss Estimates*

The "gain-loss" of a large antenna is the quotient of two ratios: the ratio of the power received by a large aperture antenna to that received by a reference antenna of small aperture on the same transhorizon path, divided by the ratio of the corresponding powers in free space, with a single plane wave incident normally on the apertures.

Yeh recently published a collection of experimental measurements of gain-loss [1], for guidance of tropospheric system designers. Unfortunately, his curve does not include two important experimental data, and two important data seem to be plotted incorrectly. When the latter are corrected and additional available data at 9000 Mc at 46 miles and 400 Mc at 600 miles are included, no curve such as Yeh's can claim to approximate the experimental facts.

The incorrectly plotted points are those taken from Crawford [2] (4110 Mc, 171 miles) and from Geiger [3] (9640 Mc, 173 miles). Crawford [4] reports the ratio of the yearly median levels of received power to be 5.7 db on the 60-foot and 8-foot antennas. The reported ratio of plane-wave gains for these two antennas is 16.5 db, so the apparent gain-loss for one 60-foot antenna, compared to an 8-foot antenna, becomes $16.5-5.7=10.8$ db, and not 5.7 db as plotted by Yeh [1]. Geiger's measurements [3] show the gain-loss for *one* 8-foot antenna, compared to a horn, to be 4.4 db rather than 4.5 db for *two* antennas as reported by Yeh. Geiger [6] estimates the total gain-loss to be 8 db when two 8-foot paraboloids replace horns.

Since the gain-loss is measured with respect to a reference antenna with varying

* Received March 26, 1962.

TABLE I
EXPERIMENTAL ANTENNA GAIN-LOSS DATA

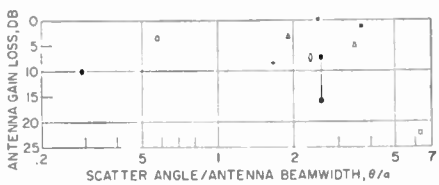
Reference Author	Frequency (Mc)	Distance (Miles)	Antenna Diameter (Feet)	θ/α	Gain-Loss	
					One Antenna	Two Antennas
Trolese [7]	9375	46.3	4	0.288	5±0.5	10±1
M.I.T. Lincoln Lab.	412.8	618	60	2.5	0	0
M.I.T. Lincoln Lab.	399	639	120	3.76	0.5	1.0
Bullington [9]	4090	173	28	2.4	3.5	7.0
Chisholm *et al.* [10]	3670	188	28	2.7	3.5	7.0
Chisholm, *et al.*	3670	188	28	2.7	7.9	15.8

directivity from one experiment to another, measured losses should be referred to comparable reference antennas before being compared. In plotting the measured data of Chisholm [8], 5-foot dishes were assumed herein for reference antennas because their directivity is roughly comparable to that of the other data plotted.

For present purposes, where gain-loss measurements for large aperture antennas at both ends of the path are not available, the loss measured for one narrow-beam antenna will be doubled to obtain the loss for two such antennas. This is by no means always true of course, but it at least approximates the truth in Trolese's experiments.

Additional available data on gain-loss is listed in Table I [7]–[10]. All data are plotted in Fig. 1 as a function of θ/α (scatter angle/antenna beamwidth). It is evident that a well-fitting empirical curve cannot be drawn through these data. The lack of fit is even more serious for the many diverse theoretical curves, especially when *all* of the data are presented. The disagreement with Trolese's 9375-Mc, 46-mile data was noted earlier both by Waterman [11] and by Armand and Vvedenskii [12]. The lack of significant gain-loss on the newer Lincoln Lab., 400-Mc, 600-mile observations serves to emphasize this disagreement. These considerations lead one to suspect that the observed antenna gain-loss on long tropospheric paths beyond the horizon is not a simple function of θ/α.

Armand and Vvedenskii [12] have recently suggested that gain-loss may be proportional to the ratio of the angular spread θ_0 occupied by the ensemble of received waves caused by the mechanism. This essentially revives the suggestion originally offered in 1951 by Schott [13] in the first paper on this phenomenon. Armand and Vvedenskii further hypothesize that this angular spread θ_0 may be inversely proportional to the path length d, and thus that the gain-loss measurements may be expected to be a function of fD/d since the numerator is proportional to the beam angle (frequency f in Gc, antenna diameter D in feet, and distance d in hundreds of miles). Fig. 2 shows the measurements plotted against fD/d. A single line seems to go well through the points in contrast to Fig. 1. We do not extend the line to the lowest point on the right because this yearly median value certainly includes losses due to local refraction variations greater than the 0.3° beamwidth used. These variations are not properly classified as gain-loss. Also for this point, the rough assumption of doubling the measured db loss for one antenna is especially an exaggeration of what would be measured with large antennas at both ends of the path. The lower of the two points at abscissa 54 is partly a sea reflection effect at a low-sited terminal. Despite the well-known time variability of the gain-loss effect, the measured median losses appear to be a monotonic function of fD/d.



Experimental Points
△ Chisholm, *et al.*, 2290 Mc.
● Chisholm, *et al.*, 3670 Mc.
★ M.I.T., Lincoln Lab., 400 Mc.
□ Crawford, *et al.*, 4110 Mc.
◇ Bullington, 4090 Mc.
■ Trolese, 9375 Mc.
○ Angell, *et al.*, 3480 Mc.
+ Geiger, *et al.*, 9640 Mc.

Fig. 1—Antenna gain-loss for two antennas as a function of θ/α.

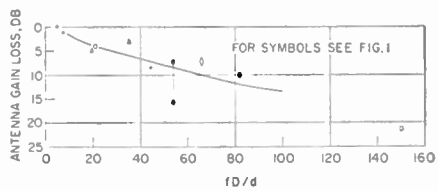


Fig. 2—Empirical curve for estimating antenna gain-loss for two antennas.

One implication of this plot is particularly interesting. It suggests that large aperture antennas may more nearly realize their plane-wave gains over long distances, thus making the use of SHF more feasible for long distance tropospheric circuits than heretofore assumed. There is some limited experimental justification for this optimism: Trolese [7] at 9375 Mc, over 46 miles, measured a gain degradation of 10±1 db for two 4-foot dishes, while Geiger [3], at 9640 Mc, over 173 miles, measured 4.4 db for one 8-foot antenna, or 8.8 db for two 8-foot antennas, showing a definite decrease in gain-loss with distance. Also, Potter [14] has reported some NEL measurements at 9375 Mc, which in two cases show a decrease of gain-loss with distance.

Until more is known, both experimentally and theoretically, about the phenomenon, Fig. 2 seems to offer a safer empirical estimate for system designers than the usual curves plotted as a function of θ/α.

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[13] F. W. Schott, "On the response of a directive antenna to incoherent radiation," PROC. IRE, vol. 39, pp. 677–680; June, 1951.
[14] C. A. Potter, "Tropospheric scattering of microwaves," *Proc. Decennial Symp. of ONR*, pp. 469–479; March, 1957.

The Assistor, a Component with Bipolar Negative Resistance*

I had lately resumed—after 36 years—a study of solid-state negative resistors. I now have evolved a new and promising basic component. In its simplest form it consists of a few grains of powdered semiconductor pressed between metal electrodes.

The voltage vs current characteristic of the new device exhibits a smoothly curved region of negative resistance for either polarity, and for rising and falling test current. Fig. 1 shows a typical plot as traced with the conventional oscilloscope test circuit of Fig. 2. For very low currents the resistance is linear and large, of the order of +50 kohm; the curve turns over, *e.g.*, at ±3 v and 0.1 ma, to drop with a slope of, *e.g.*, initially −5 kohm and more to level out at, *e.g.*, ±2 volt. The positive and negative branches are, as a rule, strictly symmetrical and identical for forward and return trace. The lowest leveling voltage observed is ±2 v with peaks then at ±3 v; with higher final levels the peaks are usually much more than proportionately higher. The characteristic does not vary from dc to—in some cases—the megacycles. One can plot the familiar

* Received March 19, 1962.

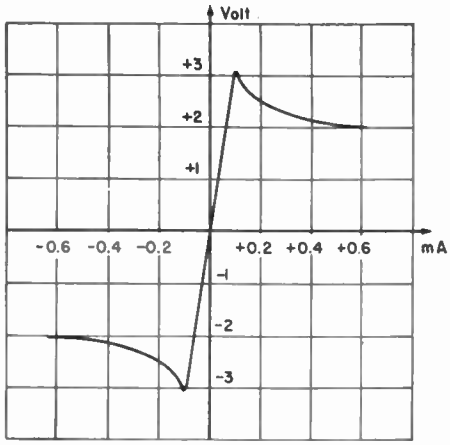


Fig. 1—Typical voltage vs current characteristic.

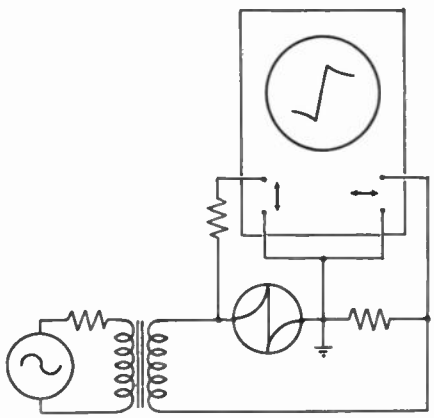


Fig. 2—Circuit for curve tracing with oscilloscope.

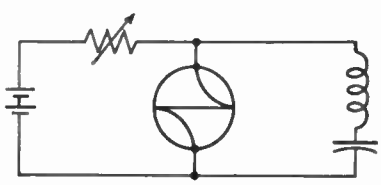


Fig. 3—Oscillator circuit.

square loop by varying a dc source voltage applied via a fixed series resistor.

Given a chance—by shunting with a series-tuned circuit—the device will readily oscillate as in the circuit of Fig. 3 from, *e.g.*, 65 cps to over a megacycle. In some (dubious) conditions the device contributes apparent inductance so that the frequency of oscillation is lower than that expected from the tuned circuit. The device differs from any now known. It has a voltage rather than the current peak of the tunnel diode and it is bipolar, but unlike the carbon arc has negative resistance for decreasing as well as for rising current.

The metal electrodes may be slightly domed flat, or simply two round wires crossing at a right angle, such as steel wires of 0.25 mm diameter, with a few grains of powder sprinkled between them; the required pressure is then of the order of a few grams which in view of the small area amounts to at least a thousand pounds per square inch. As semiconductors I have tested a few metal oxides, none failed, powdered to about 0.1 mm grain size. I use—mainly from habit—zinc oxide, ZnO. Even table salt, NaCl, finely powdered, will work at least in the kilocycle region. In its present shape the device is easily disturbed; and it will seldom dissipate more than about 2–5 mw without early harm.

I feel that, rather than the double negation "negative resistor," the device merits a new name, such as "assistor" and a new circuit symbol, such as that used in Figs. 2 and 3.

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Microwave Phonon Instabilities Associated with Ferromagnetic Resonance*

Parametric coupling between the uniform precession and the low-frequency resonant vibrational modes of a ferrimagnetic insulator has been reported by Spencer and LeCraw[1] and more recently by Comstock, Auld, and Wade.[2] In this note conditions are given under which microwave phonons, propagating in the cubic lattice of a ferrimagnetic insulator, can be parametrically excited by the uniform precession undergoing ferromagnetic resonance at frequency ω_0.

Let $\varrho = i\xi + j\eta + k\zeta$ be the displacement vector in a deformable cubic crystal where the (100) directions correspond to $x_i(x, y, z$ for $i = 1, 2, 3)$. The strain tensor components are $\epsilon_{ij} = \partial\rho_i/\partial x_j$ and the elastic energy U (including magnetoelastic energy) is of the form

$$\begin{aligned} U = {} & \tfrac{1}{2}\gamma(\epsilon_{11}^2 + \epsilon_{22}^2 + \epsilon_{33}^2) \\ & + \lambda(\epsilon_{11}\epsilon_{22} + \epsilon_{11}\epsilon_{33} + \epsilon_{22}\epsilon_{33}) \\ & + \tfrac{1}{2}\mu[(\epsilon_{12} + \epsilon_{21})^2 + (\epsilon_{13} + \epsilon_{31})^2 \\ & + (\epsilon_{23} + \epsilon_{32})^2] + a(M_1^2\epsilon_{11} + M_2^2\epsilon_{22} \\ & + M_3^2\epsilon_{33}) + b[M_1M_2(\epsilon_{12} + \epsilon_{21}) \\ & + M_1M_3(\epsilon_{13} + \epsilon_{31}) + M_2M_3(\epsilon_{23} + \epsilon_{32})] \\ & + \tfrac{1}{2}c(M_1^2\epsilon_{11}^2 + M_2^2\epsilon_{22}^2 + M_3^2\epsilon_{33}^2) \\ & + d(M_1^2\epsilon_{22}\epsilon_{33} + M_2^2\epsilon_{11}\epsilon_{33} + M_3^2\epsilon_{11}\epsilon_{22}) \\ & + \tfrac{1}{2}e[M_1^2(\epsilon_{23} + \epsilon_{32})^2 + M_2^2(\epsilon_{13} + \epsilon_{31})^2 \\ & + M_3^2(\epsilon_{12} + \epsilon_{21})^2] + \cdots . \end{aligned}$$

The differential equations of motion governing ϱ are given by

$$D\ddot{\rho}_i + \alpha\dot{\rho}_i - \frac{\partial}{\partial x_j}\left(\frac{\partial U}{\partial \epsilon_{ij}}\right) = 0 \quad (i = 1, 2, 3) \quad (1)$$

where a repeated index (in this case j) indicates summation over that index. D is the mass density of the crystal and α is a damping parameter for the elastic waves.

Assume that the crystal is magnetized along some z' axis and is undergoing ferromagnetic resonance. Then

$$\boldsymbol{M} = M\sin\theta(i'\cos\omega_0 t + j'\sin\omega_0 t) + k'M\cos\theta$$

where i, j and k are related to i', j' and k' by a pure rotation transformation so that $M_i = l_{ij}M_j'$.

Longitudinal Waves

Consider a longitudinal wave propagating in the x direction. In this case $\eta = \zeta = 0$ and $\xi = \xi_0(t)e^{-ikx}$. Eq. (1) becomes

$$D\ddot{\xi}_0 + \alpha\dot{\xi}_0 + k^2(\gamma + cM_1^2)\xi_0 = 0. \quad (2)$$

If $l_{ij} = \delta_{ij}$ (the Kronecker delta),

$$M_1^2 = M^2\sin^2\theta\tfrac{1}{2}(1 + \cos 2\omega_0 t).$$

Since (2) is a Mathieu equation there will be a second-order instability, provided that

$$k \simeq \frac{\omega_0}{\sqrt{\gamma/D}} \quad \text{and} \quad \sin\theta \geq 2\sqrt{\frac{(\alpha/D)\gamma}{\omega_0 cM^2}}.$$

The threshold is naturally the same for longitudinal waves propagating along y; there is no coupling to longitudinal waves propagating along z.

If

$$\begin{aligned} & l_{ij} \neq \delta_{ij}, \\ & M_1^2 \simeq l_{13}^2M^2 + 2l_{13}\sqrt{1 - l_{13}^2}\,M^2 \\ & \cdot\sin\theta\cos(\omega_0 t + \phi_0) \end{aligned}$$

(neglecting terms in $\sin^2\theta$ etc.). A first-order instability can occur if

$$k = \frac{\omega_0}{2}\Big/\sqrt{\frac{\gamma + l_{13}^2cM^2}{D}}$$

and

$$\sin\theta \geq \frac{2(\alpha/D)}{\omega_0}\,\frac{\gamma + l_{13}^2cM^2}{cM^2l_{13}\sqrt{1 - l_{13}^2}}.$$

If $cM^2 \ll \gamma$, θ will be minimized when $l_{13} = 1/\sqrt{2}$.

Transverse Waves

Consider a transverse wave linearly polarized along x and propagating in the z direction. In this case $\eta = \zeta = 0$ and

$$\xi = \xi_0(t)e^{-ikz}.$$

Eq. (1) becomes

$$D\ddot{\xi}_0 + \alpha\dot{\xi}_0 + k^2(\mu + eM_2^2)\xi_0 = 0. \quad (3)$$

If $l_{ij} = \delta_{ij}$, the instability occurs when

$$k \simeq \frac{\omega_0}{\sqrt{\mu/D}} \quad \text{and} \quad \sin\theta \geq 2\sqrt{\frac{(\alpha/D)\mu}{\omega_0 eM^2}}.$$

The threshold is the same for transverse waves polarized along y propagating in the z direction; however, there is a 90° phase shift between the $\xi(z)$ and $\eta(z)$ components (circularly polarized if amplitudes are equal). Transverse waves $\xi(y)$ or $\eta(x)$ are not coupled to the uniform precession.

* Received March 27, 1962; revised manuscript received April 10, 1962. This research was sponsored by the Ford Foundation.

[1] E. G. Spencer and R. C. LeCraw, "Magnetoacoustic resonance in yttrium iron garnet," *J. Appl. Phys.*, vol. 30, pp. 149S–150S; April, 1959.

[2] R. L. Comstock, B. A. Auld, and G. Wade, "Parametric effects in magnetoacoustic resonance," *J. Appl. Phys.*, vol. 32, pp. 225S–226S; March, 1961.

If $l_{ij} \neq \delta_{ij}$, a first order instability occurs when

$$k = \frac{\omega_0}{2} \Big/ \sqrt{\frac{\mu + l_{23}^2 eM^2}{D}}$$

and

$$\sin\theta \geq \frac{2(\alpha/D)}{\omega_0} \frac{\mu + l_{23}^2 eM^2}{eM^2 l_{23}\sqrt{1 - l_{23}^2}}.$$

If $eM^2 \ll \mu$, θ will be minimized when $l_{23} = 1/\sqrt{2}$.

Conclusion

The instability thresholds given are for phonons propagating along principal axes of the crystal and may be considered representative of thresholds for other propagation directions. The ratio of $\mu c/\gamma e$ together with the magnetizing geometry will determine which type of phonon (longitudinal or transverse) will have the lowest threshold.[3] If $\omega_0/2$ spinwaves do not exist (this can always be arranged by making ω_0 sufficiently large) degenerate spinwave instabilities occur if

$$\theta \geq \sqrt{\frac{2\Delta H_k}{M}}$$ [4]

where $2\Delta H_k$ is the pertinent spinwave linewidth). Assuming that the first-order phonon coupling is dominant one must compare

$$\frac{4(\alpha/D)}{\omega_0}\frac{\mu}{eM^2} \text{ or } \frac{4(\alpha/D)}{\omega_0}\frac{\gamma}{cM^2} \text{ with } \sqrt{\frac{2\Delta H_k}{M}}$$

where the intrinsic Q of the phonon is

$$\left(\frac{\omega_0}{2}\right) \Big/ \left(\frac{\alpha}{D}\right).$$

It would appear that if α is independent of ω_0 the phonon coupling dominates above some characteristic frequency.

For yttrium iron garnet (YIG):

$$\sqrt{\frac{2\Delta H_k}{M}} \simeq 1/40.$$

This means that for a supposed Q of 10^5 one would expect the phonon coupling to dominate provided that

$$\frac{eM^2}{\mu} \text{ or } \frac{cM^2}{\gamma} \gtrsim 10^{-3}.$$

Even if this condition is not satisfied (or the Q's are lower) it might be possible to increase ΔH_k through rare earth additions to the point where the inequality is satisfied. This might work if α were relatively insensitive to the rare earth impurity.

If the phonon coupling is desired and YIG turns out to be unsuitable, one might profitably investigate one of the high anisotropy materials such as barium ferrite which undoubtedly has greater "eM^2/μ" ratios. A study of phonon Q's and ΔH_k's in these materials is essential. Of course, the results derived here are for cubic symmetry and would have to be modified accordingly.

The author[5] has suggested that the thresholds of spinwaves degenerate with the uniform precession can be raised by modulating the spinwave frequencies without at the same time modulating the uniform precession. Such a technique will *not* raise the phonon thresholds and so apart from other limitations of the method there is a new one. If large-amplitude resonance is the desired goal, it appears that one wants to magnetize along a (100) direction of the crystal so as to forbid first-order phonon coupling and, if possible, to use a material with low eM^2/μ, cM^2/γ ratios. Another suggestion made by the author[6] concerns the possibility of dropping the uniform precession *below* the spinwave manifold so as to weaken spinwave coupling and allow large-amplitude resonance. In this regard it should be noted that one *always* has phonons which are degenerate with the uniform precession and they may well be the dominant limitation. This point was not considered in the previous reference.[6]

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[3] It is possible that hybrid modes may have comparable or even lower thresholds.

[4] H. Suhl, "The nonlinear behavior of ferrites at high microwave signal levels," Proc. IRE, vol. 45, pp. 1270–1284; October, 1956. See especially p. 1271.

[5] F. R. Morgenthaler, G. E. Bennett, and F. A. Olson, "Suppression of spinwave instabilities associated with ferromagnetic resonance," *Proc. 1961 Internat'l. Conf. on Magnetism and Crystallography*, Kyoto, Japan, September 25–30, to be published, *J. Phys. Soc. Japan*, vol. 17, Suppl., 1962.

[6] F. R. Morgenthaler, "On the possibility of obtaining large amplitude resonance in very thin ferrimagnetic disks," *J. Appl. Phys.*, vol. 33, Suppl., pp. 1297–1299; March, 1962.

Two Inequalities Useful in Transient Analysis*

In a recent communication[1] the following inequality, which is useful in the transient analysis of physical systems, was announced:

Let $w(t)$ be a real-valued function of the real variable t, and let it be zero for $t<0$. Furthermore let its Laplace transform $W(s)$ be a rational function denoted by

$$W(s) = \frac{a_n s^n + a_{n-1}s^{n-1} + \cdots + a_0}{s^m + b_{m-1}s^{m-1} + \cdots + b_0}, \quad (1)$$

where $m \geq 2n$ and all the roots of the denominator have nonpositive real parts. Then for $t \geq 0$

$$|w(t)| \leq \frac{|a_n| t^{m-n-1}}{(m-n-1)!} + \frac{|a_{n-1}| t^{m-n}}{(m-n)!} + \cdots + \frac{|a_0| t^{m-1}}{(m-1)!}. \quad (2)$$

This result was obtained as a special case of a more general conclusion whose proof consisted of some comparatively involved arguments. The first purpose here is to present a direct and considerably simpler proof of this special case. In the following discussion an upper case letter will denote a rational function of s and the corresponding lower case letter will designate that function of t which is zero for $t<0$ and is the inverse Laplace transform of the function of s.

First let us note that every rational function of the form

$$F(s) = \frac{s}{s^2 + c_1 s + c_0}, \quad (3)$$

where c_1 and c_0 are real and the roots of the denominator have nonpositive real parts, is a positive-real function. This means that the function $f(t)$ will satisfy the following inequality for $t \geq 0$:

$$|f(t)| \leq 1. \quad (4)$$

Similarly for

$$G(s) = \frac{1}{s+d}; \quad (d \geqq 0), \quad (5)$$

the function $g(t)$ satisfies

$$|g(t)| \leq 1. \quad (6)$$

Also consider the expression

$$H(s) = \frac{1}{(s+\gamma)(s+\bar{\gamma})}; \quad (\mathrm{Re}\,\gamma \geqq 0) \quad (7)$$

where $\bar{\gamma}$ denotes the complex conjugate of γ. The corresponding function of t satisfies

$$|h(t)| \leq t. \quad (8)$$

Secondly, consider the function

$$M(s) = \frac{s^l}{s^m + b_{m-1}s^{m-1} + \cdots + b_0} \quad (9)$$

where $q \leq m/2$. It can be factored into a product of functions such that q of them have the form of (3) and if $q < m/2$, the remaining part is the reciprocal of a polynomial. This remaining part can be factored into a product of terms, each of which has the form of either (5) or (7). The function of t, corresponding to the first product of q factors of the form (3), is the convolution of q functions, each of which is zero for $t<0$ and bounded by one for $t \geq 0$. It therefore is bounded in magnitude by

$$\frac{t^{q-1}}{(q-1)!}u(t) \quad (10)$$

where $u(t)$ is the unit step function. By a similar argument we see that when $q < m/2$, the function of t, corresponding to the product of factors such as (5) and (7), is bounded in magnitude by

$$\frac{t^{m-2q-1}}{(m-2q-1)!}u(t). \quad (11)$$

Again, convolving (10) and (11), we obtain

$$|m(t)| \leq u(t)\int_0^t \frac{x^{q-1}(t-x)^{m-2q-1}}{(q-1)!(m-2q-1)!}dx = \frac{t^{m-q-1}}{(m-q-1)!}u(t). \quad (12)$$

* Received March 23, 1962.

[1] A. H. Zemanian, "Further properties of certain classes of transfer functions: II," *Quart. Appl. Math.*, vol. 19, pp. 158–159; July 1961.

Since $W(s)$ is a linear combination of terms of the form (9), it follows that (10) and (12) imply (2). Q.E.D.

It can be shown through examples that the conclusion (2) cannot be strengthened if nothing more is assumed known about $W(s)$.

On the other hand if we weaken our assumptions on the denominator of $W(s)$ and the degrees of the numerator and denominator of $W(s)$, a similar inequality can be obtained when we add the assumption that $w(t)$ is non-negative. More specifically, *if $w(t)$ is a real-valued non-negative function of the real variable t and is zero for $t<0$, and if its Laplace transform is given by* (1) *where $m>n$ and the $b_i (i=1, 2, \cdots, m)$ are non-negative, then for $t\geq 0$*

$$w(t) \leq \frac{a_n t^{m-n-1}}{(m-n-1)!} + \frac{a_{n-1} t^{m-n}}{(m-n)!} + \cdots + \frac{a_0 t^{m-1}}{(m-1)!} \tag{13}$$

and thus $W(s)$ has no poles in the half plane Re $s>0$.

The usefulness of this result, whose proof can be found elsewhere,[2] is enhanced by the fact that a number of criteria exist by which one can ascertain whether $w(t)$ is non-negative merely by inspecting the pole and zero locations of $W(s)$.[3,4]

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[2] A. H. Zemanian, "An upper bound on nonnegative transient responses," *Quart. Appl. Math.* (to be published).
[3] A. H. Zemanian, "The properties of pole and zero locations for nondecreasing step responses," *Trans. AIEE (Communications and Electronics)*, no. 50), pp. 421–426; September, 1960.
[4] A. H. Zemanian, "On the pole and zero locations of rational Laplace transformations of nonnegative functions," *Proc. Am. Math. Soc.*, vol. 10, pp. 868–872; December, 1959.

Some Comments on "A New Precision Low-Level Bolometer Bridge"*

The following comments on the above article[1] are unfortunately necessary:

1) Reisener and Birx misrepresent when they claim, "we have been able to achieve an over-all detection limit of 3×10^{-10} w." While an article in the Proceedings is not a patent application, generally authorship implicitly, and here explicitly, testifies to technical, not merely literary, achievement. I had reported[2] the theory of the over-all system of the instrument about a year before Mr. Reisener became familiar with it. I had the burden for all aspects of the bridge development including the construction and testing of both breadboard and prototype models. Mr. Reisener designed many of the circuits and was invaluable in the trouble-shooting, but he quite properly refused to be cited as coinventor for the patent application. Dr. Birx's contribution was very slight. The bridge described is exclusively my invention.

2) What the article refers to as "bolometer instability" is really ambient temperature fluctuation. Ambient temperature fluctuation has greatly limited the sensitivity of all bolometer bridges. The method indicated by Fig. 1 dynamically compensates for temperature fluctuations sufficiently so that in principle substitution bolometric measurements can be made to the limits of sensitivity inherent in the bolometer used. To establish the correspondence of the principle of the bridge to its performance requires the determination of the sensitivity of the bolometer.

3) The statement, "tests showed that the predominant sources of noise were the Johnson noise of the fixed resistors and bolometers and the temperature fluctuation noise of the bolometers," cannot be sustained. To identify the nature of the noise it would be necessary to obtain the noise spectrum over the frequency band 0 cps to about $1/20t$ (t = thermal response time of the bolometer) and to measure the absolute value of the noise. To the best of my knowledge this has not yet been done.

4) There are several errors in the analytical evaluation of the bolometer sensitivity. a) The effective temperature of the bolometer is not 90.8°C as given, but 189°C. The bolometer is a fine platinum wire. Its effective temperature at nominal resistance can be calculated by dividing the difference between the nominal resistance, 100 ohms, and the "cold" resistance, about 60 ohms, by the temperature coefficient of resistance of platinum, 0.00352 according to the International Critical Tables, and the cold resistance. The formula used in the article is not applicable to large temperature changes since the heat loss characteristic β is not a constant but a function of temperature. b) Since both β and T, the temperature, vary along the length of the fine wire constituting the bolometer, and since T occurs as a squared term, the effective temperature cannot be used and the formula for total noise is not valid. c) As Van der Ziel[3] shows, the sensitivity of a bolometer and a bolometer bridge are two different things. No such difference is acknowledged in this evaluation. d) The expression for Johnson noise is too large by a factor of 4. The "maximum available power" is the most that can be drawn from a signal source.

5) If the analysis referred to above were accurate and the tests had been made, the latter would validate the former. While a ratio of the inherent sensitivity of the bolometer to the measured sensitivity of the bridge, as attempted in the article, would be a valid measure of bridge performance, it says nothing as to the correspondence of the principle of the bridge to its performance. For this it would be necessary to calculate the degradation caused by design limitations, *i.e.*, noise figure of the amplifier, band-pass of amplifier, equal arm basic bridge, etc. Time constants greater than the response time of the bolometer are, of course, not relevant to either evaluation.

6) The precise calculation of the bolometer sensitivity is very difficult, but a fair approximation could be made analytically. Furthermore there is sufficient test data to check this approximation with an accuracy comparable to that of the noise figure of the ac amplifier " . . . estimated to be 1.8 db." The law of the instrument allows the calculation of the sensitivity in accordance with the nature and the measured performance of the components so that the correspondence of principle to performance can be estimated also.

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Authors' Comments[4]

The authors are pleased to acknowledge that the conception and initial development of the low-level bridge were the contributions of Mr. Hirsch.

The Franklin Institute has made application for a patent in his name.

To reduce complexity and in the interest of saving space, only a simplified analysis was presented in the paper. Mr. Hirsch is quite correct in pointing out possible errors resulting from the use of mean values for the bolometer temperature rise and the heat loss coefficient. The Johnson noise mentioned is expressed in the paper as an equivalent input power fluctuation.

The performance data presented are reliable and the analysis, while not rigorous, does convey a proper conception of the nearness with which the performance approaches that theoretically possible. It was our purpose, not to present an exhaustive treatise, but to bring to the attention of the engineering community the capabilities of this bridge which we feel contributes significantly to the art of low-level RF measurement.

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[4] Received April 18, 1962.

Causality*

It is most surprising and disappointing to find Deutsch's reflections on Causality[1] in a scientific journal of the year 1962. One hundred years ago, his remarks would have been timely and appropriate. At that time,

* Received March 26, 1962.
[1] W. C. Reisener and D. L. Birx, "A new precision low-level bolometer bridge," Proc. IRE, vol. 50, pp. 39–42; January, 1962.
[2] M. M. Hirsch and D. L. Birx, "Investigation of Low-Level RF Voltage and Power Measurement," Franklin Institute, Philadelphia, Pa., 7th Quart. Engrg. Rept., Signal Corps Contract DA-36-039-SC-72827; March, 1958.
[3] A. Van der Ziel, "Noise," Prentice-Hall, Inc., Englewood Cliffs, N. J., pp. 409–412; 1954.

* Received March 30, 1962.
[1] S. Deutsch, "Causality," Proc. IRE (*Correspondence*), vol. 50, p. 222; February, 1962.

the work of the great mathematicians Laplace and Lagrange had led to the belief that a super-mathematician would be able to predict the entire future of the universe if the present conditions could fully be made known to him. This situation was entirely changed by the advent of quantum mechanics. It is now established that the laws of nature can be formulated only in a statistical manner. The classical physical laws are averages that are accurate with a very high degree of probability, but they are not fundamentally correct. For example, in a radioactive substance it is *in principle* impossible to predict which atom is going to disintegrate next, and no super-computer will be able to provide this knowledge. The laws of quantum mechanics have been verified by such an overwhelming mass of experimental evidence that today there is not the slightest experimental indication that the probabilistic interpretation of these laws is incorrect.

Turning now to the human brain and the human mind, Deutsch correctly states that the nerve cells are complicated electrochemical engines. Their behavior is fairly well understood and, of course, obeys the laws of physics and chemistry. It is further known that the number of molecules involved in a single occurrence at the synapsis is so small that there is, in accordance with Heisenberg's indeterminacy relation, a significant degree of uncertainty as to their location over the period of time of the process. Therefore, our thoughts are *in principle* unpredictable, and certainly a billion years ago neither Golay's paper nor Deutsch's paper (nor any other paper) could have been predicted.

If Deutsch feels that free will depends on the absence of causality, then the existence of free will has abundantly been proved.

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Author's Comment[2]

In 1922, the living cell was seen as a spherical blob having few structural details; in 1962, the electron microscope reveals an unbelievably complex electrochemical engine, with over 1000 different kinds of molecules. A short time ago, a protein molecule was little more than a tiny dot with an impressive atomic weight and some vague chemical bonds; today, the three-dimensional structure is an awe-inspiring miracle.[3] A few years ago, matter was composed of electrons and protons; today there is a bewildering array of "fundamental" particles, with electrons that somehow jump from one quantum orbit to another, with nuclei that remain stable for a million years and then disintegrate for no apparent reason. A multitude of questions remain unanswered, and will never be answered if we accept the naïve viewpoint that present concepts, including quantum theory, give more than a dim outline of reality. Matter may originate in particles that are much smaller than an electron; the units that we regard as fundamental may be the "microorganisms" that have survived in an evolutionary sense. The uncertainty principle, if anything, begs for this interpretation.[4]

Statistical methods are necessary in describing the activities of large populations. The conclusion that individuals do not obey causality, however, is most unjustified. We may not know when a particular atom will disintegrate, but we can be certain that a definite combination of circumstances *causes* this *effect* each time, without exception. A corollary of all this is that the human brain can be duplicated, on a small scale, by means of pattern-recognition units and memory devices.[5]

Although every facet of man's future is written in the present state of the universe, the record is too complicated for us to read. We may yet destroy ourselves because of the hostility that evolution has bestowed upon our genes.

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[4] D. Bohm, "Causality and Chance in Modern Physics," D. Van Nostrand Co., Inc., Princeton, N. J.; 1957.
[5] S. Deutsch, "Causality, consciousness, and creativity," *Cybernetica*, vol. 4, pp. 154–170; 1961.

[2] Received April 19, 1962.
[3] J. C. Kendrew, "The three-dimensional structure of a protein molecule," *Sci. American*, vol. 205, pp. 96–110; December, 1961.

Capacitance Between Thin Film Conductors Deposited on a High-Dielectric-Constant Substrate*

Introduction

In microsystems electronics functional assemblies, it is sometimes desirable to use as a substrate a material of very high dielectric constant in order to achieve capacity between the opposite sides of the substrate. This may be required for lumped capacitors wherein two film conductors are deposited on either side of the substrate, or for distributed parameter *RC* networks, in which case a film resistor replaces one or both of the conductors.

In either case, the presence of the substrate may have deleterious effects on the over-all circuit performance by introducing stray capacitance between various conductors, where no capacitance is acceptable.

It is necessary to have a means for estimating these stray capacitances in order to account for their effects in the completed function morphology.

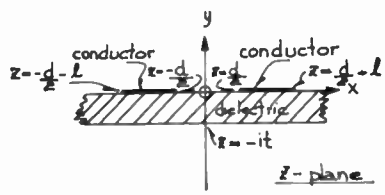
The idealized case of two parallel conductors of equal width on the same side of a plane dielectric of finite thickness and infinite extent may be analyzed analytically. This case is a very close approximation to the practical case of finite substrate dimensions, provided the substrate extends a few thicknesses beyond the conductors.

* Received April 6, 1962.

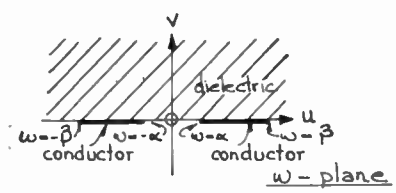
Analysis

The conductor-substrate morphology to be considered is illustrated in Fig. 1(a). The conductors and dielectric are of infinite length (perpendicular to the Z plane) and the capacitance between these conductors will be calculated per unit length. Since the dielectric constant of the substrate is assumed very large compared to that of the surrounding medium, all of the flux is considered confined to the dielectric.

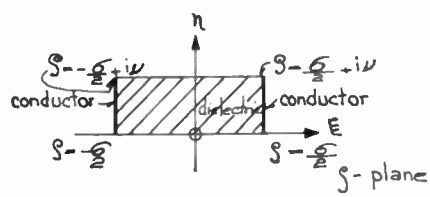
A Schwarz-Christoffel transformation is used to map the Z plane of Fig. 1(a) onto the ω plane of Fig. 1(b). An inverse Schwarz-Christoffel transformation is then used to map the ω plane of Fig. 1(b) onto the ζ plane of Fig. 1(c). The dielectric region shown shaded in the Z plane maps onto the shaded regions in the ω and ζ planes. Thus, the capacitance calculation is reduced to the elementary calculation of the capacitance of parallel plates without fringing fields.



(a)



(b)



(c)

Fig. 1.

The transformation relating the Z and ω planes is readily obtained by the method of Schwarz-Christoffel as

$$Z = \int_0^{\omega} \frac{d\omega}{\omega^2 - \gamma^2} \tag{1}$$

or

$$Z = \frac{1}{2\gamma}\left[\log\frac{\omega - \gamma}{\omega + \gamma} - i\pi\right]. \tag{2}$$

The quantity γ is evaluated as $\pi/2t$ by noting that the point $Z = -it$ goes into the point $\omega = \infty$.

From (2) we obtain

$$\omega = -\frac{\pi}{2t}\tanh\frac{\pi Z}{2t} \tag{3}$$

so that

$$\alpha = \frac{\pi}{2t}\tanh\frac{\pi d}{4t} \tag{4}$$

and

$$\beta = \frac{\pi}{2t}\tanh\left(\frac{\pi l}{2t}+\frac{\pi d}{4t}\right) \tag{5}$$

in Fig. 1(b).

Similarly, the transformation relating the ζ and ω planes is

$$\zeta = \int_0^{\omega} \frac{d\omega}{\sqrt{(\alpha^2-\omega^2)(\beta^2-\omega^2)}} \tag{6}$$

which yields

$$\nu = \int_{\alpha}^{\beta} \frac{d\omega}{\sqrt{(\omega^2-\alpha^2)(\beta^2-\omega^2)}} \tag{7}$$

and

$$\sigma = \int_0^{\alpha} \frac{2d\omega}{\sqrt{(\alpha^2-\omega^2)(\beta^2-\omega^2)}} \tag{8}$$

in Fig. 1(c). The elliptic integral (7) can be put into the Legendre normal integral form of the first kind by means of the transformation

$$\omega = \beta\sqrt{1-x^2\left(1-\frac{\alpha^2}{\beta^2}\right)}$$

to yield

$$\nu = \frac{1}{\beta}F\left(\sqrt{1-\frac{\alpha^2}{\beta^2}},\frac{\pi}{2}\right) \tag{9}$$

in standard elliptic integral notation.

Similarly, the elliptic integral (8) is transformed by putting $\omega=\alpha x$ to yield

$$\sigma = \frac{2}{\beta}F\left(\frac{\alpha}{\beta},\frac{\pi}{2}\right). \tag{10}$$

The capacitance in the ζ plane [Fig. 1(c)] is simply

$$C = \frac{\epsilon_0 K\nu}{\sigma}$$

and from (9) and (10)

$$C = \frac{\epsilon_0 K}{2}\frac{F\left(\sqrt{1-\frac{\alpha^2}{\beta^2}},\frac{\pi}{2}\right)}{F\left(\frac{\alpha}{\beta},\frac{\pi}{2}\right)} \text{ farad/meter.} \tag{11}$$

From (4) and (5)

$$\frac{\alpha}{\beta} = \frac{\tanh\frac{\pi d}{4t}}{\tanh\left(\frac{\pi l}{2t}+\frac{\pi d}{4t}\right)}$$

or

$$\frac{\alpha}{\beta} = \frac{1+\tanh\frac{\pi l}{2t}\tanh\frac{\pi d}{4t}}{1+\tanh\frac{\pi l}{2t}\Big/\tanh\frac{\pi d}{4t}}. \tag{12}$$

To calculate C for given l, d, and t, the quantity α/β is first calculated from (12) and this value used to evaluate (11) by means of tables of elliptic integrals.

Numerical Results

The capacitance in $\mu\mu$f/inch is given by

$$C(\mu\mu f/\text{inch}) = 0.1125K\frac{F\left(\sqrt{1-\frac{\alpha^2}{\beta^2}},\frac{\pi}{2}\right)}{F\left(\frac{\alpha}{\beta},\frac{\pi}{2}\right)}$$

and is plotted in Fig. 2 for unit dielectric constant K vs d/t for various values of l/t.

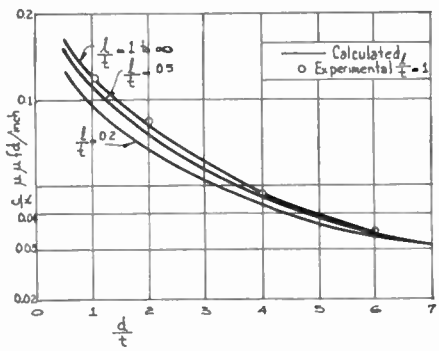


Fig. 2.

Measurements performed on a barium titanate substrate of 0.020 inch thickness and dielectric constant 1773 are also shown in Fig. 2 as circled points.

Acknowledgment

The preparation of substrates and the measurements are due to A. J. Nichols III.

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Hilbert Transforms of Band-Pass Functions*

The Hilbert transform $\hat{x}(t)$ of a (generally complex-valued) function $x(t)$ of a real variable t is defined by

$$\hat{x}(t) = \frac{1}{\pi}\int_{-\infty}^{\infty}\frac{x(v)dv}{t-v} \tag{1}$$

where the integral is a Cauchy principal value. The form of (1) shows that $\hat{x}(t)$ is the convolution of $x(t)$ and $1/(\pi t)$. Therefore, the Fourier transform of $\hat{x}(t)$ is the product of the Fourier transform of $x(t)$ and that of $1/(\pi t)$. That is, if $X(\omega)$, $\hat{X}(\omega)$ are the Fourier transforms of $x(t)$ and $\hat{x}(t)$, respectively, then

$$\begin{aligned}\hat{X}(\omega) &= -jX(\omega), & \omega > 0\\ &= jX(\omega), & \omega < 0.\end{aligned} \tag{2}$$

* Received April 5, 1962.

Now, let

$$x(t) = a(t)\exp(j\omega_c t), \qquad \omega_c > 0$$

where, in general, $a(t)$ may be complex-valued. If $A(\omega)$ is the Fourier transform of $a(t)$, the Fourier transform of $x(t)$ will be

$$X(\omega) = A(\omega - w_c).$$

Then

$$\begin{aligned}\hat{X}(\omega) &= -jA(\omega-\omega_c), & \omega > 0\\ &= jA(\omega-\omega c), & \omega < 0.\end{aligned}$$

If $x(t)$ is a band-pass function, $a(t)$ will be slowly varying compared to ω_c. In particular, let it be required that the total, nonzero extent of $A(\omega)$, including negative ω, be such that $A(\omega-\omega_c)=0$ for $\omega<0$. This will ordinarily require that the "video" bandwidth of $a(t)$ be less than ω_c. With this requirement,

$$\hat{X}(\omega) = -jA(\omega-\omega_c)$$

and

$$\hat{x}(t) = -ja(t)\exp(j\omega_c t).$$

Generally, for any real ω_c,

$$\hat{x}(t) = -j(\operatorname{sgn}\omega_c)a(t)\exp(j\omega_c t) \tag{3}$$

where "sgn" stands for "signum" and is $+1$ if $\omega_c>0$ and -1 if $\omega_c<0$.

Eq. (3) is the basic formula from which the entries in Table I may be derived. In every case, $a(t)$ is a function with transform $A(\omega)$, such that $A(\omega-\omega_c)=0$, for $\omega<0$, $\omega_c>0$ and for $\omega>0$, $\omega_c<0$.

TABLE I

Function, $x(t)$	Hilbert Transform, $\hat{x}(t)$
$a(t)\exp(j\omega_c t+j\theta)$	$-j(\operatorname{sgn}\omega_c)a(t)\exp(j\omega_c t+j\theta)$
$a(t)\cos(\omega_c t+\theta)$	$(\operatorname{sgn}\omega_c)a(t)\sin(\omega_c t+\theta)$
$a(t)\sin(\omega_c t+\theta)$	$-(\operatorname{sgn}\omega_c)a(t)\cos(\omega_c+\theta)$
θ = constant	

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A Large Signal Behavior of a Low-Noise Traveling-Wave Tube and Its Application to Pulse Radar*

A novel pulse radar employing a circulator and low-noise traveling-wave tube as a switching device is suggested [1] and experimental results in X-band 50-kw and S-band 5-Mw radars are reported. In this duplexing system, by taking advantage of limiter action of the traveling-wave tube, the mixer crystals are completely protected from the leaking power of the transmitter. Differing from usual radar, our radars use no TR tubes and naturally there is no spike and therefore no fear about deterioration of the mixer crystals. It is another merit that

* Received April 3, 1962.

some extension of the radar range can be obtained in a low-noise traveling-wave tube. In *S*-band experiments, a ferrite switch is needed in order to protect the traveling-wave tube. High-power behavior of low-noise traveling-wave tubes is shown in Figs. 1 and 2. Gain of the *X*-band tube which was made on a trial basis at our laboratory bench is more than 8 db for the input power less than −20 dbm, for more power than −20 dbm the gain turns to decrease and gets into the attenuation band. In the *S*-band tube

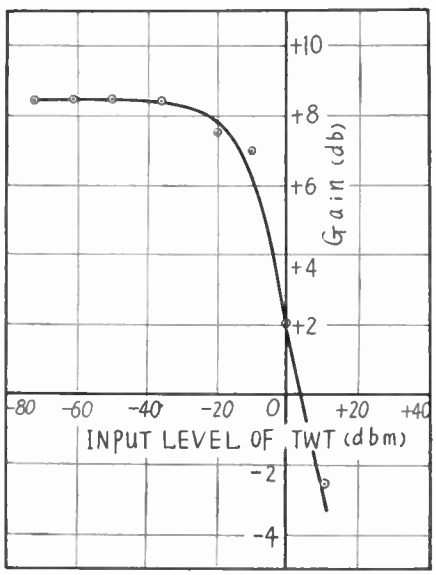


Fig. 1—High-power behavior of the *X*-band traveling-wave tube (f=9375 Mc).

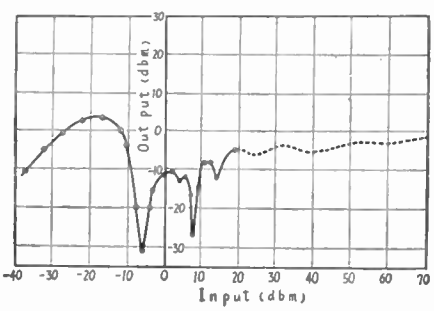


Fig. 2—High-power behavior of the traveling-wave tube 6861.

RCA 6861, however, the attenuation characteristics as a function of input power show a complicated figure. The output power of the tube drops sharply to −30 dbm for the input power of −5 dbm and +8 dbm. This complicated behavior seems to stem from the close coupling between the helix and electron beam.

A schematic diagram of the duplexing device in *S*-band 5-Mw radar is shown in Fig. 3. Though decoupling between ports 1 and 3 of the circulator is fairly large—say 25 db—the reflected power from the antenna, which is about 20 db down from the radiation power, adds to the former. Therefore, some kind of protection to the TWT was needed. For this purpose a Faraday-type ferrite switch was employed where insertion loss was 0.7 db and attenuation 30 db with an exciting coil current of 0.9 a. These were measured at the ambient temperature of 40°C.

Over-all noise figure was measured as 8.7 db in which 6.9 db was due to the TWT only. In an ordinary 5-Mw radar where GE 6621 was employed for the duplexing device, the NF of this system was 9 db in normal operation which sounded almost the same as our radar. However, the striking merit of our radar was its small recovery time. The recovery time of GE 6621 was more than 160 μsec in 5-Mw operation, but in our radar, there was no spike and no recovery time. However, some kind of pulse delay due to magnetizing coil in the ferrite switch had to be considered which measured less than 10 μsec.

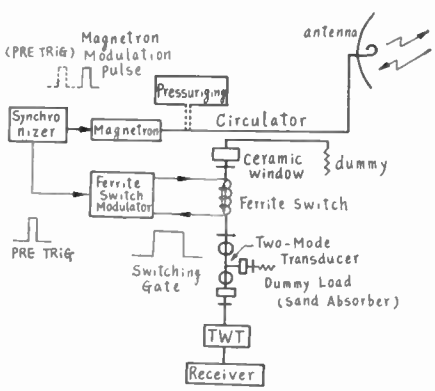


Fig. 3—A schematic diagram of a circulator-TWT-radar in 5-Mw operation.

In the experiment of the *S*-band 500-kw radar, the arrangement was almost the same as in the 5-Mw radar except for the circulator. As the 5-Mw circulator was too bulky, a complete tee-type circulator was employed where the applied magnetic field was above resonance of the ferrite.

Field tests of the two radars have been successfully done through the courtesy of the Japan Self-Defense Force. Sensitivity of the receivers were compared by echoes from a target plane. It was confirmed in 500-kw radar that the radar range was extended by 27 per cent in our TWT radar whose over-all NF was measured as 7.3 db. This was due to 4-db improvement of the NF compared to TR radars.

The authors are indebted to K. Kurihara who developed the ferrite and S. Takeuchi and T. Nagai for making the necessary measurements.

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T. Takeya
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[1] S. Mita, S. Hayasi, Japanese Patent No. 219,555, February 2, 1956; Brit. Patent No. 812,082, April 26, 1960; U. S. Patent No. 2,934,638, April 26, 1960.
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Modified Radar Range Equation Relating Pulse Energy to Antenna Aperture or Gain in a Radar with a Specified Angular Accuracy*

The specifications for new radars usually include not only the requirement of maximum range but also angular accuracy both of which are related to signal-to-noise ratio. The relationship between transmitted power and antenna gain or area for a given maximum range differs from that where both maximum range and angular accuracy must be considered. This difference is discussed briefly in the following.

The usual radar range equation when the same antenna is used both on transmission and reception is

$$R^4 \propto \frac{P_p \tau G^2 \lambda^2 \sigma}{KTL(S/N)} \propto \frac{P_p \tau A^2 \sigma}{KTL(S/N)\lambda^2}$$

where

P_p = peak power transmitted
τ = pulse length
G = antenna gain
λ = wavelength
σ = target area
S/N = signal-to-noise power ratio
K = Boltzmann's constant
T = system noise temperature
L = system losses
A = antenna area.

These expressions imply that, for given values of R, required S/N, T, σ, and L, the energy

$$P_p \tau \propto \frac{\lambda^2}{A^2} \propto \frac{1}{G^2 \lambda^2}.$$

Thus, if the antenna size is changed so that the gain is reduced by 3 db, the required power must be increased by 6 db in order to maintain the range capability. This commonly used relationship applies if maximum range is the prime consideration.

If the basic radar equation is modified slightly, a similar expression is available relating energy to gain or antenna area with both range and rms angular error fixed. The rms angular error in the x plane of an antenna is[1]

$$\sigma_x \propto \frac{\lambda}{D_x} (S/N)^{1/2}$$

where D_x is the x dimension of the antenna. Similarly,

$$\sigma_y \propto \frac{\lambda}{D_y} (S/N)^{-1/2}$$

and

$$\sigma_x \sigma_y \propto \frac{\lambda^2}{D_x D_y} (S/N)^{-1} \propto \frac{\lambda^2}{A} (S/N)^{-1}.$$

Substituting for S/N in the basic radar equation,

$$R^4 \propto \frac{P_p \tau G^3 \lambda^2 \sigma \sigma_x \sigma_y}{KTL} \propto \frac{P_p \tau \sigma A^3 \sigma_x \sigma_y}{KTL\lambda^4}.$$

* Received April 2, 1962.
[1] R. Manasse, "Maximum angular accuracy of tracking a radio star by lobe comparison," IRE Trans. on Antennas and Propagation, vol. AP-8, pp. 50–56; January, 1960.

Now, with $\sigma_x\sigma_y$ fixed, along with R, T, L, and σ,

$$P_{pT} \propto \frac{\lambda^4}{A^3} \propto \frac{1}{G^3\lambda^2}.$$

Returning to the example used previously, if, in this case, the antenna size is changed so that the gain is reduced by 3 db, the power must be increased by 9 db in order to maintain both range and angle measuring capabilities instead of the 6 db required to measure range only.

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A Relation Between α and Q*

In 1944, Davidson and Simmonds[1] derived a relation between the Q of a cavity composed of a uniform transmission line with short-circuiting ends and the attenuation constant α of such a transmission line. Later, in 1950, Barlow and Cullin[2] rederived this relation. These authors showed that this relation is quite general and is applicable to arbitrary cross-section uniform metal tube waveguides. Since then one of the standard techniques for the measurements of the attenuation constant α has become the use of the cavity method.[3] This method offers an excellent way of measuring the attenuation constant of a waveguide when the loss is quite small. Later on this method was generalized and applied to open waveguides,[4] such as the single-wire transmission line and the dielectric waveguide, by various authors.

However, it is noted that the formula by Davidson, etc., was derived under the assumption that there exists a single equivalent transmission line for the mode under consideration. This assumption is true for a pure TE, TM, or TEM mode, but it is not clear that such a single equivalent transmission line exists for a hybrid wave. This suspicion originates from the fact that 1) the TE and TM waves are intimately coupled to each other, and 2) the characteristic impedance defined by Schelkunoff[5] is not constant with respect to the transverse coordinates. It is, therefore, very difficult to conceive the possibility that there exists a single equivalent transmission line for this hybrid wave; at best the hybrid wave may be represented by a set of transmission lines coupled tightly with one another. Hence the formula by Davidson, etc., may not be applicable to a hybrid wave.

A more general relation[6] between Q and α can be obtained without using the transmission line equivalent circuit, provided that α is very small compared with the phase constant β and that the loss contributed by the short-circuiting end plates is negligible compared with the total loss of the waveguide section under consideration. The propagation constant of a guided wave with a small attenuation constant at ω_0 is

$$\Gamma(\omega_0) = \alpha(\omega_0) + i\beta(\omega_0). \tag{1}$$

At resonance, the following relation is true:

$$\Gamma(\omega_0) + \frac{\partial\Gamma}{\partial\omega}\Delta\omega \approx i\beta(\omega_0). \tag{2}$$

Combining (1) and (2) gives

$$\alpha(\omega_0) = -i\frac{\partial\beta}{\partial\omega}\Delta\omega. \tag{3}$$

Since the group velocity v_g and Q are given by the relations

$$v_g = \frac{\partial\omega}{\partial\beta},$$

and

$$Q = \frac{\omega_0}{2(\Delta\omega/i)},$$

one arrives at the relation

$$\alpha = \frac{\omega_0}{2Qv_g} = \frac{v_p}{v_g}\frac{\beta}{2Q} \tag{4}$$

where v_p is the phase velocity of the wave. This is the general relation that is sought.

Substituting the values of v_p/v_g for a TE, TM, or TEM wave into (4), one gets the relations derived by Davidson, etc. For a TM or TE wave in a metal waveguide,

$$\frac{v_p}{v_g} = \frac{1}{1-\left(\frac{\lambda}{\lambda_c}\right)^2}, \tag{5}$$

where λ_c is the cutoff wavelength of the wave under consideration. Thus

$$\alpha_{TE,TM} = \frac{1}{1-\left(\frac{\lambda}{\lambda_c}\right)^2}\frac{\beta}{2Q}. \tag{6}$$

For a TEM wave,

$$\frac{v_p}{v_g} = 1. \tag{7}$$

Hence

$$\alpha_{TEM} = \frac{\beta}{2Q}. \tag{8}$$

For a hybrid wave, v_g and v_p are not simply related. They may be obtained graphically from the ω-β diagram. However, for a dominant hybrid wave on a dielectric rod at very low frequencies or at very high frequencies, the relation (8) is a good approximation since at these frequencies $v_p \approx v_g$.[7]

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* Received April 9, 1962.

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[6] Private communication with Prof. R. W. Gould, Calif. Inst. Tech., Pasadena, Calif.

[7] C. Yeh, "Electromagnetic surface wave propagation along a dielectric cylinder of elliptical cross section," Ph.D. thesis, Calif. Inst. Tech., Pasadena, 1962.

Man-Machine Communication

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Mr. Radcliffe appears to have overlooked most of the important points concerning the man-computer communication problem.[1] He seems to propose a new class of specialist trained to respond to the characteristics of the machine (whatever that may mean). This is precisely the situation we are trying to avoid in the construction of new computer "languages," for example, for reasons which hardly bear repeating. In any case a central point is that the characteristics of computer hardware and software are changing in a time scale incomparable to that of human characteristics. What long-term good, then, would "two semesters of computer pseudo-speech" do a twelfth grader?

Even more strenuous objection might be raised due to what Mr. Radcliffe does not say. First, in referring to man's need "to give up his native language . . . in order to earn his living," one may remark that this is hardly the "last straw" but rather the first since machine language has been constantly giving way to "people-oriented" language. *A fortiori*, to refer in this context to "a machine of lightning speed and powerful memory" is totally inept in the light of the clearly demonstrated properties of self-organizing systems and "intelligent" machines. It is analogous to designing the Niagara Falls power station as though it were to be driven by the steam engine of James Watt. Ultimately the only mismatch which will exist between man and machine will be due to his relatively low upper bound on communication (and *conscious* computation) rate, his fallibility and his need to operate with comparatively informal structures. Every effort must be made to take advantage of whatever sensory motor discrimination a man has,[2] but it is too much to hope that the computer user of the future will adapt very much to the machine.

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* Received April 23, 1962.

[1] A. J. Radcliffe, Jr., "Man-machine communication," Proc. IRE, vol. 50, p. 482; April, 1962.

[2] J. C. Bliss, "Kinesthetic-tactile communications," IRE Trans. on Information Theory, vol. IT-8, pp. 92–98; February, 1962.

P. I. Hershberg[3]

In a recent paper,[1] Radcliffe suggests the interconnection of the human mind to the computing machine for purposes of obtaining advantages absent in either device. In one case, he advises " coupling a living organism in a feedback loop with a computer in a manner devised to condition the organism to communication with the machine." Is not this condition present in the many teaching machines utilizing digital computers, with the reward and punishment signals appearing in the form of wrong and right answers?

Problems involved in man-machine communication are similar to those consequent to machine-machine communication. While both machines may have capabilities realized through megacycle operating speeds, the communication facility between computers may be capable of only kilocycle data transmission rates. The result is that if we wish to have extensive communication between computers, the performance of each must be degraded to that of the communication facility.

This situation is analogous to man-machine communication. If we are to couple the machine to some peripheral nerve of the body, the maximum data rate of the nerve is given by

$$V_n = Kmd^2$$

where V_n is the maximum data rate in bits per second, K is a constant, m is the myelin factor of the nerve fiber and d is the nerve fiber diameter. Since most peripheral nerves have relatively low myelin factors and diameters, the maximum data rate for these nerve fibers is of the order of fifty bits per second.

If the optic nerve is utilized, it must be recognized that a basically slow chemical reaction converts light to nervous impulses (rhodopsin-iodopsin reaction). This may be verified by walking from a bright room to a dark one and noting that it requires a few seconds for the eyes to once again pick up input information.

In contrast with these lower data rates, the human brain has an extremely fast clock rate. For example, it takes no more than a few seconds for the cortex to scan several hundreds of thousands of bits of data "stored" in the human memory. This clock rate is of the order of megacycles and is comparable with the digital computer.

Of course if one is to couple data into the cortex, it is necessary to understand the function of the cortex in terms of its component circuitry. Yet Eccles has stated that there has been little correlation between the enormous amounts of data gathered on the brain and the "activity of cortical neurones."[4] In other words we don't really know anything about how the brain functions in terms of its basic building blocks.

Clearly, a better understanding of the human data processor is necessary before we attempt to couple this device to another machine.

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[3] Received April 26, 1962.

[4] J. C. Eccles, "The Neurophysiological Basis of the Mind," Oxford University Press, New York, N. Y., p. 258; 1953.

Author's Comment[5]

In response to the letter of Mr. Pedelty, I wish to state two of the basic premises of my previous letter:

1) Verbal or written language, man's primary means of conveying concepts which he might wish to communicate to a computer, has evolved over only a relatively short period of time and has been shaped toward a maximum basic survival capability in a crudely hostile environment over most of that period.
2) The speech process and the sonal structure of our various languages are inherently limited by evolutionary processes, extending over a million years, which shaped the structure of our vocal and hearing organs on the basis of an environment encompassing factors quite foreign to the man-machine relationship.

With reference to 1) man has difficulty in conveying concepts clearly to another human due to the uniquely implicit values tied to words by each individual as a consequence of his own living experience. If semantic problems present a barrier to human-human communication, then communication with a machine may similarly be plagued. While humans might communicate with each other purely on the basis of the "dictionary meaning" of words, this is not the actual case. Human communication is assisted materially by interpretations on the part of the listener which are influenced by past experience with the particular speaker. Much meaning is added to speech by accompanying gestures or facial expressions.

My second premise leads to a questioning of the efficiency of speech for man-machine communication. The speech process involves concessions on the part of our vocal organs to limitations of the ear; and very probably the listening process has limitations related to limitations of the vocal organs. These mutual limitations are not inherent in man-machine communications.

It must also be recognized that the human brain has adaptive and learning capability which can be brought to use at rather low cost to overcome limitations in everyday language and speech.

Why then employ expensive input-output machine hardware to circumvent weaknesses in the structure of our language and inefficiencies in the speech process? It appears to me that we will not continue indefinitely to base computer input and output languages (which need not even be the same for man-machine communication) on the language and speech that evolution and circumstance have left for us. When machines of magnitude sufficient to generate genuine semantic problems in man-machine communication have been built, it may prove more practical for man to adapt to the machine, as its potentialities are explored, rather than continually modify a massive piece of hardware.

One of the best ways to train man, as Mr. Hershberg points out, is to employ a "teaching machine" which involves programmed responses by the machine that are conditional upon input signals from the learner. This is feedback of a sort, but not what I had in mind for improving man-machine communication.

Two strangers, motivated to communicate but lacking a common language, may be expected to evolve means for communicating, given sufficient time. Recognizing in advance that a problem of this sort was to be encountered would lead them to preparations that would improve the efficiency of this mutual learning process. It appears that man and machine might be coupled in a similar closed learning loop with hopeful prospects for improved communication as a consequence. The time constants of this closed loop would be a great deal shorter than the time constants of the communication loop, concerned with this same problem, of which this letter is part.

Both Mr. Hershberg and Mr. Pedelty refer to the discrepancy in speed between computation rate in man or machine and in the information rate available in visual, aural or tactile communication channels to or from a human. While we cannot change the propagation velocity of nerve impulses, we can perhaps speed communication by greater emphasis of the parallel mode of operation as opposed to series. There is a great deal more information in a symbol from the Chinese written language than in a letter of the Latin alphabet. Granted that one is more amenable to typesetting than the other, still one presents more information than the other. Is there not a lesson to be learned here in respect to improved communication between man and machine? Similarly, it takes no longer to say crag, steed or gloom than rock, horse or dark, but the former certainly seem to convey more meaning with the same input-output effort.

The associative and adaptive capabilities present in the human brain are not likely to be exceeded by a machine for some time to come. In the meantime it appears to be desirable to determine how man can modify his customary means of communication to improve man-machine communication.

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[5] Received May 8, 1962.

The Discrete Binomial Time Interval Distribution*

The time interval distribution as derived from the Poisson distribution is a useful relation when dealing with random pulses. In some pulse systems, the time interval is restricted to multiples of a finite resolving time. Hence, it would be desirable to have a discrete time interval distribution describing this condition. A discrete distribution is derived from the binomial distribution that can be numerically evaluated from standard tables.

* Received April 11, 1962.

For Poisson-distributed pulses the probability of x pulses in time t is

$$P_x(t) = \frac{(at)^x}{x!} e^{-at} \tag{1}$$

where a is the average pulse rate. The probability of no pulses ($x=0$) in time t is

$$P_0(t) = e^{-at}. \tag{2}$$

If the period t is to be a time interval between successive pulses, the interval must terminate with the next pulse. The probability of a pulse in dt at t is (adt). The combined probability of no pulse in time t followed by a pulse in dt at t is the product of these two probabilities.

$$dP(t) = ae^{-at}dt. \tag{3}$$

The probability of a time interval less than t is obtained by integrating (3)

$$P(t) = \int_0^t dP(t) = (1 - e^{-at}). \tag{4}$$

Note that (3) and (4) are continuous functions.

A discrete relation analog to (3) and (4) can be derived from the binomial distribution. The probability of x pulses in r time increments of length τ is

$$B_x(r) = \binom{r}{x} (a\tau)^x(1 - a\tau)^{r-x} \tag{5}$$

where $\binom{r}{x}$ is the binomial coefficient and $r=t/\tau$ is an integer. The probability of no pulses ($x=0$) in r increments is

$$B_0(r) = (1 - a\tau)^r. \tag{6}$$

The probability of a pulse in the $(r+1)$ increment is $a\tau$. The combined probability of no pulses in r increments followed by a pulse in the $(r+1)$ increment, *i.e.*, the probability of a time interval r increments long is

$$B(r) = a\tau(1 - a\tau)^r. \tag{7}$$

The probability of a time interval less than n increments long is obtained by summing (7)

$$B(n) = \sum_{r=0}^{n} B(r) = 1 - (1 - a\tau)^{n+1}. \tag{8}$$

Eqs. (7) and (8), the desired discrete relations, are analogs to (3) and (4), respectively.

It is readily shown that (4) is the limiting expression for (8), just as the Poisson distribution can be considered as the limiting distribution for the binomial distribution. Writing (8) in terms of t and τ

$$B(n) = 1 - (1 - a\tau)^{(t/\tau)+1}. \tag{9}$$

If $a\tau \ll 1$, $(1-a\tau)$ is approximately $e^{-a\tau}$ and

$$B(n) \sim 1 - e^{-a(t+\tau)}. \tag{10}$$

Letting the increment τ approach zero and n approach infinity such that $t=n\tau$ is finite yields

$$\begin{aligned} B(n) &\sim 1 - e^{-at} \\ B(n) &\sim P(t). \end{aligned} \tag{11}$$

Numerical values for $B(n)$ are given in the Table of the Binomial Probability Distribution.[1] For table use $B(n)$ is rewritten in terms of the binomial expansion as

$$B(n) = \sum_{r=1}^{n+1} \binom{n+1}{r} (a\tau)^r(1 - a\tau)^{n+1-r}. \tag{12}$$

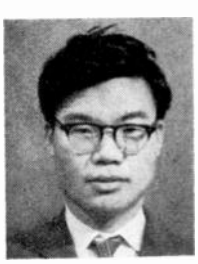
This form is tabulated in the tables.

The preceding relations have proven useful in evaluating sampling rates for digital data systems handling random Poisson-distributed pulses.

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[1] "Table of the Binomial Probability Distribution," Natl. Bureau of Standards, Applied Mathematics Series No. 6, 1950.

Contributors



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Following his graduation he joined the Digital Computer Section of the OKI Electric Company, Japan. He is presently at the University of Tokyo as a Research Assistant in applied physics.

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Lee F. Barash (A'50–M'55) was born in New York, N. Y. on December 28, 1927. She received the B.A. degree in physics from New York University, in 1947, and the M.S. degree in physics from Stanford University, Calif., in 1948.

From 1948 to 1954 she worked in the Electron Tube Department of the Sperry Gyroscope Company, Great Neck, N. Y. She was engaged in research and development of klystrons and traveling-wave tubes. In 1955 she worked at Jasik Laboratories, Westbury, N. Y., on antenna design. During the years 1959 to 1961 she was at Columbia University, New York, N. Y., for graduate study and held a Graduate Assistantship at Columbia Radiation Laboratories. At present she is retired for the third time.

Mrs. Barash is a member of the American Physical Society and Sigma Xi.

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Edward Bedrosian (S'51–A'53–M'54–SM'56) was born in Chicago, Ill., on May 22, 1922. He received the B.S.Ae.E. degree from the Aeronautical University, Chicago, in 1942, and the B.S.E. and M.S. degrees, from Northwestern University, Evanston,

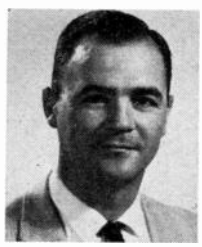


Ill., in 1949 and 1950, respectively. The study for the M.S. degree was conducted under a fellowship granted by the Airborne Instruments Laboratories. He received the Ph.D. degree from Northwestern, in 1953.

After receiving the B.S.Ae.E. degree, he worked as an Aeronautical Engineer at Consolidated-Vultee, San Diego, Calif., and Hughes Aircraft Company, Culver City, Calif. During World War II he served in the U. S. Marine Corps as a Radar Technician. He joined the research staff of the Communication and Electronics Division of Motorola, Inc., Chicago, in 1953, where he engaged in studies of antennas and communications systems. Later he became a Senior Staff Member at the company's research laboratory in Riverside, Calif., and conducted missile and radar systems analyses. Since 1957 he has been a member of the Electronics Department of the RAND Corporation, Santa Monica, Calif., participating in studies of antennas and communications systems.

Dr. Bedrosian is a member of Eta Kappa Nu, Tau Beta Pi, Pi Mu Epsilon, and Sigma Xi.

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Hewitt D. Crane (S'47–A'48–M'55–SM'57) was born in Jersey City, N. J., on April 27, 1927. He received the B.S. degree from Columbia University, New York, N. Y., in 1947, and the Ph.D. degree from Stanford University, Calif., in 1960, both in electrical engineering.

During 1945 and 1946 he served in the U. S. Navy. From 1947 to 1949 he worked for the David Bogen Company, New York, N. Y., and did development work on facsimile equipment for Western Union Telegraph Company, New York. For the next two and one-half years he was involved in maintenance of IBM's SSEC computer. From 1951 through 1954, at the Institute for Advanced Study, Princeton, N. J., he was concerned with the over-all logic of the IAS computer and was in charge of designing and building the input-output equipment, a graphing device, and a magnetic secondary storage device. He then worked a year at the David Sarnoff Research Center, Princeton, on magnetic devices, in particular on multiaperture magnetic devices and a large-scale magnetic memory. He joined the staff of Stanford Research Institute, Menlo Park, Calif., in 1956. During his first six months at the Institute he was concerned with an analysis of the ERMA logic. Subsequently he initiated work on the development of techniques for the realization of digital logic circuits with all magnetic circuitry, in particular, using MAD's (multiaperture magnetic devices). More recently, he has initiated a program of study of a new class of computer devices termed neuristors.

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Charles A. Hauer (M'57) was born in Erie, Pa., on May 26, 1932. He received the B.S.E.E. degree (*cum laude*) from Pennsylvania State University, University Park, in 1954.

He then joined the Philco Research Division (now called the Scientific Laboratory). From 1954 to 1956 he was engaged in a study of communication systems where he analyzed methods of modulation and detection, studied methods for reducing the vulnerability of communication systems to natural and man-made interference, and made extensive evaluations of existing voice, digital, and multiplex communication systems. From 1956 to 1959 he was responsible for the conception and design of a jam-resistant data link (Spread Eagle). During 1958, in a systems analysis of the AN/GRC-59 microwave delay, he made a detailed study of FM transient response. In 1959 he managed a program to reduce the time-bandwidth product required for transmission of pictorial data. He managed a theoretical investigation of radar resolution from June, 1960 to May, 1961. From May, 1961 to May, 1962, he managed a program of research and investigation of advanced sensor techniques for space-intercept vehicles. At present he is a Research Specialist at the Philco Scientific Laboratory and is engaged in studies of radar resolution and accuracy. His specialties are data processing, signal detection and system analysis as applied to both radar and communication systems.

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John F. Koval (S'57–M'59) was born in Cleveland, Ohio, on July 16, 1936. He received the B.S. degree in electrical engineering from Ohio University, Athens, Ohio, in 1958, and received the M.S. degree in electrical engineering from Drexel Institute of Technology, Philadelphia, Pa., in 1962.

He joined the Philco Research Division (now called the Scientific Laboratory) in 1958. He has engaged in the development of a low-noise, high information-rate sensor technique for interception of ICBM targets, signal-to-noise analysis of multiplex FM systems, advanced studies on spread spectrum communications systems, development of coaxial and waveguide parametric amplifiers, and design and development of circuitry for airborne moving target indicators.

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John D. Kraus (A'32–M'43–SM'43–F'54) was born in Ann Arbor, Mich., on June 28, 1910. He received the B.S. degree, in 1930, the M.S. degree, in 1931, and the Ph.D. degree, in 1933, all from The University of Michigan, Ann Arbor.

From 1934 to 1935 he did research on industrial noise-reduction problems, and from 1936 to 1937 was engaged in nuclear research with the newly completed University of Michigan cyclotron. From 1938 to 1940 he was an Antenna Consultant. He joined the Naval Ordnance Laboratory, Washington, D. C., in 1940, working on the degaussing of ships, and in 1943 he became a member of the Radio Research Laboratory, Harvard University, Cambridge, Mass. He joined the faculty of The Ohio State University, Columbus, in 1946, where he is now Professor of Electrical Engineering, Professor of Physics and Astronomy, and Director of the Radio Observatory. He is the inventor of the helical beam antenna, the corner-reflector antenna, and other antenna types. He is the author of textbooks on antennas and electromagnetic theory.

Dr. Kraus is a member of the American Astronomical Society, the American Rocket Society, the American Physical Society, the American Geophysical Union, the International Astronomical Union, and the International Scientific Radio Union.

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Jin-ichi Nagumo (M'60) was born in Tokyo, Japan, on October 14, 1926. He received the Ph.D. degree in applied physics from the University of Tokyo, Tokyo, Japan, in 1954.

From 1953 to 1959, he was an Assistant Professor of Applied Mathematics at the Keio University, Tokyo, Japan. Since 1959 he has been an Assistant Professor of Applied Physics at the University of Tokyo. Since 1948 he has been working on the theory of nonlinear oscillations, nonlinear circuit theory and nonlinear electronics; since 1959 he has been working on cybernetics, bionics and bio-medical electronics.

Dr. Nagumo is Vice-Chairman of the Professional Group on Nonlinear Circuit Theory of the Institute of Electrical Communication Engineers of Japan.



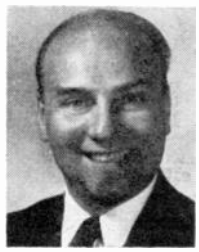
Akira Okaya was born in Kobe, Japan, on October 10, 1920. He received the Ko-Gakushi degree in aeronautical science (corresponding to the Master's degree) from Nagoya University, and the Rigaku-Hakase degree (corresponding to the Ph.D. degree) from Kyoto University, in 1956, his thesis was on millimeter microwave spectroscopy.

During World War II he was engaged in the design of aircraft at the Japan Naval Aeronautical Research Laboratory in Yokosuka. He started his career in physics in 1946 at the Physics Department of Kyoto University. An Instructor there from 1948 to 1957, he began work on microwave spectroscopy in 1950, a study which continues to absorb much of his attention. He was a member of the Japanese Government Committee on Radio Technique and engaged in atomic clock experiments in Japan prior to 1956. In 1955 he became a Research Associate in the Chemistry Department at Columbia University, New York, N. Y., on a Fulbright travel grant, where he worked with nuclear magnetic resonance of liquids and gases. In 1957 he became a staff member of the Columbia Radiation Laboratory (connected with the Physics Department) where, for four years, he was engaged in work on gas maser spectroscopy, the solid-state maser, electron paramagnetic resonance, and microwave electronics. He joined the IBM Communications Systems Center, Rockville, Md., in September, 1961, and is now doing research in advanced communications systems, especially laser and special light communications.

Dr. Okaya is a member of the American Physical Society and the Physical Society of Japan.

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Charles Süsskind (S'47–A'52–M'53–SM'54) was born in Prague, Czechoslovakia, on August 19, 1921. He received the B.S. degree from the California Institute of Technology, Pasadena, in 1948, and the M.Eng. and Ph.D. degrees from Yale University, New Haven, Conn., in 1949 and 1951, respectively.



During the war he served with the Eighth Air Force in Europe as an airborne-radar specialist. His interest in the microwave field led to a doctoral thesis on artificial dielectrics. An article based on this dissertation, published by the *Journal of the British IRE*, earned him the Clerk-Maxwell Premium for "the most outstanding paper" published in the Institution's Journal in 1952. From 1951 to 1955, he was a Research Associate at Stanford University, Calif.; between 1953 and 1955, he was also Assistant to the Director of Stanford's Microwave Laboratory and Lecturer in the Department of Electrical Engineering. In 1955 he joined the faculty of the University of California at Berkeley, where he is now Associate Professor of Electrical Engineering. He is editor of the "Encyclopedia of Electronics" (Reinhold, 1962).

Dr. Süsskind is an associate member of the British IRE, and a member of the American Physical Society, the ASEE, The History of Science Society, the Society for the History of Technology, Sigma Xi, and Tau Beta Pi.

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Martti E. Tiuri (M'61) was born in Koski, Finland, on November 13, 1925. He received the B.S. degree in electrical engineering from the Institute of Technology, Helsinki, Finland, in 1950, the M.S. degree from Stanford University, Calif., in 1956, and the Doctor of Technology degree from the Institute of Technology Finland, in 1960.

Since 1951 he has been Research Engineer in the State Institute for Technical Research, Helsinki, Finland, where his work has been primarily in microwaves, and low-noise receivers. From 1961 to 1962 he has been a Visiting Professor at the Radio Observatory of The Ohio State University, Columbus.



Harry Urkowitz (S'48–A'49–SM'54) was born in Philadelphia, Pa., on October 1, 1921. He received the B.S. degree in electrical engineering from the Drexel Institute of Technology, Philadelphia, Pa., in 1948, and obtained the M.S. degree in electrical engineering from the University of Pennsylvania, Philadelphia, in 1954.

In World War II he served as a Radar Navigator with the U. S. Army Air Forces. In 1948 he joined the research division (now called the Scientific Laboratory) of the Philco Corporation, where he now holds the rank of Senior Research Specialist. He has engaged in research and development in ferroelectric amplifiers, noise and clutter in radar and communication systems, analysis and design of special purpose video and IF amplifiers, analysis and synthesis of special filters, including delay line periodic filters for use in radar communication systems, transient analysis of FM systems, analysis and design of airborne and ground-based MTI systems, detection of weak signals in noise and clutter and the study of randomly time-varying systems. He also lectures in the Graduate School of Engineering at Drexel, where he presently teaches a course in random processes.

Mr. Urkowitz is a member of Phi Kappa Phi, Tau Beta Pi and Eta Kappa Nu.

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Shūji Yoshizawa was born in Kanagawa, Japan, on November 27, 1938. He received the B.S. degree in applied physics from the University of Tokyo, Tokyo, Japan, in 1962.

He is now a post-graduate student engaged in studies of nonlinear circuit theory at the University of Tokyo.

Mr. Yoshizawa is a member of the Institute of Electrical Communication Engineers of Japan.

Books

High Magnetic Fields: Proceedings of the International Conference on High Magnetic Fields, 1961, H. Kolm, B. Las, F. Bitter, R. Mills, Eds.

Published (1962) by John Wiley and Sons, Inc. 440 Park Ave. S., New York 16, N. Y. 736 pages +xv pages. Illus. 6½×9¼. $15.00.

This is an important and in some ways impressive book. The reader is apt to be disappointed, however, unless he takes its nature into account. It is actually the *Proceedings* of a recent large topical Conference devoted to the technology of magnetic fields. The nature of a *Proceedings*, especially one such as this, in which essentially all papers were published, is that there is no especially logical development of the subject, or comprehensive and orderly discussion of the progress in the technology. Rather there is an almost fantastic array of papers on many, many aspects of magnetic fields.

An attempt has been made by the editors to establish a logical grouping of the material, although it is at variance with the order of presentation at the actual Conference. Whether the regrouping has really helped much is somewhat questionable.

The book deals with the theory and then the practice of the generation of steady (dc) high magnetic fields. This is followed by a treatment of pulsed fields and pulsed-power sources. Superconducting magnets are discussed in two places, partly in Section I D, (magnets) and partly in III C (materials). This division had to be rather arbitrary as the speakers in both sections really discussed both topics. Parts II and III A and B include a review of many fine experimental programs in numerous fields of physics, all related by their requirement for high magnetic fields. Concluding papers are on plasma research, fusion research, and particle physics research (Part IV, A, B, and C).

The greatest strength of this work is that it is an almost inexhaustible source of reference and bibliographical material in the subject of magnetic fields. There are a few papers that are very distinguished, such as Professor Bitter's review of the present and future of high-field technology, and there are a number of elegant discussions of research underway or possible with high fields.

One cannot help mentioning that there is also some nondescript science that crept into the Conference and subsequently into the book. There is little science and too much bureaucratic boasting in the papers on the one minute per 24 hours neon-cooled magnet.

The use of superconducting materials in the design of magnets has become important in the last two years, with the discovery of hard superconductors which retain their properties in the presence of high magnetic fields. The superconducting papers in this book constitute a good summary of the state of the technology at the time of the Conference.

One aspect of these *Proceedings* that is undesirable is that the Conference discussion has been entirely omitted. One of the greatest virtues of topical conferences is that it permits a lively interchange of up-to-the-minute professional opinion. Most *Proceedings* include at least selected portions of this valuable commentary. Here it has regrettably been completely ignored.

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Radio Surface Waves, by H. M. Barlow and J. Brown

Published (1962) by Oxford University Press, 417 Fifth Ave., New York 16, N. Y. 170 pages+4 index pages+xi pages+16 appendix pages. Illus. 6½×9¼. $6.75.

Professor Barlow and Dr. Brown have made many important contributions to the subject of electromagnetic surface waves in recent years. Therefore, it is quite fitting that they should co-author a monograph on surface waves. Essentially it is a detailed exposition of their own views on this rather controversial subject. The electromagnetic waves discussed in the monograph are of a non-radiating type. Thus, the subject matter really deals with *evanescent* waves, and this might have been a better title for the book. It is just this evanescent character which is the basis of a working definition of a "surface wave," although as the authors point out, not all evanescent fields are associated with surface waves. Explicitly, their definition reads: "A surface wave is one that propagates along an interface between two different media without radiation; such radiation being construed to mean energy converted from the surface wave field to some other form." As the authors admit, the definition is highly idealistic and realizable only with a smooth and plane interface between homogeneous media. They propose that waves on actual structures, such as when a wave propagages around a smooth cylindrical supporting surface, be regarded as a "perturbed surface wave." (No suggestions are put forth as to how one would classify the various field components produced by a dipole, in the presence of a plane interface between homogeneous media.)

The book treats, in detail, the conditions for the support of surface waves on various planar and cylindrical structures. Many interesting physical descriptions are put forth, such as the significance of a complex Brewster angle and its relation to surface waves. By first leaving aside the question of excitation, the early chapters use only quite elementary methods. Later on in the book the question of launching of surface waves is considered in an elegant fashion. (Unfortunately, however, the reader is never told if the Zenneck surface wave can be excited over a homogeneous conducting half-space.)

Most noteworthy is the material on approximate methods of analysis and design procedures for surface wave antennas. Such problems as discontinuities in surface properties and the resultant radiation are dealt with by several approaches, which have various degrees of rigor. Furthermore, the analytical data are checked by careful microwave experiments to a high precision. It is in this area that the authors are in their element.

On the whole the book is clearly written and is a worthwhile purchase. The quality of the printing and illustrations is excellent.

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Principles and Applications of Electromagnetic Fields, by R. Plonsey and R. Collin

Published (1961) by McGraw-Hill Book Co., Inc., 330 W. 42 St., New York 36, N. Y. 554 pages+10 index pages+xvi pages+1 bibliography page+problems. Illus. 6½×9½. $12.75.

This is a very complete and well-written text on electricity and magnetism, designed for use at about the senior year in a modern Electrical Engineering or Applied Physics curriculum. It is written in the "standard" manner, *i.e.*, electrostatics and magnetostatics treated separately, followed by time varying field and Maxwell's equations. The last third of the book takes up applications of Maxwell's equations to plane waves, waveguides, cavities, radiation including antennas, and the interaction of charged particles with fields. This last chapter includes such topics as Llewellyn-Peterson equations, the klystron, and space charge waves. There is also a very short introduction to magnetohydrodynamics, but no treatment of relativity or radiation from higher multipoles. In addition to these electrical topics there is also a thorough discussion of the necessary vector analysis, the solution of Bessel's and Legendre's equations, and the properties of these functions. The presentation is very well done, being neither too wordy nor too concise, but pitched at just about the right level for intelligent seniors or poorly prepared first-year graduate students. Many excellent problems are included.

This book covers so many topics that, as a whole, it would make a rather long course. It is so constructed, however, that many topics can easily be left out. Many modern electrical engineering curricula, for instance, treat vector analysis, the advanced functions and conformal transformations in a separate course in "Engineering Mathematics," which precedes the course in Electricity and Magnetism. In a certain sense this book is too complete. The reviewer does not believe that many courses will be taught which will go through the book from beginning to end. On the other hand, just this completeness will make the book an excellent reference, as well as a very good text.

In conclusion, it is very well written, very complete, and highly recommended.

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Radio-Electronic Transmission Fundamentals, by B. Whitfield Griffith, Jr.

Published (1962) by McGraw-Hill Book Co., Inc., 330 W. 42 St., New York 36, N. Y. 602 pages+xi pages+10 index pages. Illus. 6¼×9¼. $10.75.

This book is an intermediate-level text which concentrates on the generation and transmission of electrical energy at radio frequencies. It should be regarded as a textbook and its pedagogical value depends, of course, on the orientation of the course and its instructor. The text consists of four parts which are subdivided into sixty-eight chapters. The parts cover topics of electrical networks, transmission lines, radio antennas and radio transmitters.

The author very carefully avoids the use of higher mathematical methods not necessary for the everyday problems encountered in this line of work. The technicians and radio operators for whom this book is intended will find presented a good geometrical interpretation of complex numbers, and a very brief introduction to the calculus. Emphasis is placed on the geometrical solution of problems as opposed to a purely mathematical concept. The text is accurate, its arrangement is orderly, the progression from one topic to the next is logical, and the most advanced sections build on that which precedes.

A beginning student will find that the text includes far too few practice problems. Problems are absolutely essential if the student is to assimilate the book's methods for later application to his own problem situations. The author could have also included more sample problems to augment private study.

Mr. Griffith has presented his subject sufficiently to stimulate the student, whet his appetite, and give him something more to think about than sets of equations. Any technician interested in broadening his knowledge will find this volume an indispensable addition to his library.

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Engineering Aspects of Magnetohydrodynamics, Clifford Mannal and Norman W. Mather, Eds.

Published (1962) by Columbia University Press, 2960 Broadway, New York 27, N. Y. 570 pages+xiv pages+8 appendix pages+references by chapter. Illus. 6¼×9¼. $13.50.

"Engineering Aspects of Magnetohydrodynamics" is an edited version of the thirty-four papers presented at the Second Symposium on Engineering Aspects of Magnetohydrodynamics, at the University of Pennsylvania, March, 1961. Consequently the volume is intended almost exclusively for workers active in this field, and perhaps for a few others who will relish several of the review articles which it contains.

Four major areas are covered, namely: Flight Applications, Power Conversion, Communications and Diagnostics, and Fusion. The Flight Applications section deals principally with experimental investigations concerned with devices for the acceleration of plasma. The portion on Power Conversion commences with several review papers on the state-of-the-art at the time of the meeting, and concludes with more specific topics (*e.g.* consideration of ac mhd-generation, conduction in gases with elevated electron temperature, etc.). The Communications and Diagnostics section considers propagation of electromagnetic waves through plasmas, nonlinear properties of plasmas, space charge waves and other diagnostic techniques. Finally, the portion of the book dealing with Fusion contains an extensive treatment of the economics of fusion reactors, and a wide assortment of aspects relating to the thermonuclear field (*e.g.* superconductivity, shear heating, transient Astron reactor, etc.). In addition, an introduction to each section by the Chairman of the corresponding session of the Symposium is included. A list of the papers presented at the First Symposium on Engineering Aspects of Magnetohydrodynamics, which were never published in book form, is also included.

A considerable amount of experimental data is contained in the papers, and the description of untried facilities and experimental configurations are kept to a minimum. Although some of the material in the book has appeared in other parts of the literature, the coverage presented is, in general, more complete, and constitutes a fuller account than was possible in oral presentation of the papers. As such it provides an excellent account of the Symposium and is one of the best edited versions of any symposium proceedings yet available. It will make a valuable addition to the library of the specialist in the field of magnetohydrodynamics. Since further series of this meeting are to be published with the same editors, a high degree of clarity and uniformity is insured.

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The Mathematical Theory of Linear Systems, by B. M. Brown

Published (1961) by John Wiley and Sons, Inc., 440 Park Ave., S., New York 22, N. Y. 267 pages+3 index pages+x pages+exercises and answers. Illus. 5½×9. $8.00.

This is the second of the books appearing in the Automation and Control Engineering Series, edited by John F. Coales. The objective of this series is to provide textbooks on principles, as opposed to application, of control engineering. A most important type of control system found in practice is one which can be characterized adequately by a set of linear differential equations. The manipulation and understanding of such equations is basic for those who would analyze and design linear systems.

Included in the treatment of linear systems is the operational method, which is handled as in elementary textbooks on differential equations. This approach leads, quite naturally, to the single-sided Laplace transform. Following this, there is an exposition of Fourier series and Fourier transform and the use of the bilateral Laplace transform. The definition of linear systems is then carefully given, along with their properties, including stability criteria. Sampled-data systems including the analytical method of difference operators, z-transform, and other techniques are discussed. The theory of linear systems subjected to random inputs and their optimization is covered, as well as definitions and basic properties of random signals.

In summary, the various mathematical techniques of treating linear differential and difference equations which describe linear dynamic systems are well covered. The book is written in the concise style typical of English technical textbooks. Though at an elementary level, the reader must be mature enough to fill in the gaps. There are a few exercises for the reader to test his mastery of the subject matter. A number of explanatory appendixes do help complete the understanding of the subject.

Generally, the book is competently written and brings together many well-known techniques in one volume. It is suitable as an undergraduate textbook but it does need support in the form of classroom exposition by the teacher. Most engineers in the field should already be completely competent in the subject matter presented. For them, this book will serve as a review or refresher textbook.

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Transient and Steady-State Analysis of Electric Networks, by Edward Peskin

Published (1961) by D. Van Nostrand Company, Inc., 120 Alexander St., Princeton, N. J. 412 pages +11 index pages +xiv pages+references and problems by chapter. Illus. 6¼×9¼. $13.00.

This book was written as a textbook for advanced undergraduates or first-year graduate students. Its unified approach to the transient and steady-state solution of linear network problems recommends its use early in the training of electrical engineering students. It should also be a useful refresher book for practicing engineers who first learned ac circuit theory, and later took a course in transient analysis which dealt primarily with pulse waveforms. The first chapter reviews the basic definitions and laws pertaining to linear, lumped, finite networks. The classical solution of the differential equations of a simple network is compared to the ac steady-state solution, as an introduction to the Heaviside operational calculus method.

The Laplace transform is then introduced in the second chapter after a review of Fourier analysis. Finally, some useful properties of complex valuables are introduced in Chapter 3, before general network analysis is undertaken in the following chapters. Chapters 4–7 cover the usual network theorems, loop and node analysis, poles and zeros, and some procedures of matrix algebra, all in terms of the Laplace operator rather than complex impedances. The ability of the Laplace operator to give both the transient and the steady-state response is thus emphasized. Chapter 8 concludes the book by presenting material on signal flow graphs, which should be of interest to most engineers.

This is a very up-to-date treatment of network analysis and a most useful addition to the literature.

G. B. Herzog
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The Junction Transistor as a Switching Device, by C. Le Can, K. Hart, and C. De Ruyter

Published (1962) by Reinhold Publishing Corp., 430 Park Ave., New York 22, N. Y. 213 pages+ix pages+3 index pages+9 appendix pages. Illus. 6¼×9. $10.00.

This book is based on some of the work which has been done in the Semiconductor Electrical Department Laboratory of Philips, in Niijmegen, Netherlands. The author has attempted to establish a link between semiconductor physics and semiconductor application, in the area of large signal behavior of alloy junction diodes and transistors. This he has done very adequately by considering the majority-carrier concentration patterns.

The first chapter of the book is devoted to the electrical behavior of junction diodes. It discusses the physics of the alloy junction and covers the effects of bias, steady-state current flow, temperature, and the current flow mechanisms. The majority-carrier concentration pattern approach is developed. The second chapter is concerned with the steady-state behavior of alloy junctions. It develops the basic current-voltage relationships, current amplification factors, and the transistor characteristics. The third chapter gives the transient behavior and fundamental transistor parameters.

The fourth chapter is concerned primarily with equivalent circuits and applications. Here the author explains some of the circuit effects of transistors which appear abnormal. The grounded emitter and grounded collector stages are analyzed for response to resistive, capacitive and inductive loads.

It is the opinion of the reviewer that this book will be very valuable to circuit and device designers, development engineers, and as a supplementary text for engineering seniors or graduate students. It is highly recommended for people in circuit applications, where transient or transistor switching behavior is of interest.

ARTHUR R. MOORE
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Handbook of Semiconductor Electronics, Lloyd P. Hunter, Ed.

Published (1962) by McGraw-Hill Book Co., Inc., 330 W. 42 St., New York 36, N. Y. 665 pages+18 index pages+xxxvii pages+162 bibliography pages +references by chapter. Illus. 6¼×9¼. $18.50.

This handbook is a valuable addition to the literature of semiconductor electronics. The contents are divided approximately into four major parts: 1) physics of transistors, diodes, and photocells; 2) technology of transistors, diodes, and photocells; 3) circuit design and application for transistors, diodes, and photocells; and 4) reference material.

The text will be of value to students for introductory explanations of the physics of electronic conduction in solids, rectification, and transistor action. Graphical representations of energy levels with the carrier density distributions are noteworthy for clarity. The advanced worker will find a source of information on special topics, and an extensive bibliography to lead him quickly to the pertinent literature.

Each major section is written by an expert in the field. For example, the surveys on materials preparation and methods for making *P-N* junctions are probably the most complete up-to-date surveys available.

It can be expected that a book of reasonable size cannot cover everything of importance. As the major subdivisions suggest, this handbook is confined to transistors, diodes, and photocells. Topics such as magnetoresistance and associated devices, Hall effect devices, thermoelectrics, and controlled rectifier applications are not discussed.

Since the handbook has many authors, integration is difficult. This reviewer looked for a discussion of the space-charge control theory of transistor operation in the transistor action section, but found it instead under high frequency and video applications. This example is used to point out that a search for a given topic should not be confined to one part of the handbook.

The circuit design and applications sections are very complete, and should be valuable to students and specialists alike. These sections are supplemented in the reference section by a survey of methods of circuit analysis. The reference materials on measurement of device and materials parameters would be difficult to find in one survey article. As usual, the literature references are extensive.

A comprehensive bibliography is attempted. It is the opinion of the reviewer that this should not be attempted in the Third Edition. Perhaps the early years of transistor development should be retained, but special bibliographies are now being published for the solid-state field.

The editor, Dr. Lloyd Hunter, is to be commended for his thorough revision of the handbook, and the authors are to be thanked for their valuable contributions.

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RECENT BOOKS

Alexander, J. E. and Bailey, J. M. *Systems Engineering Mathematics.* Prentice-Hall, Inc., Englewood Cliffs, N. J. $13.00. Discussion of the behavior of electrical, mechanical, fluid, and thermodynamic elements and the mathematics for analyzing systems made up of these elements.

Eckman, Donald P. *Systems: Research and Design.* John Wiley and Sons, Inc., 440 Parr Ave. S., New York 16, N. Y. $8.50.

Forrester, Jay W. *Industrial Dynamics.* The MIT Press, Cambridge, Mass. and John Wiley and Sons, Inc., 440 Park Ave., S., New York 16, N. Y. $18.00. A result of the first five-year phase of a research program at MIT School of Industrial Management, the book is intended both as a classroom text and as a practical guide for managers. It analyzes the management process through evaluation of realistic models that incorporate decision-making policy and the interactions of information, money, orders, material, personnel, and capital equipment.

LaFuze, David L. *Magnetic Amplifier Analysis.* John Wiley and Sons, Inc., 440 Park Ave. S., New York 16, N. Y. $9.75.

Parzen, Emanuel, *Stochastic Processes.* Holden-Day, Inc., 728 Montgomery St., San Francisco, Calif. A textbook on the theory and application of stochastic processes.

Samaras, Demetrios G. *Theory of Ion Flow Dynamics.* Prentice-Hall, Inc., Englewood Cliffs, N. J. $16.00. The basic facts of the theory of ion flow dynamics are explored and a number of typical examples of their application are examined.

Savant, C. J., Jr. *Fundamentals of the Laplace Transformation.* McGraw-Hill Book Co., Inc., 330 W. 42 St., New York 36, N. Y. $7.75. An undergraduate text presenting the transform method of analysis from the point of view of the engineer.

Shiers, George. *Electronic Drafting.* Prentice-Hall, Inc., Englewood Cliffs, N. J. $11.70. Interpretation of industrial drafting practices for the beginner. Introduces the various drafting techniques and types of drawing used in the design and construction of electronic equipment.

Van de Hulst, H. C., de Jager, C., and Moore, A. F., Eds. *Space Research II: Proceedings of the Second International Space Science Symposium, Florence, 1961.* John Wiley and Sons, Inc., 440 Park Ave., S., New York 16, N. Y. $29.50.

Scanning the Transactions

The Eternal Search for faithful recording and reproduction of sound and other information has led in many directions. A giant step was taken with the development of magnetic tape. This versatile method of recording has seen diverse applications in many areas besides the high-fidelity reproduction of sound; for example, data of a great variety are often recorded on magnetic tape. Such data are sometimes analyzed by playing the tape back very slowly, an action convenient to the analysis but very demanding of the low-frequency response of the entire playback system. Other applications of tape likewise require good reproduction of low frequencies, ideally down to zero frequency. Magnetic-tape pickup heads commonly used for playback have a basic limitation imposed on their low-frequency response by the principle by which they operate: they are sensitive to rate of change of the flux picked up from the tape, and hence their output falls off as the frequency decreases, becoming zero at zero frequency. If a pickup were used which was sensitive to magnetic flux rather than to its rate of change, this limitation would be eliminated and the frequency-response curve would be flat down to zero frequency.

Analytical and experimental investigations have been carried out using Hall-effect heads. Such devices consist of a semiconductor element which is placed where magnetic flux from the tape can act upon it. Fundamentally the element will have four terminals, two for impressing a steady direct current, and two for extracting the output Hall voltage whose magnitude is dependent on the magnetic flux. A number of variations of geometry and arrangement of connections are possible. The linearity of the experimental Hall pickups was found to be practically perfect. The sensitivity is a function of the semiconductor material used; indium antimonide proved to be superior to other materials tested. An efficient magnetic design is more important for Hall-effect heads than for ordinary types since the low output voltages—roughly a fraction of a millivolt—do not allow a margin of safety. The experimental devices fully met expectations for low-frequency response; exhibited high-frequency response which, although excellent, showed a need for high-frequency equalization; and had a signal-to-noise ratio of more than 50 db over audio bandwidths. (M. Camras, "Some Experiments with Magnetic Playback Using Hall-Effect Sensitive Elements," IRE TRANS. ON AUDIO, May–June, 1962.)

Are you reading more now but enjoying it less? Problems caused by an expanding technical literature include, besides eyestrain, difficulty in recalling needed information, innocent duplication of effort, and the frustration experienced by any engineer or scientist struggling to keep abreast of new developments.

Estimates based on lists used by abstracting services indicate that there are about 1000 serial publications having information of interest to electronics engineers. These provide us with from 30,000 to 70,000 articles a year. If we take the low figure, and decide that only one-third of these are worthy of attention, we are left with a daily reading assignment of 27 articles. The United States is by far the most prolific publisher in the electrical engineering field, followed by Great Britain, Germany, the USSR, and either Japan or France, depending on the estimating method used. Almost two thirds of the E. E. literature is in English, 11.5 per cent in German, and 8.2 per cent in Russian. (C. P. Bourne, "Problems Posed by an Expanding Technical Literature," IRE TRANS. ON ENGINEERING WRITING AND SPEECH, August, 1962.)

More on Reading. With the prodigious amount of technical literature available, what periodicals does the technical and scientific worker actually subscribe to and read? A readership survey covering 2400 professional employees at nine U. S. research and development laboratories has revealed some very interesting data. Although persons in all specialties were included, they were primarily concerned with rocket and space activities and antisubmarine defense. About 35 per cent were electrical and electronics engineers, with the remainder heavily concentrated in physics, chemistry, and other engineering disciplines. The project was initiated to determine in which of the well-known technical, business, and news publications professional employment advertisements could most effectively be placed. Thirty-seven periodicals which carry such advertisements (thus excluding publications such as the TRANSACTIONS) were incorporated in the survey. The results showed that reading habits can be readily differentiated by level of academic attainment and academic major. Lesser but important influences on readership appear to be pay level and type of work within the various fields (research, test, or development). *Time* proved easily the most popular magazine from the standpoint of subscriptions, but in terms of actual claimed readership it was outdistanced by *Scientific American*. Both of these magazines showed a pronounced rise in claimed readership as the level of formal education increased from no degree to the Ph.D. Furthermore, subscribers having no degree appear to tend more toward the electronic trade magazines than those with more formal education. The following table lists the ten periodicals in order of subscription popularity for all professional workers at the surveyed laboratories (column A), and for electrical and electronics engineers only at the same locations (column B):

A	B
Time	PROCEEDINGS OF THE IRE
Scientific American	*Electronic Design*
Electronic Design	*Electrical Design News*
PROCEEDINGS OF THE IRE	*Time*
Newsweek	*Electronics*
Electrical Design News	*Scientific American*
American Scientist	*Western Aviation*
Electronics	*Newsweek*
Western Aviation	*Electrical Engineering*
Physics Today	*Electronic News*

(J. W. Hoyt, "Periodical Readership of Scientists and Engineers in Research and Development Laboratories," IRE TRANS. ON ENGINEERING MANAGEMENT, June, 1962.)

Merger may not come to pass until next January, but the IRE and AIEE student publications have already blazed a trail in that direction. The September, 1962 issues of the IRE STUDENT QUARTERLY and the AIEE's *EE Digest* were combined into a joint publication with the not-too-surprising title of STUDENT QUARTERLY AND EE DIGEST. A new name will soon be adopted for this merged student journal, and it is planned that it will be issued six times yearly. It seems very appropriate that the first concrete step toward consolidation of our two societies should take place in the area of student affairs where for a number of years an unusually close working relationship has existed.

Abstracts of IRE Transactions

The following issues of TRANSACTIONS have recently been published, and are now available from the Institute of Radio Engineers, Inc., 1 East 79 Street, New York 21, N. Y., at the following prices. The contents of each issue and, where available, abstracts of technical papers are given below.

Sponsoring Group	Publication	IRE Members	Libraries and Colleges	Non-Members
Antennas and Propagation	AP-10, No. 4	$2.25	$3.25	$4.50
Audio	AU-10, No. 3	2.25	3.25	4.50
Automatic Control	AC-7, No. 4	2.25	3.25	4.50
Communications Systems	CS-10, No. 1	5.00	7.50	10.00
Electronic Computers	EC-11, No. 2	2.25	3.25	4.50
Engineering Management	EM-9, No. 2	2.25	3.25	4.50
Engineering Writing and Speech	EWS-5, No. 1	2.25	3.25	4.50
Military Electronics	MIL-6, No. 3	2.25	3.25	4.50
Product Engineering and Production	PEP-6, No. 2	2.25	3.25	4.50
Reliability and Quality Control	RQC-11, No. 2	2.25	3.25	4.50

Antennas and Propagation

VOL. AP-10, NO. 4, JULY, 1962

Biographies—(p. 107)

Radiation Fields from a Horizontal Electric Dipole in a Semi-Infinite Conducting Medium—Albert W. Biggs (p. 358)

The radiation fields of a horizontal electric dipole in a semi-infinite conducting medium are developed to yield a ground wave near the surface and a space wave above it. Previous work for points of observation slightly above the conducting medium is extended to the entire region by including the height of the observation point in the evaluation of the integral by the saddle point method. The effect of burying a horizontal electric dipole is to modify the field intensity by exp $(-h/\delta)$, where h is the depth of burial and δ is the skin depth of the medium. As h approaches zero, the expressions for radiation fields are identical with those developed by Norton.

An Investigation of Near-Field Radiation Patterns Measured with Large Antennas—E. V. Jull (p. 363)

The plane-wave spectrum concept is used to derive an expression for the radiation pattern of an antenna in a two-dimensional system and an investigation is made of the effect of the aperture size of the measuring antenna on the pattern obtained. It is shown that measured sidelobe levels either above or below the far-field values may be obtained. Measurements of near-field antenna radiation patterns are carried out in a parallel-plate pattern range, and the predicted effects of measuring antennas of finite size on the beamwidth and sidelobe levels of the measured patterns are confirmed experimentally.

Space Tapering of Linear and Planar Arrays—Robert E. Willey (p. 369)

The recent activity in electronically scanned arrays has stimulated interest in means of reducing both the quantities of elements required for a given size aperture and the number of different types of transmitters which would be necessary in an array using an illumination taper. One method of doing this is presented in this paper. The conventional amplitude tapered array is simulated by varying the spacing of equally excited elements, hence the name "space tapering." Space tapered arrays with predictable gains, beamwidths, and sidelobe levels can readily be designed using graphical techniques and simple mathematics. Reduction in the number of elements of from 50 to 90 per cent for moderate and large size planar arrays are possible while retaining good pattern characteristics. Although only linear and circular planar arrays are discussed, the technique can be applied to other planar and three dimensional arrays.

A Quasi-"Isotropic" Antenna in the Microwave Spectrum—Dimitri S. Bugnolo (p. 377)

The general problem of radiation by apertures on spherical surfaces was first considered theoretically by Stratton and Chu in 1941. More recently, the problem was again considered by Bailin and Silver and Mushiake and Webster.

All numerical and experimental efforts to date have been restricted to the special case of a very small sphere as measured in wavelengths. This paper is in effect an extension of the previous results to the case of large spheres uniformly excited by an equatorial slot of varying size. The detailed numerical and experimental results are restricted to the first TM and TE modes.

In general, it is possible to show that the large spherical slot antenna can be made unusually "isotropic" with a far-field coverage factor of over 98 per cent.

The Dipole Antenna Immersed in a Homogeneous Conducting Medium—K. Iizuka and R. W. P. King (p. 384)

It is the object of this investigation to measure the driving-point admittance, the amplitude distribution of the current, and the phase distribution of the current, relative to the phase at the driving-point for a dipole antenna immersed in a homogeneous conducting medium.

Measurements of the driving-point admittance have been made for a range of values of the ratio $\sigma/\omega\epsilon_r\epsilon_0$ of the medium taking the electrical height of the antenna βh as a parameter. The ratio $\sigma/\omega\epsilon_r\epsilon_0$ is varied from $\sigma/\omega\epsilon_r\epsilon_0=0.036$ to $\sigma/\omega\epsilon_r\epsilon_0=8.8$ and the antenna height βh is varied from $\beta h=0.1$ to 2π at intervals of 0.1.

The amplitude and phase distributions of the current have been measured for $\beta h=\pi/2$, $3\pi/4$, π, and $5/4\pi$ in a homogeneous conducting medium whose conductivity is varied from $\sigma/\omega\epsilon_r\epsilon_0=0.036$ to $\sigma/\omega\epsilon_r\epsilon_0=8.8$.

An Experimental Study of the Half-Wave Dipole Antenna Immersed in a Stratified Conducting Medium—K. Iizuka and R. W. P. King (p. 393)

Measurements are described of the driving-point admittance and the amplitude and phase distributions of the current for a half-wave dipole antenna immersed in a stratified medium for various gradients S of the ratio $\sigma/\omega\epsilon_r\epsilon_0=b+sx$, where the origin of the coordinates is at the driving point of the dipole antenna and the x axis is along the antenna.

The study includes a description of the apparatus, a discussion of the effect of the nylon films separating the layers, and an interpretation of the data. Reference is made to the possible application of a half-wave dipole antenna to the detection of an inhomogeneity in a medium.

Properties of Focused Apertures in the Fresnel Region—John W. Sherman, III (p. 399)

The diffraction field of continuous, rectangular apertures is analyzed when the system is focused in the Fresnel region. The Fresnel region is defined by phase and amplitude considerations, and the boundary separating the near and Fresnel regions is given as well as the conventional boundary distinguishing between the Fresnel region and far field. Curves have been plotted for several square apertures of varying size for both uniform amplitude illumination and tapered illumination while the focused condition is maintained. A theorem is given proving that the far-field pattern occurs in the focal plane of a continuous, focused aperture. Also included is the expression for gain of a focused rectangular antenna, and a discussion of the concept of gain where strict $1/R^2$ dependence does not exist. A comparison of focused circular and rectangular apertures is made in regard to beamwidth, energy between 3-db points, and energy in the main beam. Consideration is also given to the problem of maximizing the field at the point of focus, and as a corollary the "depth of focus" is examined.

A Wire-Grid Lens Antenna of Wide Application Part I: The Wire-Grid Lens-Concept and Experimental Confirmation—R. L. Tanner and M. G. Andreasen (p. 408)

A novel broad-band lens antenna consisting of a pair of circularly symmetric wire grids with varying grid spacing is described. The antenna is useful for beam scanning at lower frequencies such as at HF where dielectric and waveguide-type lenses are impractical. It should also prove useful in HF communications applications. A single lens structure with multiple feeds could serve as a high gain, high resolution antenna on many different circuits simultaneously. In addition, it could become useful as a printed-circuit lens antenna at microwave frequencies. The property of a pair of grids that makes it suited for lens structures is that the velocity of a wave propagating between two grids can be varied by varying the ratio between the grid spacing and the mesh size.

The measured performance of an experimental model of the HF wire-grid lens antenna has demonstrated the feasibility of the basic idea underlying the antenna. The model was tested in a 2:1 bandwidth and a sidelobe level of lower than -20 db was measured. In principle, the bandwidth of the lens can be made as large as desired.

A Wire-Grid Lens Antenna of Wide Application Part II: Wave Propagating Properties of a Pair of Wire Grids with Square, Hexagonal or Triangular Mesh—M. G. Andreasen and R. L. Tanner (p. 416)

A study of the wave-propagating properties of a pair of wire grids with square, hexagonal or triangular mesh is presented. The results of the

investigation contain the information needed for designing a wire-grid lens antenna. Of the three mesh types investigated, the hexagonal mesh is found to exhibit the least amount of dispersion and anisotropy for a constant mesh area.

Pattern Limitations in Multiple-Beam Antennas—W. D. White (p. 430)

It is shown that in a lossless passive antenna radiating multiple beams from a common aperture, the radiation pattern and the crossover levels cannot be specified independently. This is due to a requirement that the beams must be orthogonal in space. It is also shown that this limitation can be overcome by using multiple apertures, lossy networks, or active networks. Suitable arrangements for obtaining low sidelobes with satisfactory crossover levels are described.

Excitation of Slots in a Conducting Screen Above a Lossy Dielectric Half Space—Janis Galejs (p. 436)

A slot in a conducting screen which separates a lossless from a lossy dielectric half space is excited by a current source which lies in the interface between the two media. This geometry idealizes slot antennas constructed by placing wire mesh on the ground surface or island antennas as proposed by Morgan and Gould.

For narrow slots of width $w<\delta$ (δ = skin-depth in the lossy medium) the transverse distribution of the electric field in the slot plane is approximated by the static solution. The longitudinal distribution of the potential across the slot is obtained from a Hallén type solution of an integral equation. For slots of negligible width the propagation constant along the slot approaches the propagation constant of a thin wire which lies in the interface between the two dielectrics.

For wide slots of $w\gg\delta$ the integral equation can be solved for the longitudinal variation of the transverse electric field by Fourier transforms, provided that the variation of the electric field in the transverse direction is negligible. The electric field exhibits approximately the same longitudinal variation in the slot plane as the electric field in a direction transverse to the axis of an infinitely long line current which lies in the interface of the two dielectrics.

The radiation efficiency of the wide slot ($w>\delta$) is larger than the radiation efficiency of narrow slots ($w<\delta$). Still, the expected efficiency of the wide slots is less than 10^{-3} for frequencies in the VLF range and below, if the ground conductivity is of the order of 10^{-3} mho per meter. This indicates an upper limit to the efficiency of an island antenna as proposed by Gould and Morgan.

A General Analysis of Geodesic Luneberg Lenses—R. C. Rudduck and C. H. Walter (p. 444)

Geodesic lenses have been developed which give a collimated beam diametrically opposite the focal point and at an arbitrary angle relative to the plane of the lens rim. The lenses may have an arbitrary variation of the index of refraction; however, only the constant index lenses are considered in detail.

Experimental results are in good agreement with the theory. The practical importance of these lenses is that as a result of their circular symmetry a lens antenna may scan a full 360° by feed movement alone or simultaneous coverage in many directions may be accomplished by use of multiple feeds.

Universal Curves for the Horizontal Polarization Reflection Coefficient—G. P. Ohman (p. 450)

A set of universal curves is presented by which the approximate magnitude and phase of the reflection coefficient for horizontal polarization can be determined at any grazing angle with relatively little computation. The approximation is excellent when the relative impedance ratio of the two mediums is high, and sufficiently accurate for almost all engineering purposes even when the impedance ratio is only moderately greater than unity. This set of curves is intended to be a companion to a similar set described in a previous article for the vertical polarization reflection coefficient.

Wave Propagation in Magneto-Plasma Slabs—H. Hodara and G. I. Cohn (p. 452)

The propagation characteristics of electromagnetic waves in magneto-plasma slabs are of interest in predicting the behavior of trapped modes in anisotropic waveguide structures such as the ion ducts formed by the lines of the earth's magnetic field. This paper derives the expression for waveguide modes present in a uniform infinite plasma slab. The slab has finite thickness, is filled with a lossless plasma and threaded by a uniform static magnetic field along the direction of propagation. It is shown that two sets of hybrid modes can propagate in a magneto-plasma slab at frequencies below plasma resonance, but no trapped models are supported above this frequency. For infinite magnetic fields, the modes uncouple and only a TM mode is supported.

The Magneto-Ionic Theory for Drifting Plasma—Hillel Unz (p. 459)

The magneto-ionic theory for the case of drifting plasma with constant drift velocity $\bar{v}_0$ is developed. The refractive index n is shown to be the solution of an algebraic equation of the sixth order. In several particular cases the algebraic equation is solved explicitly.

A Note on Radiation in a Gyro-Electric-Magnetic Medium—An Extension of Bunkin's Calculation—Yutze Chow (p. 464)

Some fundamental results on radiation in anisotropic media have been worked out by F. V. Bunkin for the case of a *gyro-electric (or plasma) medium* using the technique of tensorial Green's function. He obtained asymptotic expressions for the far-zone radiation fields in a gyro-electric medium and therefore demonstrated actually the validity of Sommerfeld's radiation condition for a gyro-electric medium. The present paper extends Bunkin's calculation to a case of a *gyro-electric-magnetic medium* and thus justifies the validity of Sommerfeld's radiation condition in a *gyro-electric-magnetic medium*.

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Audio

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The Editor's Corner—Marvin Camras (p. 59)
PGA News—(p. 60)

Absolute Measurements of Magnetic Surface Induction—Frank A. Comerci (p. 64)

Two methods have been used for obtaining an absolute measurement of normal surface induction of a medium wavelength signal recorded on magnetic tape. One employs a nonmagnetic loop and an electrical procedure for determining its active rectangular cross-sectional dimensions. The normal surface induction is computed from these dimensions and the output voltage induced in the loop as the tape is transported at a prescribed speed across the active portion of the loop. The other is based on recording a dc signal of current level equal to the rms value of the ac recording current which produces the equivalent recorded signal as that for which the measurement is desired. A calibrated search coil is then used to measure the total remanent flux in a bundle of many layers of the dc recorded tape. The surface induction is then calculated for medium wavelength signals using existing formulas.

Measurements of surface induction obtained using the two methods provide equivalent results. The surface induction could be measured for a wide range of wavelengths using the nonmagnetic loop.

A Study of the Chemical and Physical Properties of Magnetic Recording Tape—F. Nesh and R. F. Brown, Jr. (p. 70)

A study has been made of the chemical and physical properties of magnetic recording tape. The chemical studies indicate the possible relationship between a series of solvent tests and the wear properties of commercially available tapes. The physical studies, including electron micrographs, indicate a smaller unit crystal size for γ-ferric oxide than previously supposed and a relationship between a better dispersion and improved magnetic properties. Both studies are being continued in greater detail.

New Approaches to AC-Biased Magnetic Recording—D. F. Eldridge and E. D. Daniel (p. 72)

Ac-biased magnetic recording is described in a brief analysis based upon the anhysteretic magnetization process. Several different models are used to illustrate particular aspects of the recording process. It is shown that, using conventional record heads, there is bound to be a conflict between the efficient recording of short and long wavelength signals. The critical requirements involved in recording 15 kc at as low a tape speed as $1\frac{7}{8}$ ips are emphasized by pointing out that, on reproduction, 75 per cent of the output comes from the first 0.7 micron (28 microinches) of the coating.

Various methods for improving the record process have been proposed from time to time. These methods consist of proposals for sharpening the recording field as a whole, sharpening the bias field alone, and increasing the uniformity of the recording field through the coating thickness. The design of heads to accomplish any of these objectives remains to be worked out in detail. However, several possible techniques are discussed, namely pole-shaping, eddy-current shielding and the use of poles on both sides of the tape.

An entirely new recording technique is described which utilizes successive recordings by two separate gaps of different sizes. The two gaps can be contained in separate heads or in a single head, provided that the time-delay between the two recordings, if significant, is compensated for prior to recording. The principle is simple. The long gap is fed only with low-frequency signals which it can record effectively throughout the whole coating; the short gap records the high-frequency signals in the outermost layer of tape. The bias field from the short gap need not significantly reduce the level of the low-frequency recorded information. Filtering requirements and methods of time-delay compensation are discussed in the text, and a mathematical analysis is given of the interference to be expected from an uncompensated system.

Fabrication of a Magnetic-Tape Transducer by Electroplating and Engraving—C. J. Peters (p. 79)

A new technique for the construction of magnetic-tape heads is described. In this technique the magnetic material in the vicinity of the air gap is deposited by electroplating and the air gap is formed by engraving. Analysis of the leakage flux and of the pole piece saturation is presented. The performance of the head is evaluated in terms of frequency response.

Some Experiments with Magnetic Playback Using Hall-Effect Sensitive Elements—Marvin Camras (p. 84)

Flux sensitive heads using thin semiconductors have been developed for playback of magnetic recordings. The elements have a frequency response inherently flat from dc to tens of megacycles, so that the main limitations are in the associated head structure. Factors entering into optimum design are discussed, including materials and configurations of semiconductor elements and cores. Experimental results for successful designs are given, and a number of applications are suggested.

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Automatic Control

Vol. AC-7, No. 4, July, 1962

Image or Mirage?—Editorial—(p. 1)
The Issue in Brief—(p. 2)
On the Real-Time Control of Time-Varying Linear Systems—J. S. Meditch and J. E. Gibson (p. 3)

A new class of sampled data controls for time-varying dynamic processes is developed and investigated.

The notions of interval control, prediction, and synthesis of the optimum control variable by a linear combination of orthonormal polynomials in time t are introduced and applied in the optimization of a modified least-squares integral index of performance. Minimization of the index of performance leads to a family of control laws which specify the controller synthesis. The resulting control configuration is optimum on a "per interval" basis and is readily realized with available physical components. The design circumvents the complex computational problems associated with the usual calculus of variations and dynamic programming approaches. All controller operations are performed in real time using analog components, and sampling and reset circuits.

Controller design is discussed in terms of the available parameters: the relative weighting of system error and control effort, the control interval length, and the degree of the polynomial sum approximation of the control variable. A method of obtaining an engineering estimate of the latter quantity is developed and illustrated by an example.

An analog simulation of the control of a second-order dynamic process whose parameter varies in such a manner that the process is unstable at one extremum and heavily damped at the other is presented. Experiments are performed to measure the quality of steady-state control and the transient response of the control scheme.

Adaptive Finite Time Filtering—R. S. Bucy and J. W. Follin, Jr. (p. 10)

A detailed analysis of a particular adaptive filter has been carried out and the required extension of the theory to the general case is indicated. The filter measures the spectral densities of the input signal and noise processes and adjusts its band-pass to give optimal filtering in the Wiener sense. The behavior is examined in the linear approximation and a crude treatment of the nonlinear transient response is given. These results compare favorably with an analog simulation.

Theory of a Two-Parameter Adaptive Control System—V. Rajaraman (p. 20)

A two-parameter adaptive control system employing multiple models and model-parameter perturbation is proposed in this paper. Theoretical studies of this system have shown that it is self-adaptive when both the input signal and the parameters of the system vary unpredictably. It is shown that the two parameters of the feedback system are adjusted automatically along the path of steepest descent of the performance measure-parameters surface. An interesting feature of this system is that the perturbation signals do not appear at the output of the controlled system.

Synthesis of an Optimal Set of Radar Track-While-Scan Smoothing Equations—T. R. Benedict and G. W. Bordner (p. 27)

Performance measures are described which realistically reflect both noise-reduction and maneuver-following capability of a radar track-while-scan (TWS) system. Illustrations are given which compare various smoothing equations on the basis of this performance measure. Finally, a set of position-and-velocity tracking equations is synthesized by a calculus-of-variations technique.

The synthesized set is optimum for both position and velocity tracking within the given performance sense, in the class of all fixed parameter, linear tracking equations. The resulting optimally synthesized set characterizes the commonly termed "α-β" tracker, with the important proviso that

$$\beta = \alpha^2/(2-\alpha).$$

On Timesharing of Control System Components—J. Peschon and W. H. Horton (p. 33)

The present paper shows that the concept of component timesharing is applicable to certain types of control systems. A model in which a group of components is commutated sequentially around several control loops is presented. Two mathematical procedures, borrowed from known sampled data and state-space techniques, permit the detailed analysis of the resulting linear system. It is seen that timesharing creates cross coupling or interaction between control loops which are independent except for the shared component. There is no cross coupling in the steady state if each loop contains at least one integrator. It is concluded that timesharing is an economic proposition in which possible savings in equipment cost are balanced out against reduced system performance.

Discussion—D. P. Lindorff and J. Peschon (p. 38)
Closed Loop Response of Nonlinear Systems by a Functional Transformation Approach—J. C. Hill (p. 39)

A graphical technique is presented for determining the closed loop response of nonlinear control systems driven by sinusoidal inputs. The nonlinear portion of the system is represented by its conventional describing function, which may be frequency dependent as well as amplitude dependent. The linear portion of the system is represented by its complex frequency response function $G(j\omega)$. A transparent overlay is used to mechanize a functional transformation similar to that performed by a Nichols chart, allowing rapid determination of the system output for sinusoidal inputs. The accuracy of the method is limited by the accuracy of the describing function approximation.

In addition to offering a rapid solution to what has been regarded as a time consuming problem, the method gives the designer sufficient insight into the behavior of the system to allow the intelligent choice of compensating networks to improve system performance.

A numerical example is used as a vehicle for discussion of compensation, and experimental results are presented to verify the analysis.

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Communications Systems

Vol. CS-10, No. 1, March, 1962

Toward a Communication System Discipline—(p. 1)
Predicting the Performance of Tropospheric Communication Links Singly and in Tandem—A. P. Barsis, K. A. Norton and P. L. Rice (p. 2)

Performance of long-distance tropospheric communication circuits, either singly or in tandem, is predicted in terms of the probability of obtaining a specified grade of service or better for various percentages of time. The grade of service is determined by the minimum acceptable ratio of hourly median predetection rms-signal to rms-noise for the type of intelligence to be transmitted.

The standard deviation of prediction errors depends upon the percentage of hours for which the specified grade of service is required and on parameters characterizing the propagation path. The possibility of reducing this standard deviation by making path-loss measurements is dissussed. It is shown that no improvement is possible unless the test path is very nearly the same as the proposed operational path; in particular, unless the test path and operational paths have terminals less than one mile apart it is shown to be doubtful in most cases whether the observations will be useful in improving the reliability of the predicted performance of the proposed system. Assuming that the test path and proposed operational paths are identical, estimates are given of the number of days of observations required to halve the prediction uncertainty or, alternatively, to reduce it to 3 db; in some cases several years of observations are required.

On Optimal Diversity Reception, II—G. L. Turin (p. 22)

In a previous paper the structure of the optimal multi-diversity receiver for a fading, noisy diversity channel was determined under the assumption that the memory of the receiver does not extend beyond the time interval of the signal currently being received. The probability of error for such a receiver is determined herein under the following assumptions: the channel is binary and symmetric; the (Gaussian) noises in the various diversity links are white and independent; and the fadings in the various links are Rayleigh distributed and slow, but not necessarily independent. Detailed curves of this probability of error are given for various orders of diversity in the two cases of independent fadings and "exponentially correlated" fadings.

Since the optimal receiver may be difficult to implement, a more easily implemented "square-law combining" receiver is also considered, and it is shown that for all practical purposes this simpler receiver behaves optimally. Finally the effect of the assumption that the receiver has a short memory is considered by comparing its performance with that of a longer-memory receiver studied by Pierce and Stein.

Prediction Diversity Combining with Selectively Fading Channels—P. Bello and B. D. Nelin (p. 32)

The effective use of predetection (or IF) linear combining of diversity channels requires that all frequency components be added in = phase. In view of the time-variant phase characteristics of fading channels, some form of instantaneous phase correction must be provided prior to the diversity combining so that proper in-phase addition will result. If the instantaneous fading is not uniform over the width of the channel, the resulting channel incoherence will reduce the efficacy of the phase correction. One way to avoid this difficulty is to divide the selectively fading channel into "flatly" fading subchannels and perform the necessary phasing and combining operations per subchannel. The question arises as to how much frequency selective fading can be tolerated within a channel, or a subchannel, before serious deterioration in performance results. This paper considers this question with reference to FDM-SSB binary data transmission and matched filter reception, wherein a pilot tone is used both for deriving proper phase correction and for providing the weighting necessary for maximal ratio combining. Analytical expressions are derived for error probabilities which show the deleterious effects of frequency selectivity and pilot tone additive noise.

Effect of Propagation Fading and Antenna Fluctuations on Communication Systems in a Jamming Environment—A. B. Glenn and G. Lieberman (p. 43)

The effect of different degrees of both slow propagation fading and antenna fluctuations on the performance of communication systems in the presence of interference was analyzed. The performance of the system was evaluated

from the aspect of the probability of error which in turn is related to the required system margin. The results are plotted for a large variety of fading conditions for both the desired and undesired signals and antenna fluctuations. Some typical results are given below:

Propagation Fading with No Antenna Fluctuations

1) For greater than 85 percentage satisfactory operation, the fading of the additive interfering signal has negligible effect on the probability of error when the desired signal experiences "strong" or Rayleigh distributed type of fading. The system margin for 99 per cent of satisfactory operation when the desired signal is Rayleigh fading and whether or not the undesired signal is fading is 20 db.
2) When the fading in the desired path is normally distributed, the effect of a normal fading interference signal is to increase the system margin compared to a nonfading interfering signal by 1 db and 1.4 db for 90 and 99 per cent satisfactory operation, respectively. The system margin for 99 per cent satisfactory operation is 5.2 db and 6.6 db when the undesired signal is not fading and is normally fading, respectively.
3) When there is no fading in the desired signal path and the undesired signal is fading, the system margin for 99 per cent satisfactory operation is 5 db for normal fading and 6.8 db for Rayleigh fading.
4) The system margin for 99 per cent of satisfactory operation when the desired signal is normal fading and the undesired signal is Rayleigh fading is 15 db. The system margin must be increased to 20 db when the desired signal is Rayleigh fading and the undesired signal is normal fading.

Propagation Fading and Antenna Fluctuations

1) For a good UHF airborne antenna, the approximate variance of 0.1 is calculated from the vertical and horizontal patterns. The system margin for 99 per cent satisfactory operation is 6.4 db for no propagation fading in the desired or undesired paths. When there is normal fading in both the transmission paths, the system margin for 99 per cent satisfactory operation increases to 8 db or only an increase of 1.6 db over the case for no propagation fading.
2) Increasing the antenna variance to 0.3 causes an increase in the system margin for 99 per cent satisfactory operation and normally distributed fading in both paths to 9.6 db or an increase of 1.6 db due to the increased antenna variance.
3) The percentage of satisfactory operation as a function of system margin for antenna variances equal to 0 and 0.1 for a Rayleigh fading desired signal and a normally fading undesired signal shows that above a 10-db system margin, the curves are indistinguishable. A 20-db system margin is needed for 99 per cent of satisfactory operation.
4) The effect of the antenna variance is very significant when the desired signal is normal fading and the undesired signal is Rayleigh fading. For 99 per cent of satisfactory operation, the system margin must be increased from 15 db to 26 db when the antenna variance is increased from 0 to 0.10.

Multiple Echo Distortion in Frequency-Modulation, Frequency-Division-Multiplex Trunk Radio Systems—R. G. Medhurst and J. H. Roberts (p. 61)

This paper is concerned with intermodulation distortion in FM, FDM trunk radio systems due to a number of echoes of differing amplitudes and delay times. General formulas are developed, and applied to the case of a 960-channel system subjected to distortion due to echoes from flanges in waveguide runs. Examples considered are 100- and 200-ft runs of RG-50/U waveguide, in 10-ft sections, for use in the 6000-Mc band. The calculations are carried out with the simplifying assumption (for which there is some experimental evidence), that the individual flange reflections are equal and of the same sign. Distortion levels are computed both with equal section lengths and with section lengths made slightly unequal, so as to destroy coherent addition of the individual echoes. While the latter arrangement gives a considerable reduction, as compared with the equal length arrangement, in the peak input standing-wave ratio, the improvement in the distortion is so small as probably not to justify, in practice, any deliberate variation of the lengths.

Performance of Error-Correcting Codes—Michael E. Mitchell (p. 72)

Binary decoding of relatively short codes corrupted by Gaussian noise or pulse interference has been considered in applications for which the minimum useful information block length is unity, equal to the information bit capacity of a code, or equal to an integral multiple of the information capacity of a code. The following results pertain to applications which are further characterized by the performance criterion of minimizing error rate and transmitter power, with only minor restrictions on transmission bandwidth and none on channel efficiency (no credit given for error detection):

1) The word error probability for any group error-correcting code is never exceeded by its corresponding bit error probability (for either Gaussian noise or pulse interference).
2) Relatively short codes (10 to 50 information bits) of moderate redundancy (40 to 50 per cent) are available which can reduce bit error rates by factors of 10 to 400, or equivalently, permit 1 to 3 db reductions in transmitter power, when the coded system bit error rate is 10^{-6} (Gaussian noise).
3) These same codes can alternatively reduce word error rates by factors of 10 to 10^4, or equivalently, permit reductions of 1.5 to 3.7 db in transmitter power, for average word error rates corresponding to values of the quantity E/N_0 (signal energy per information bit/Gaussian noise power density) in the neighborhood of 8 db for the coded systems.
4) Under worst-case pulse interference, these codes permit 4 to 7 db reductions in transmitter power for average word error rates less than 20 per cent. (Considerably greater improvements result for the pulse interference expected.)
5) All of these codes are easily encoded and decoded.

Error Rates in Coherent Communication Systems—A. V. Balakrishnan and J. E. Taber (p. 86)

In coherent PCM communication systems where the noise is additive and Gaussian (such as in space communication systems) two classes of nonredundant coding are of special interest. One is Regular Simplex Coding which yields the lowest possible error rate for fixed SNR. The other is Bi-orthogonal Coding because of its many desirable features from the practical point of view. In this paper the error rates per word and per bit are computed for both of these classes. The communication efficiency—a quantity of major importance in space communications—is also calculated for fixed error rate as a function of code size.

A Coherent Digital Amplitude and Phase Modulation Scheme—C. N. Campopiano and B. G. Glazer (p. 90)

In this paper, we present a technique for a combination of amplitude and phase modulation for digital communication. This technique assumes phase coherence between transmitter and receiver. For purposes of analysis, we also assume baud synchronization. The technique involves the simultaneous and independent amplitude modulation of two carriers of the same frequency which are in quadrature with each other. The demodulation process consists of the separation of the in-phase and quadrature components followed by amplitude detection. This scheme appears to be easier to implement than either the system studied by C. R. Cahn or the system investigated by J. C. Hancock and R. W. Lucky. Furthermore this system is comparable in reliability to the latter scheme.

A Comparison of Several Types of Modulation—C. S. Weaver (p. 96)

This report presents a brief mathematical analysis of various modulation-demodulation combinations which may be used in data-link systems. Both digital and analog transmission systems are discussed.

For some of the more important digital (binary) systems, the probability of making an error in detecting a binary symbol (using the solutions of Cahn, and Berry and Garbedian) is used to compare the various systems. It is shown that there is no one binary information transmitting system that is superior under all circumstances. Analog systems are discussed in terms of mean-square error (variance), and this criterion provides a convenient method for comparing digital and analog systems.

There is a discussion of transmitter and receiver bandwidth—an important consideration for spectrum utilization.

A time multiplex PAM system is discussed, along with crosstalk and noise level considerations for such a system.

Most of the methods presented here are not new; however, it is hoped that the compilation and direct comparisons of the various modulation systems will be useful and timely for the systems engineer.

Asymmetrical Sideband Transmission of Position Modulated Pulses—Georges L. Lukatela (p. 102)

The present analysis is a theoretical examination of the asymmetrical sideband transmission in PPM systems. It is shown that the synchronously detected component of the envelope of an asymmetrically transmitted impulse can possess a steeper slope than the envelope itself and that over-all transient time of this component may be reduced.

The effect of various degrees of transmission asymmetry is calculated for "squared cosine" type transmission-frequency characteristic. The conclusion is reached that for a given channel capacity an asymmetrical sideband transmission in PPM systems should be distinctly superior to a symmetrical one.

A Synchronized Pulse Communication System with Pseudorandom Interpulse Period—Richard C. Mackey (p. 109)

The security of a pulse type communication system can be improved by the incorporation of an element of randomness in the time of transmission. A system is simulated in which a remote receiver is caused to be gated on at apparently random intervals which are in synchronism with the arrival of the signal from the controlling local transmitter. The time of transmission is not random, but is determined by an extremely complex program. By duplicating the program in the remote equipment, the receiver can anticipate the time of arrival of the local transmissions.

The time of transmission is varied ± 10 per cent about a nominal period which corresponds to a message repetition frequency of 100 cps. The pseudorandom inter message variations are governed by the output of an eighteen stage magnetic core shift register. The period of repetition of the pseudorandom transmission is in

excess of six minutes and much longer periods are readily obtainable. Special scrambling techniques are used to increase the burden of a would-be code breaker.

The system can be operated in either the periodic or pseudorandom mode, and a fail-safe feature is provided to return the system to the periodic mode if the local and remote units become unsynchronized while in the pseudo-random mode. Receiver synchronization is maintained by a voltage controlled clock frequency oscillator which is phase locked to the local transmission. The phase lock compensates for Doppler shift in the received frequency and for any drift inherent in the remote clock source.

A New Pulse Modulated Speech Communication System Having a Minimum Duty Cycle—R. O. Eastman (p. 114)

The development of the Navy Electronics Laboratory of a low duty cycle, multichannel pulse modulation communication system is described. The system provides adequate naturalness of speech while maintaining a quantized sampling rate which is proportional to the information bandwidth.

First, a general description of the speech processing and reprocessing techniques is given. This includes techniques for generating infinitely clipped signals, and for deriving an optimum sampling rate from these signals. Next is shown how two separate forms of information, amplitude and frequency, are conveyed by a time modulated pulse. Finally is presented the results of tests conducted through the system.

Digital Computer Simulation of Communications Networks—R. L. Dujack and D. I. Epstein (p. 118)

In planning large scale communications networks, the systems engineer must define the best mix of system trunking assignments, switching logic, and system operating policies. Field tests are not usually a practical method of evaluating network performance as the above design parameters are changed—particularly if the networks are complex and if many operating personnel are involved. An alternative to field testing is offered by the use of a model which can duplicate with reasonable fidelity the performance characteristics of the real system. Digital computer simulations are a practical implementation of the modeling approach. If the input-output routines of the computer program are skillfully planned, the computer model can be used to evaluate a broad spectrum of system configurations and operating doctrines under arbitrarily selected changes in the system environment.

The construction of the appropriate model must be performed with care. If too many facets of the real life communication process are ignored, the simulation results are meaningless. If the model is too complex, excessively long computer running times might ensue. When the simulation technique is intelligently matched to the problem under consideration, results of considerable practical value are obtainable.

In this paper a digital computer simulation technique is discussed and some examples of its application are presented.

Overlapping Tessellated Communications Networks—L. J. Craig and I. S. Reed (p. 125)

In this paper we present a theory of interconnecting communication networks which have the appearance of overlapping tessellations. We have derived the bandwidth requirements for two types of networks—the linear spanning and the inclusive spanning. The efficacy of one network as compared to the other is measured by the additional switching requirements for various spanning doctrines.

Grade of Service of Direct Traffic Mixed with Store-and-Forward Traffic—Joseph Otterman (p. 130)

The dual use of trunks for both direct and store-and-forward (S/F) traffic makes high trunk efficiency possible. The resultant trunk savings are important in communication systems in which long-haul trunks contribute heavily to the cost of the system. This paper reports work on the computation of trunking tables that could be used to engineer trunk requirements for prescribed loads and distributions of direct and S/F traffic.

A method of computation and some specific results in terms of grade of service of direct traffic and traffic capacity for S/F traffic are presented. The numerical results are for two to forty-eight trunks. The results apply to the case of exponentially distributed holding times of both the direct and the S/F traffic.

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Electronic Computers

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Iterative Switching Networks Composed of Combinational Cells—William Kilmer (p. 123)

The networks considered in this paper consist of n identical combinational logic cells connected in cascade through bi-directional discrete information channels. All switching is done synchronously with unit time delay through each cell. Three classes of networks are formed according to whether or not information flow in one direction along the cascade is dependent upon that in the other. Steady-state behavioral relationships between the three classes are discussed, and the class having mutually dependent information flow is shown to be the only one which can exhibit steady-state memory properties. The main object of the paper is to derive several theorems concerning these memory properties. The last section interprets some previous results of Hennie as a start on the transients and cycling problems.

Examples of Abstract Machines—Seymour Ginsburg (p. 132)

Numerous physical situations related to data processing are shown to be modeled by a mathematical entity called a quasi-machine. The situations described include 1) single inputs producing multiple outputs, 2) machines yielding no outputs upon insertion of certain inputs, 3) the retention of the last n outputs only, 4) "erase left" on tape, 5) different input routines doing the same work, and 6) certain types of asynchronous switching circuits. The first five may be modeled by quasi-machines with a special property, such quasi-machines being called abstract machines.

Cascaded Switching Networks of Two-Input Flexible Cells—K. K. Maitra (p. 136)

This paper studies a class of switching networks which are constructed by cascading two-input, single-output completely flexible binary logical cells. It is assumed that each cell can be adjusted at will in order that any one of the sixteen possible switching functions of its two binary inputs may be obtained at its single output lead. An n-input cascaded network is composed of $n-1$, two-input, single output logical cells. The inputs to a typical cell A_i consist of a signal variable X_i and the output a_i from the preceding cell A_{i-1}. For the end cell A_1 the inputs are the signal variables X_0 and X_1. The output is derived from the terminal cell A_{n-1}.

In view of the growing interest in the area of "self-organizing" or "adaptive" switching networks, the study of the type of network just discussed seems to be pertinent.

The results presented here are pertinent to typical n-input cascaded structures, and consist of characterization and enumeration of the class of all n-place realizable switching functions, test procedure for the determination of the realizability of arbitrary n-place switching functions, and finally determination of all possible cascaded structures for synthesis of realizable n-place switching functions.

The paper develops some new techniques especially suited to the present study. These techniques may also hopefully find applications in other types of logic synthesis.

The Vertex-Frame Method for Obtaining Minimal Proposition-Letter Formulas—Theodore M. Booth (p. 144)

The vertex frame is similar to an n-dimensional cube or Tychonoff frame that has been cut and unfolded into the shape of a Karnaugh map. The methods of use are similar to those for Karnaugh maps. Minimal disjunctive and conjunctive normal formulas are found for problems with or without don't-care cases. The selection graph, a linear graph, is used to enhance the prime antecedent (=prime implicant) selection procedure.

The vertex frame readily handles most problems of up to six variables. Problems with seven and eight variables have been worked successfully, but this is an area where more experience is needed in working out actual problems that arise in engineering practice.

As with any map method, pattern recognition plays an important role, and thus it takes longer to become proficient in this method than in some of the formula-manipulation methods (*e.g.*, Quine's, McCluskey's, Mott's).

The problem of recognizing plots, on a vertex frame, of symmetric and unate truth functions is discussed.

Logarithmic and Exponential Function Evaluation in a Variable Structure Digital Computer—D. Cantor, G. Estrin, and R. Turn (p. 155)

Sequential table look-up algorithms are proposed for the evaluation of ln x and exp x. Tables of pre-computed constants are utilized in transforming the argument into a range where the function may be approximated by a simpler polynomial. In the case of ln x, x is transformed so that it falls into a close neighborhood of 1; in the case of exp x, x is transformed to a close neighborhood of 0. These algorithms are particularly effective when mechanized so as to carry out the predetermined sequence of operations without waste manipulation. Such special purpose organization is reasonable within the variable structure part of UCLA's proposed Fixed-Plus-Variable structure computer, as the same equipment may be reorganized for use in other special purpose configurations at other times.

In this paper the sequential table look-up algorithms and their speed advantages over existing subroutine approximation procedures are described. The design of special purpose configurations which leads to another factor of speed increase as compared to a stored program implementation of the algorithms is presented. The properties of the supervisory control which integrate the operation of the "fixed structure" general purpose computer and the "variable structure" special purpose configurations are specified and the over-all computational gains evaluated. An order-of-magnitude increase in speed compared to existing subroutines is predicted in both cases.

A Computer for Solving Linear Simultaneous Equations Using the Residue Number System—Ronald M. Guffin (p. 164)

The design of a special-purpose digital computer for solving simultaneous equations which operates with numbers coded in the residue number system is described. Since addition, subtraction or multiplication can be done in one-bit time using this coding, Gauss-Seidel iteration can be done in a very fast and efficient manner. The computer has been arbitrarily designed to solve dense systems of equations with as many as 128 unknowns and sparse systems with as many as 512 unknowns. Operating at a 500-kc clock rate, the computer would be able to perform one complete iteration on a system with 128 unknowns 30 times faster than an IBM 704. Using a 7 digit residue code requiring a 42-bit word, the computer would provide solutions of up to 4 significant figures. By using the best presently obtainable components, computing speed can be increased by a factor of 5. The size of the system which

can be handled and the number of significant digits which can be obtained in the solutions can also be extended if desired.

The speed of computation obtained with this computer is made possible by the combination of the one-bit-time arithmetic operations obtainable with residue numbers, the high data rate possible with a magnetic drum, and the sequential nature of the Gauss-Seidel iteration procedure. The digital techniques which have been developed to realize a computer of this type include methods of encoding decimal numbers into residue representation, rescaling residue numbers, and decoding residue numbers into binary coded decimal form. A method of comparing two residue numbers which is suitable for this particular application is also discussed.

Computational Chains and the Simplification of Computer Programs—Thomas Marill (p. 173)

In the synthesis of switching circuits, a formal representation of the function to be realized by the circuit is first established and simplified as much as possible. Only then is construction of the circuit undertaken. It is argued that an analogous strategy should be followed in the synthesis of digital computer programs: the function to be realized by a program should first be established in a suitable formalism; the resulting formal expression should then be simplified as much as possible; only at this point should translation into the final "machine" program be undertaken.

In the light of this discussion, the simplification of a certain type of elementary program, containing no branching or internal modification, is considered in detail. It is argued that the analysis of this type of program, whose formalization is called a "computational chain," is a prerequisite to the analysis of more general programs. A system of notation is developed, and rules are given for minimizing the temporary storage requirements associated with a computational chain, for eliminating vacuous and redundant parts, and for forming combinations of chains.

Information Processing by Data Interrogation—J. Atkin and N. B. Marple (p. 181)

A data-processing technique is described in which a function is evaluated by rapid interrogation of the given data for the presence of combinations of data variables giving rise to values of the function which are of interest. Testing unlabeled measurements from several observers to find data sets which satisfy a test function is discussed as an example. For many problems the required time is much shorter than that needed when processing is carried out by a high speed general purpose machine. Logical design and equipment requirements for solution of such a problem by data interrogation are discussed.

The Stenowriter—A System for the Lexical Processing of Stenotypy—E. J. Galli (p. 187)

The system design and evaluation of an experimental system are presented. The "Stenowriter" utilizes a special-purpose translation computer, equipped with a large random-access memory, to provide real-time processing of spoken material into typewritten English. The input is provided by an operator using a special keyboard with which information may be stenographically encoded as rapidly as it is spoken. The stenographic code is automatically processed by the translation system into corresponding English with proper spelling and format. The output appears on an electric typewriter. Through multiplexing techniques, information for a large number of input-output units can be processed, each on a real-time basis, by one computer.

Time Average Thermal Properties of a Computer Utilizing Thin-Film Superconducting Elements—Harold Sobol (p. 200)

It is necessary to understand the static thermal properties of cryotron gates before one can predict the limitations associated with dissipative heating of a complex cryotronic computer. An experimental program has been conducted to determine the thermal properties of isolated and of thermally coupled gates. All experiments reported in this paper were performed on tin gates evaporated onto glass substrates.

The total thermal conductance K of a gate is defined, and experimental values of K are presented as a function of power and gate geometry. An analysis of the heat flow is given, based on a temperature-dependent coefficient of heat transfer. Theoretical values of K and temperature distribution are derived. The theory qualitatively predicts the temperature variation. Finally, the results are extrapolated to estimate the number of cryotrons that can be used safely in a thermally coupled system.

Tunnel-Diode Full Binary Adder—C. A. Benton and B. Rabinovici (p. 213)

A full binary adder utilizing a single tunnel diode and three resistors as its basic components is described, the circuit design criteria is outlined and values for a typical case are computed. The operating rate of the adder was determined analytically to be approximately 3 nsec. Experimental results obtained with a practical circuit that employed a 1N2939 tunnel diode were in good agreement with the analytical findings. The Sum and Carry outputs were 130 mv and 300 mv, respectively. These outputs need to be amplified if used to drive identical stages since there is no gain in this circuit. Since a full binary adder is a multipurpose logic element, a simple, compact, full binary adder that contains few components presents attractive possibilities for new computer logical schemes.

This adder offers circuit simplicity and stability and is particularly suitable for application where low-power dissipation and compactness are desired.

Circuits Employing Toroidal Magnetic Cores as Analogs of Multipath Cores—J. A. Baldwin, Jr. (p. 218)

Circuits which are analogous to multipath magnetic cores are described. Toroidal cores are employed to simulate the branches of the multipath core. The conservation of flux at a node is simulated by means of shorted windings connecting groups of cores. These insure that the sum of the flux changes in the cores on a given winding will vanish. Ampere's law is simulated by proper choice of both number of turns and core dimensions.

Equations are derived which determine the parameters of the analog in terms of those of the multipath core. These state that the volume of a core must be proportional to the volume of the corresponding branch, that the number of turns a shorted winding makes on a core should be proportional to the ratio of core diameter to branch length, and that the number of ampere-turns of externally supplied drive should also be proportional to this ratio.

Analog circuits employing single-turn coupling loops and cores of different diameters are discussed as a special case. Another special case employs multiturn windings but identical cores.

It is shown that the current flowing in a shorted winding is proportional to the magnetic potential of the corresponding node of the multipath core. The use of the core analog for the direct observation of internodal magnetic potential differences is illustrated by means of oscilloscope photographs of the magnetic potential difference between the nodes of a three-rung laddic.

One-Level Storage System—T. Kilburn, D. B. G. Edwards, M. J. Lanigan, and F. H. Sumner (p. 223)

After a brief survey of the basic Atlas machine, the paper describes an automatic system which in principle can be applied to any combination of two storage systems so that the combination can be regarded by the machine user as a single level. The actual system described relates to a fast core store-drum combination. The effect of the system on instruction times is illustrated, and the tape transfer system is also introduced since it fits basically in through the same hardware. The scheme incorporates a "learning" program, a technique which can be of greater importance in future computers.

Design of Memory Sense Amplifiers—G. H. Goldstick and E. F. Klein (p. 236)

A fundamental problem in the design of fast (4 to 6 μsec) word oriented memories using S-1 core material is the realization of a sensing amplifier which has the multitude of required properties. The sense amplifier is usually required to provide stable gain for bipolarity signals, to introduce a negligible delay into the signal path, to avoid dc level shift, to recover within a short time following the enable (or inhibit) noise, to provide a stable discrimination level, etc. Although many sense amplifier designs have appeared in the literature, none, to the authors' knowledge, are eminently suitable for a 4- to 6-μsec memory under the condition of "worst case input pattern"—a long sequence of unipolarity pulses interspaced with "blasts" of enable noise.

The basic requirements for a sense amplifier suitable for a word oriented memory in conjunction with a required memory cycle are discussed. This newly developed sense amplifier and its operation and design are presented. Emphasis is placed on the evolution of the techniques and circuits presented, and several other approaches pursued during the design are also described. The authors also comment on possible future approaches to the sensing problem.

A Harmonic Analysis of Saturation Recording in a Magnetic Medium—Bohdan Kostyshyn (p. 253)

A single write-read head coupled with saturation recording techniques is commonly used to store digital information. Because of the experimental difficulty in determining the contribution of individual parameters to the system, a theoretical equation describing the output of a system in terms of the physical, magnetic, and electrical parameters of the system has been developed and programmed for the IBM 704 Data Processing System.

When applied to an experimental NRZ system, excellent agreement between observed and calculated results was obtained. The phenomena of peak or phase shift and amplitude shift occurring for an isolated pair of adjacent ones at high bit densities is demonstrated and discussed. The calculated dependence of the zero-to-peak output and the peak-and-amplitude shifts is shown for variations in the thickness, retentivity, coercivity and squareness of the storage medium, and for variations in the write current, the head-to-medium spacing, and the pole gap dimensions.

The Magnetic Configuration of Stylus Recording—H. J. Kump (p. 263)

A cylindrical magnetic stylus driven by a coil can be used to transfer bits of information onto a magnetic surface. These bits are recorded as round spots and combinations of these spots can form a desired pattern, *e.g.*, numbers or letters.

A mathematical model corresponding to the magnetic properties of a recorded spot is proposed and analyzed. The major considerations in the analysis are the field pattern existing about the spot, the self-demagnetization of the spot, and interference resulting from crowding the spots. Field mappings, obtained with a miniature Hall probe, were used to experimentally verify the proposed model. The experimental results show that the analytic model can predict the magnetic configuration of the spot at distances away from the recording surface that are considerably greater than the surface thickness.

An alternate model is proposed for studying spot properties in regions quite close to the recording surface.

Stochastic Model for the Browning-Bledsoe Pattern Recognition Scheme—G. P. Steck (p. 274)

A stochastic model is presented which gives the probabilities of successful recognition of the Browning-Bledsoe recognition scheme as a function of scheme parameters and pattern variability parameters. Also, procedures are given for estimating the variability parameters from data so that the model can be used to predict readability. The adequacy of the model is checked by comparing estimated readability with observed readability for two sets of data, one with high variability and one with low variability.

The Browning-Bledsoe recognition scheme is also treated as a coding and decoding problem in which case the concepts of information theory are useful. Finally, brief mention is made of the connection between pattern recognition problems and classification problems in general, and the Browning-Bledsoe recognition scheme is compared and contrasted with other recognition schemes which make use of measurements on patterns.

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Engineering Management

Vol. EM-9, No. 2, June, 1962

About This Issue—The Editor (p. 45)

A Model of the Imitation Process—Edwin Mansfield (p. 46)

A simple model is developed to help explain differences among process innovations in the rate of imitation. The major hypothesis around which the model is built is the following: The probability that a firm will adopt a new technique is an increasing function of the proportion of firms already using it and the profitability of adopting it. The probability of adoption is also hypothesized to be a decreasing function of the size of the investment required. The model is tested with twelve innovations from three industries. Interindustry differences show up. An equation of the form predicted by the model can explain most of the variation among the rates of imitation.

The Creative Individual in Industrial Research—S. L. Jones and J. E. Arnold (p. 51)

A project was undertaken to analyze the processes associated with individual creativity in industrial research organizations and the effects of the company environment and management policies on the results of research and development (R and D) activities. Interviews were held with two hundred and nine technical researchers drawn from the fundamental, applied, and development staffs of seventeen private industrial firms, one research institute, one university laboratory, and the members of a creative engineering seminar for industry held at Stanford. One aspect of the study concerned the dichotomy of research specialist vs generalist. Most of the men participating in this study had broad technical and personal interests even though their current job assignments might each involve a definite specialization. Data on a wide range of characteristics and attitudes of research people are presented.

Selective Dissemination of Information—A New Approach to Effective Communication—C. B. Hensley, T. R. Savage, A. J. Sowarby, and A. Resnick (p. 55)

A first pilot test of a system to disseminate pertinent documents to individual persons selectively has been tried using IBM Research personnel and a variety of document sources. A detailed study of the user populations was made by means of diaries, interviews, and a mail survey. The effectiveness of the system is measured by 1) the percentage of unwanted or "trash" documents sent to the user, and 2) the percentage of documents the user wants but are not sent by the system. User responses to systematically and randomly selected documents, both integral parts of the system, were used to estimate these parameters.

The Use of Random Alarm Devices in Studying Scientists' Reading Behavior—Miles W. Martin, Jr. (p. 66)

An intensive study of the scientific journal reading behavior of chemists and physicists was made using almost exclusively the direct observation of behavior (rather than questionnaires, interviews, or diaries) to determine what they read, why they read what they do, and what differences in reading behavior exist between the groups. A device which generated an audible alarm at random moments of time was carried by scientists and used to signal the times at which observations were to be made.

Periodical Readership of Scientists and Engineers in Research and Development Laboratories—J. W. Hoyt (p. 71)

The reading habits of scientists and engineers engaged in research and development are shown to be readily differentiated by academic attainment and field of academic discipline. Lesser but important influences on readership appear to be pay level and general type of work (research, test, or development). From these data, advertising programs appealing to definite categories of research and development personnel can be shaped.

An important attitudinal question as to the usefulness of journal advertising as a source of employment information was clearly answered in the affirmative. Junior scientists and engineers seem to rely on employment advertising to a somewhat greater extent than those in more senior positions.

The Use of Technical Literature by Industrial Technologists—Christopher Scott (p. 76)

A study was made of the role of scientific information in the work of a sample of technologists in the British electrical and electronics industry. Interviews lasting up to one hour each were conducted with the people in the sample, using a standard set of questions. The sample consisted of 1082 people in 127 establishments. The term "technologist" covered all persons engaged in research, all those with technical qualifications, and all those responsible for planning development work. Among the findings are that the principal role of the literature to the people studied is to supply useful information which is not being deliberately sought by the reader. Its role as a references source is less significant.

A Manual Information Retrieval System—P. C. Constant, Jr. and P. C. Patton (p. 86)

A manual information retrieval system which is applicable to a library or bibliography of a few thousands of documents is described. It was designed for an engineering research program and has been used successfully on several relatively small research programs in the past few years. The system uses a Royal-McBee Keysort (KS-3718) card as the medium of information storage. The coding system is described in detail. The "keyword" concept is discussed and an example is given.

An Information Retrieval System for a Small Research Department—Leslie R. Axelrod (p. 92)

A modest information retrieval system is described for a small electronics laboratory. Initially it employs punched cards, but will be converted to edge-notched cards when the coding system is "debugged." The advantages and deficiencies of the system are discussed. The coding system was developed from key words and descriptive phrases in the laboratory's 700-document collection.

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Engineering Writing and Speech

Vol. EWS-5, No. 1, August, 1962

Editorial—(p. 1)

Problems Posed by an Expanding Technical Literature—Charles P. Bourne (p. 2)

A quantitative estimate is made of the magnitude of the world's electronics and electrical engineering journal literature problem. By the use of a number of basic sources of information, a composite picture is established to show total volume, linguistic and national origins, degree of coverage by the abstracting and indexing services, and the availability of special information centers. Some criticism is directed at the electrical engineering professional societies for not taking more positive action on this problem of concern to their entire membership. The paper concludes with suggestions for coping with this general problem.

The Construction of Statistical Tables—Christian K. Arnold (p. 9)

This paper outlines, in some detail, accepted rules for constructing statistical tables. Among the topics covered are the criteria for choosing vertical or horizontal presentation of information, use of rules, and table and column captions.

Technical Writing in Industry: Technique and Morale—B. C. Beookes (p. 15)

The author draws upon his experiences in teaching report writing in industrial organizations to illustrate some factors, other than the technique of the authors, which can seriously affect the quality of written reports. These factors include unsuitable working conditions, vagueness of specification, slow editing, denial of responsibility, criticism of trivia, and the constant attempt to simplify all things, however complex.

A Small-Scale Experiment in the Application of Creativity to the Teaching of Technical Communication—Richard L. Venezky (p. 21)

At Cornell University a course in technical writing and speaking, required for fifth-year electrical engineers, was redesigned for the summer co-op session to encourage creative thinking, both as an aid to the students in their general engineering work and as a more efficient medium for conveying good technical communicating processes. Students were required to "invent" devices, find answers to imaginary problems, and convey the ideas derived from these projects to other people. Emphasis was also placed on fast, efficient methods for finding information, and coherent explanations of technical problems to many different types of audiences. No attempt has been made to evaluate the long-range effects of this course on the creativity of the individuals concerned. It is felt, however, that the major accomplishments along these lines were the introduction of techniques for ideational fluency and creative problem solving and the constant reinforcement of creative efforts. Included in this paper are the original assignments for the course and a proposal for an ideal, but wholly practical, new course in technical communications based upon the results of this course.

Preparation of Commercial Instruction Manuals: A Survey and Some Comments—Ephraim M. Miller (p. 26)

A questionnaire was used to survey the operating practices of commercial instruction-manual groups in the process instrument industry. The results are presented and discussed. Subjects examined include 1) the amount of time the writer must spend on research; 2) areas of responsibility of the instruction department and other departments in the preparation of an instruction manual; 3) relations with customers; and 4) manpower shortage.

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Military Electronics

VOL. MIL-6, NO. 3, JULY, 1962

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Editorial—David B. Dobson (p. 238)

Quantitative Test Techniques for Advanced Automatic Checkout—M. C. Peterson and R. A. Kirkman (p. 240)

High-speed, digital, internally-programmed, automatic checkout equipment is a powerful test tool whose impact has not yet been felt in testing large complex systems. It has, in certain ways, much higher test capabilities than either a sequentially programmed machine or a human operator. For these capabilities to be harnessed, new test techniques are required to ensure that failures or weaknesses are detected and separated from normal drifts. These techniques involve the design-of-the-prime system to secure through instrumentation, wiring, test points, and other built-in provisions, precisely the kind and quantity of test data needed.

The test techniques involve improvement of several present-day test practices. The selection of test points, for example, is based upon both broad and detailed quantitative factors related to probability-of-mission-success, and reliability, rather than upon intuition, but with ample opportunity for the exercise of the designer's judgment. In general, more data are required, but the pay-off is in greatly increased reliability and availability, together with lower net costs.

The immediate objectives of failure detection, prediction, prevention, and isolation are defined and a test program is described whose purpose is to maximize the failure information content of the test data rather than to reconfirm abstract design parameters. One technique described is called "inferential testing" or "introspective testing" which, complemented by end-to-end testing, is especially powerful in testing closed-loop systems. Another approach and technique described is that of wringing all possible test data from a fixed set of test points by "network testing." The point is made that a good test approach really involves several complementary techniques, all related by a unified set of objectives.

The ground is laid for consideration of this "design-for-testability" of the prime system by classifying the kinds of failures and the reasons for them. The approach is taken that since failures are really deterministic, rather than random, with sufficient data of the right kinds, many can be predicted. This concept is reduced to practice by applying certain quantitatively based assumptions and a simple example given.

The kinds of resulting physical provisions in the prime system are discussed briefly. These include electrical test points and stimulus inputs and their selection, ingenious methods of instrumentation, wiring and provisions to handle prime system modifications and changes. Means of preventing the test provisions from degrading the reliability of the prime equipment are discussed. These amount to isolation means to implement the principle of "monitoring without affecting."

An example of the use of electronic test point selection techniques on portions of an aerospace system are shown, together with the quantitative analysis which led to them and the results obtained.

The discussion is summarized by reviewing the steps in design for testability, the physical results upon the prime system, and how one handles the resulting test data.

Operational Criteria for Automatic Missile Readiness Testing—Sidney I. Firstman (p. 251)

One of the activities in which automatic testing plays a key role is missile readiness testing, which is that set of tests performed regularly for the purpose of detecting failures that have occurred in the missile and launch equipment. This is done in order to maintain a high state of operational readiness, measured in terms of readiness probability. An important adjunct to criteria for physical characteristics of readiness-testing techniques and equipment are statements derived from operational analyses that specify 1) what tests are done best by each testing method, 2) the best test frequencies, and 3) the preferred equipment locations for each test; these are operational design criteria and they depend greatly upon the obtainable readiness probability. Depending on the test techniques employed, different testing methods (check periodically, monitor continuously, leave alone) and equipment locations (van, silo, missile) could yield different readiness probabilities for each missile subsystem. These probabilities depend on the particular missile's characteristics (test-point availability, failure rates, modes of operation, etc.) and test technique and equipment characteristics (test completeness, accuracy, failure rates, etc.). In addition, missile system concepts or constraints such as fixed periods for some inspections and planned delays before repair must be accounted for in the readiness determination. Using probabilistic models to describe the system operation, computational forms are developed that account for each of the above physical and operational factors, and estimate the readiness probability of either an entire missile or each of its parts. This is done for each of the three generic testing methods.

An Application of Inverse Probability to Fault Isolation—F. D. Brown, N. F. McAllister, and R. P. Perry (p. 260)

Inverse probability is applied to the problem of isolating faults in an electrical network by an analysis of failure symptoms. Prior to performing fault isolation, the performance of a network with respect to possible failures is established. Then, identifying each possible failure as a separate event, the procedure for isolating a fault is presented in terms of extracting a "signal" from random amplitude "noise." An example demonstrating these procedures is presented. Finally, an approximate solution involving less computation is summarized.

Self-Testing and Verification Techniques as Applied to an Airborne Military Guidance Computer—Wallace R. Pfeil and Norman H. Quan (p. 268)

A major problem in the field of unattended military airborne guidance equipment is the reliability and verification of proper operation. The testing techniques used to provide the maximum assurance of correct operation are a function of the level of complexity of testing. The levels of testing are:

1) Depot
2) Field Maintenance
3) Preflight
4) Inflight

This paper examines the testing requirements at each level of test. A hypothetical guidance digital computer is used as an example to present some of the techniques that could be used to meet these requirements. Techniques discussed include the computer self-test, an automonitor program, fill-compare process, protection against program lock-up, and the return echo method of detecting information transfer errors.

Computer Logic Testing by Simulation—G. N. Stockwell (p. 275)

With the advent of high-speed computers having large memory capacities, it has become possible to automate one of the more difficult and time-consuming areas of computer developmental work; namely, that of checking the logical design.

This paper describes one such method which utilizes an IBM 704 to simulate the logical equivalent of a specific computer under development by Nortronics Division of Northrop Corporation.

Such logic testing before fabrication and checkout of the first computer eliminates much of the expensive checkout time usually required. The method employed by this simulation technique can readily be adapted to other types of digital equipment.

Statistical Considerations in Element Value Solutions—R. S. Berkowitz and R. L. Wexelblat (p. 282)

This paper presents a method of fault isolation in electrical networks by determination of circuit element values from a sufficient set of measurements at the available or partly available terminals. The method requires that the circuit configuration be known; the measurements taken are voltage and current responses to sinusoidal stimuli of specified frequencies.

The method presented is an iteration one using maximum likelihood estimation techniques at each stage, the iteration accounting for the nonlinearity of equations used with respect to the element value unknowns. A specific computational procedure is evolved. An example of a sample two-terminal passive circuit with 3 elements is given in detail, complete with computer flow charts, and numerical results.

An Alternative Method for Obtaining Performance Figures—Magnus Moll (p. 288)

A performance figure indicates the ability of a system to perform a task. At present, several checkout systems obtain performance figures by employing the prime equipment directly in a mission simulation. The measure-compute approach is proposed as an alterna- which might sometimes be applied more advantageously. In the measure-compute approach, the prime equipment is not directly involved in the mission simulation; hence, the simulation need not occur in real time.

This paper first discusses general considerations related to performance figures. Then, the foundations for the measure-compute approach are presented, including techniques for overcoming certain measurement problems. Finally, two possible applications are discussed.

Implementation of Logic—R. W. House and T. Rado (p. 297)

In the development of computer programs for automatic checkout and the simplification of switching networks, certain simple logical routines occur frequently. Only a few such routines are needed to cover a large number of situations. A concise list or catalog of such routines is given in this paper. The shortest and simplest of these routines are called micro-programs. Somewhat more involved routines applying one or more micro-programs are given also. The use of the micro-programs and the more involved routines in automatic checkout is discussed.

Diagnosis in Automatic Checkout—S. E. LaMacchia (p. 302)

This paper summarizes several approaches to the problem of isolating faults in a complex system. The formulation of a diagnostic procedure involves 1) logical analysis techniques for the interpretation of results obtained from testing and 2) optimization techniques to insure that procedures employed in performing tests, selecting or sequencing tests, processing data, etc., are as efficient as possible. A procedure developed at Syracuse University, applicable when deterministic tests are used, is described. A model which extends the logical aspects of this procedure to the case when mul-

tiple failures can occur is presented. The problem of applying the figure of merit used in the single error case is discussed briefly. It is pointed out that the extended model is capable of handling a more general class of tests than those usually considered. A simple logic circuit is used to illustrate a test which will fail if two particular element failures occur together, but which will not fail if either one of the two failures occurs alone.

An Electro-Optical Test System—David M. Goodman (p. 310)

This paper describes how a host of sensors sensitive to X rays, ultraviolet and infrared, and to other electric or magnetic fields are distributed throughout a complex electronic system in such a manner that the system is continuously monitored to determine its condition of operability. The large quantity of data derived from these sensors are scanned by television techniques in order to meet the needs of rapid and automatic test. The test results are displayed on a cathode-ray tube in cyclogram fashion so that departures from normal operation may be easily and rapidly detected.

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Product Engineering and Production

VOL. PEP-6, NO. 2, JULY, 1962

High Production Encapsulation of Electronic Devices—Robert F. Zecher (p. 1)

Thermal Considerations for Plastic Encapsulation or Coating in Electronic Product Design—D. R. Fairbanks and M. Mark (p. 9)

The effects of using plastic compounds for encapsulation or coating of electronic circuitry are discussed with respect to cooling. The basic thermal mechanisms are presented. It is concluded that encapsulation or coating can significantly improve the cooling of small compact circuitry not employing direct metallic conduction paths or direct forced convection.

Vibration Slip Tables—W. X. Lamb and D. R. McCoy (p. 11)

Vibration slip tables have been developed by the testing industry to provide a means of vibrating new products and equipment in the horizontal planes, while the item under test is positioned in its normal stance. The slip tables described in this paper are oil-lubricated by a continuous oil pressure system, which permits maximum efficiency of the equipment.

A number of tests have been run on the tables to determine the response characteristics of the table. From these, data curves have been generated to show the table response at various frequencies in terms of transmissibility.

A Technique for Transient Thermal Prediction in Microminiaturized Circuits—D. M. Cawthon (p. 15)

Reliability and Quality Control

VOL. RQC-11, NO. 2, JULY, 1962

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Weapon System Design and Supportability, A Function of Failure Prediction—Vernon J. Taylor (p. 13)

A Survey of Techniques for Analysis and Prediction of Equipment and Reliability—H. E. Blanton and R. M. Jacobs (p. 18)

This paper presents a brief synopsis of representative techniques that are used in the analysis and prediction of equipment reliability during the design phase. In particular, attention is directed to: 1) techniques commonly employed for the prediction of circuit or module reliability, given part reliability, circuit configuration, and environment; 2) techniques commonly employed for the prediction of equipment (or systems) reliability, given module reliability, equipment configuration, and operational and environmental requirements; and 3) advanced statistical techniques which are useful under certain conditions to supplement those techniques previously mentioned.

Reliability's Role in Management—Harold L. Gilmore (p. 36)

IDEP—Year One at Librascope—Leonard G. Rado (p. 38)

The Interservice Data Exchange Program (IDEP) is an important new tool in the reliability test effort of General Precision's Librascope Division. IDEP's effectiveness for Librascope has expanded rapidly through use of a summary which codes salient features of pertinent test reports. The summary, issued semiannually, has helped reduce wasteful duplication of tests already conducted by IDEP member organizations. The program's manifold benefits, as reviewed in this paper, have saved Librascope about $8000 to date. While IDEP is still not functioning perfectly, the company sincerely believes that certain minor changes in the manner of failure-rate data exchange, for example, will prompt even greater benefits in the future.

Statistical Distributions in Reliability—John E. Nylander (p. 43)

This paper gives an introduction to the important probability models and statistical techniques that can be used in reliability studies.

In Section I several important ideas are stressed. They are

1) That it is generally impossible to fit a statistical distribution from data alone, but that one must usually select first a family of distributions based on some theoretical and/or physical considerations and then estimate the parameters in this family of distributions.
2) That it is important to know something of the properties of the estimators used since not all estimators are necessarily good and could possibly be worse than just picking a value for the estimate at random.

In addition to the above points important statistical quantities are defined and discussed. Also, in Section I the ideas of testing hypothesis are discussed and illustrated.

In Section II a very brief discussion of some of the important discrete distributions is made.

In Section III the probability distributions under discussion are briefly introduced.

In Section IV some guides and examples are presented on how the different family of distributions may arise.

Tolerance Limits for the Output of Systems of Cascaded Elements—Richard E. Beckwith (p. 54)

A methodology is given for analyzing the cumulative effect of a cascaded system of transducers on an input signal, where the transducers themselves activate peripheral activities, and are designed to leave the input signal unchanged, except for random magnification perturbations. Relationships are developed which relate component tolerances with the system output tolerance, and hence with system reliability. Examples are given which illustrate the type of problems that are amenable to solution by the general methodology developed, in electronics applications and otherwise. They include the determination of system output tolerances from given component tolerances, a system reliability determination when the input represents a binary digit, and a determination of minimal component precisions that are required to achieve given system tolerances. Care has been given to present the basic methodology in a complete, detailed, and self-contained manner.

Individuality in Small-Plant Reliability—Irvan J. Bearer (p. 59)

Increases in performance requirements and environmental severity, coupled with customer demands for analysis, review, testing, demonstration, training and documentation have pushed reliability costs to a critical stage. To prevent short-cutting an adequate program, reliability manpower costs must be reduced. The contrasting situations between plants with independent reliability organizations and plants with the reliability tasks distributed among existing departments will be discussed. The work atmosphere that is characteristic of many small plants and its adaptation to the organizational technique of task integration will be described. Examples will be presented of reliability work assigned according to engineering planning and procedures to those departments with personnel capable of doing the work. Guidance, unity, strength and control of a task-integrated reliability program will be shown to be the responsibility of a management staff representative. Replies to a questionnaire on the distribution of reliability manpower will be analyzed with respect to the principle that, in small plants, reliability activities are performed by each of the major departments.

Relationship of Field Reliability to AGREE Bench-Test Reliability—J. F. Thiel and H. J. Kennedy (p. 70)

The purpose of this test was to establish a quantitative relationship between AGREE bench-test results and field-test results, as measured in terms of mean time-between-failures (MTBF). The scarcity of field data on AGREE-tested equipments and the variations in field conditions and definitions of failure prevented direct comparison of field and bench results. Therefore, AGREE results were compared with composite field data on many airborne equipment types, the comparison being made in terms of a factor reflecting the relationship between equipment complexity and observed MTBF. The comparison took into account 1) two levels of test severity, 2) two degrees of optimism in interpretation of test results, and 3) differences between tubed and transistorized equipments.

Comparison shows that the observed field MTBF was generally higher than the observed bench MTBF for similar equipment types. These findings are reasonably indicative of what might be expected when additional AGREE and field test data are accumulated. In order to improve feedback of information to AGREE, it is recommended that methods and definitions used in military-sponsored field tests be standardized. These changes would appreciably increase not only the predictive value of bench tests but also the usefulness of future field evaluations to all users of field data.

A Selected Bibliography on Reliability—Harold S. Balaban (p. 86)

The Role of Components in Satellite Reliability—J. A. Morton (p. 104)

The economics of satellite communications depend critically upon the reliability of the satellite components. Assuming that the prelaunch and launch requirements on reliability have already been demonstrated, attention is focused on the orbital life requirements and the failure modes to be expected. It is proposed that our experience with electronic components for submarine cable communications is applicable to the orbital phase of satellite component reliability. The philosophy and methodology of design, fabrication, test and certification used in providing components for the Telstar experiment are outlined briefly. It is concluded that all four steps coupled with a thorough knowledge of the operating environment are required to produce an economical system.

Biographies—(p. 107)

Abstracts and References

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NOTE: The Institute of Radio Engineers does not have available copies of the publications mentioned in these pages, nor does it have reprints of the articles abstracted. Correspondence regarding these articles and requests for their procurement should be addressed to the individual publications, not to the IRE.

The Index to the Abstracts and References published in the PROC. IRE from February, 1961 through January, 1962 is published by the PROC. IRE, May, 1962, Part II. It is also published by *Electronic Technology* and appears in the March, 1962, issue of that Journal. Included with the Index is a selected list of journals scanned for abstracting with publishers' addresses.

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The number in heavy type at the upper left of each Abstract is its Universal Decimal Classification number. The number in heavy type at the top right is the serial number of the Abstract. DC numbers marked with a dagger (†) must be regarded as provisional.

UDC NUMBERS

Certain changes and extensions in UDC numbers, as published in PE Notes up to and including PE 666, will be introduced in this and subsequent issues. The main changes are:

Artificial satellites:	551.507.362.2	(PE 657)
Semiconductor devices:	621.382	(PE 657)
Velocity-control tubes, klystrons, etc.:	621.385.6	(PE 634)
Quality of received signal, propagation conditions, etc.:	621.391.8	(PE 651)
Color television:	621.397.132	(PE 650)

The "Extension and Corrections to the UDC," Ser. 3, No. 6, August, 1959, contains details of PE Notes 598–658. This and other UDC publications, including individual PE Notes, are obtainable from The International Federation for Documentation, Willem Witsenplein 6, The Hague, Netherlands or from The British Standards Institution, 2 Park Street, London, W. 1, England.

A list of organizations which have available English translations of Russian journals in the electronics and allied fields appears each June and December at the end of the Abstracts and References section.

ACOUSTICS AND AUDIO FREQUENCIES

534.232 2864

Radiation from Unbaffled Bodies of Arbitrary Shape at Low Frequencies—M. Strasberg. (*J. Acoust. Soc. Am.*, vol. 34, pp. 520–521; April, 1962.) A simple relation for calculating the sound pressure in the distant radiation field is discussed.

534.241 2865

The High-Frequency Echo Structure of Some Simple Body Shapes—A. Freedman. (*Acoustica*, vol. 12, no. 2, pp. 61–70; 1962.) Echo structure and back-scattering cross section are calculated for several shapes on the basis of results previously obtained (2521 of August).

534.32 2866

The Audibility of Irregularities of the Frequency Response in Acoustic Transmission—R. Bücklein. (*Frequenz*, vol. 16, pp. 103–108; March, 1962.) Report on subjective listening tests using music, speech and white noise. "Humps" in the response curve are more objectionable than corresponding depressions, particularly as far as speech intelligibility is concerned.

534.417 2867

Hydrophone Design for a Receiving System in which Amplifier Noise is Dominant—R. S. Woollett. (*J. Acoust. Soc. Am.*, vol. 34, pp. 522–523; April, 1962.) The importance of transducer efficiency in determining the merit of a hydrophone system [see 3222 of 1961 (Kendig)] is questioned, and a fresh approach to optimum design is considered.

534.6–8:621.385.83:537.228.1 2868

Electronic Acoustic Image Converter—R. G. Goldman. (*J. Acoust. Soc. Am.*, vol. 34, pp. 514–515; April, 1962.) A note on early design and an outline of two current models.

534.614–8 2869

Measurement of Ultrasonic Wave Velocities for Solids in the Frequency Range 100–500 Mc/s—H. J. McSkimin. (*J. Acoust. Soc. Am.*, vol. 34, pp. 404–409; April, 1962.) Phase-comparison methods are used.

534.75 2870

The Formation of Audibility Thresholds and Masking Thresholds for Pulses—H. Scholl. (*Acoustica*, vol. 12, no. 2, pp. 91–101; 1962, In German.) An investigation of the dynamic characteristics of the ear at the threshold of hearing. For corresponding work on the audibility of continuous sound see 1795 of June.

534.75 2871

The Dynamic Characteristics of the Ear in the Division of the Sound Spectrum into Frequency Groups—H. Scholl. (*Acustica*, vol. 12, no. 2, pp. 101–107; 1962. In German.) A scanning method is used to measure the time taken by the ear to separate the sound spectrum of a noise pulse into frequency groups. Calculated masking thresholds based on the measured value of about 10 msec are in good agreement with threshold values obtained by measurement. See also 2870 above.

534.76:534.61 2872

A Direction Meter for Stereophony—J. J. Geluk. (*E.B.U.*, no. 72A, pp. 50–55; April, 1962.) A simple system using two volume meters is described. One meter monitors the modulation levels of the two stereophonic channels and the other the difference in level between the channels, giving an indication of the direction in the case of intensity stereophony.

534.76:621.396.97 2873

Possibilities of Stereo-ambiophonic Sound Transmission on Two Channels—L. Keibs. (*Acoustica*, vol. 12, pp. 2, pp. 118–124; 1962. In German.) With the method discussed a sound field can be created in the reproduction room, which resembles the sound field in the room where the sound originated. See also 1796 of June.

534.793 2874

The Drive Reserve Required in Practice for Objective Loudness Meters—H. Niese. (*Hochfrequenz. und Elektroak.*, vol. 7, pp. 59–64; April, 1962.) The overload capacity of loudness meters is determined, which is required to ensure correct measurements of various types of noise at different levels.

534.84 2875

Sound Propagation in Anechoic Chambers—H. G. Diestel. (*Acustica*, vol. 12, no. 2, pp. 113–118; 1962. In German.) Theoretical investigation of the deviation of sound pressure from the $1/r$ law, where r is the distance from the sound source. The effect of the directivity of the sound source and of the microphone on the accuracy of measurements carried out in anechoic chambers is calculated.

534.84:621.395.623.8 2876

The Improvements of 'Presence' and Naturalness using Loudspeakers with Spatial Sound Pattern—W. Kuhl. (*Frequenz*, vol. 16, pp. 86–90; March, 1962.) The results of tests

with a new type of loudspeaker array for use in broadcast studios [see 1399 of 1961 (Harz)] are discussed.

534.844 2877
Artificial Reverberation—H. Kuttruff. (*Frequenz*, vol. 16, pp. 91–96; March, 1962.) Several methods of generating artificial reverberation are reviewed.

621.395.625.3 2878
Theory of Noise in Magnetic Recording Tape Coated with Magnetic Particles—I. Mikami. (*J. Appl. Phys.*, vol. 33, pp. 1591–1596; April, 1962.) For tape, magnetized to saturation, noise is shown to originate from an irregular spatial distribution of particles. This does not affect demagnetized-tape noise. The theory is shown to agree with experiment.

ANTENNAS AND TRANSMISSION LINES

621.315.212.014.12 2879
Reduction of Skin Effect by the Use of Aeolotropic Conductors—P. M. Prache. (*Câbles & Trans.* (Paris), vol. 16, pp. 73–84; April, 1962.) A general theoretical treatment of laminated and similar types of cylindrical conductor whose electrical parameters vary with the direction of measurement. Calculations are made to determine their effective resistance when used as inner conductors in coaxial circuits. Ways of reducing skin effect are demonstrated.

621.372.2 2880
Coupled Transmission Lines with Equal and Different Phase Velocity of the In-Phase and Opposite-Phase Wave—W. Burkhardtsmaier and W. Buschbeck. (*Arch. elekt. Übertragung*, vol. 16, pp. 192–197; April, 1962.) When coupled lines are used as directional or 3-db couplers wide frequency coverage can only be obtained under conditions of equal phase velocity for the two modes.

621.372.8 2881
On the Theory of Shrink Fits with Application to Waveguide Pressure Seals—A. J. Schepis. (*Bell Sys. Tech. J.*, vol. 41, pp. 885–907; May, 1962.) A theoretical and experimental investigation of the stresses and displacements which result from shrink fits between rectangular cross sections.

621.372.8:621.396.4:621.376 2882
Fundamental Consideration of Various Methods of Modulation with Regard to Waveguide Link Techniques—W. Lorek. (*Nachrtech. Z.*, vol. 15 pp. 125–129; March, 1962.) A comparison indicates the superiority of PCM for long-distance transmission by waveguide. The possibilities of using FM and AM are also investigated.

621.372.821:621.374 2883
A Strip Line for the Investigation of Sharp Pulses—D. Seitzer. (*Frequenz*, vol. 16, pp. 125–130; April, 1962.) Description of the design of, and measurements on, a homogeneous coplanar strip line with coaxial terminations having a reflection coefficient <1 per cent at frequencies up to 2.7 Gc. It is intended for use as a reference standard in the investigation of pulse transmission by inhomogeneous types of transmission line (see *e.g.*, 360 of February).

621.372.823 2884
Approximate Solutions for the Coupled-Line Equations—H. E. Rowe. (*Bell Sys. Tech. J.*, vol. 41, pp. 1011–1029; May, 1962.) The coupled-line equations for only two modes, representing the TE_{01} signal mode and a single spurious mode in circular waveguide, are solved in series form by the method of successive approximations.

621.372.823 2885
Transmission in Multimode Waveguide with Random Imperfections—H. E. Rowe and W. D. Warters. (*Bell Sys. Tech. J.*, vol. 41, pp. 1031–1170; May, 1962.) A detailed analysis is made of the effects on the TE_{01} wave transmission of random geometric imperfections produced by manufacture and assembly of circular waveguide.

621.372.823 2886
Circular Electric Wave Transmission through Helix Waveguide—K. Noda, Y. Yamaguchi and N. Suzuki. (*Rev. Elec. Comm. Lab., Japan*, vol. 10, pp. 49–70; January/February, 1962.) Experimental investigations have been undertaken to study additional losses due to mechanical imperfections and axial deformation of helix waveguide to determine the allowable dimensional tolerances and curvature of the waveguide.

621.372.823 2887
Fabrication and Characteristics of Helix Waveguide by the Vacuum Impregnation Method—K. Noda, K. Yamaguchi, N. Suzuki, H. Murasame, and C. Tanaka. (*Rev. Elect. Comm. Lab., Japan*, vol. 10, pp. 31–48; January/February, 1962.) The average attenuation constant of a straight waveguide made by this method is 1.7 db/km at 50 Gc. A bending radius of 210 m increases the attenuation constant by 1 db/km.

621.372.823:537.56 2888
Slow Wave Propagation along the Boundary of Two Coaxial Anisotropic Media—S. N. Samaddar. (*J. Electronics and Control*, vol. 12, pp. 353–368; May, 1962.) The results of a study of propagation in a metallic cylindrical waveguide excited by a magnetic ring source are applied to propagation in plasmas with zero or infinite uniform dc magnetic fields in the axial direction.

621.372.852.15 2889
Tunable Band-Pass Filters and Directional Couplers for Centimetre Waves—F. Künemund and G. Enszlin. (*Frequenz*, vol. 16, pp. 149–156; April, 1962.) The filter described consists of a waveguide resonator with coaxial terminations; accurate tuning over a wide range of mid-band frequency is achieved by means of dielectric plungers. Directional couplers comprising such band-pass filters are also discussed.

621.372.852.15 2890
YIG Microwave Electronically Tunable Preselectors—R. Blau. (*Microwave J.*, vol. 5, pp. 98–102; May, 1962.) It is possible to design light-weight efficient voltage-tuned filters using commercially available single-crystal spheres of Y-Fe garnet. Thermal drift is small and can be controlled. The design of a single unit covering the range 2–12 Gc is described. Test results indicate that units could be constructed for frequency ranges extending from 240 Mc to 40 Gc.

621.372.852.4 2891
Polarization Changers in Waveguides—H. G. Unger. (*Frequenz*, vol. 16, pp. 117–120; April, 1962.) By an appropriate combination of dielectric insert with cross-sectional deformation of the waveguide it is possible to achieve polarization changing with minimum frequency dependence of the phase delay.

621.372.853.2 2892
Backward Waves in Longitudinally Magnetized Ferrite Rods—J. Spector and A. W. Trivelpiece. (*J. Appl. Phys.*, vol. 33, pp. 2142–2143; June, 1962.) Reply to the comment by Thompson (739 of March) on 1729 of 1961 (Trivelpiece, *et al.*).

621.396.67:621.391.812.624 2893
The Aperture-to-Medium Coupling Loss in Beyond-Horizon Propagation—D. C. Hogg and L. P. Yeh. (PROC. IRE, vol. 50, pt. 1, pp. 1529–1530; June, 1962.) Comment on 1455 of May, giving further experimental and theoretical plots of coupling loss, and author's reply.

621.396.67.095.3 2894
The Representation of the Radiation Field for a Given Current Distribution—R. Kühn. (*Hochfrequenz. und Elektroak.*, vol. 71, pp. 53–59; April, 1962.) Approximation formulas for the determination of the radiation pattern of surface radiation and aerial arrays are derived and their limitations are discussed.

621.396.677(083.57) 2895
Design Charts for Microwave Antennas—G. J. Monser. (*Electronics*, vol. 35, pp. 46–50; May, 1962.) For determining gain bandwidth and volume occupied by rectangular horn and parabolic dish antennas.

621.396.677.8 2896
Off-Axis Illumination by a Plane Circular Radiator—R. G. Clarke. (*J. Appl. Phys.*, vol. 33, pp. 1614–1615; April, 1962.) Radiation flux is shown to approximate to inverse square law for points at a large distance relative to radiator radius.

621.396.677.83 2897
Stepped Parabolic Antennas—E. W. Richter. (*Microwave J.*, vol. 5, pp. 95–97; May, 1962.) To get rid of the unwanted concentrations of ultraviolet energy at the focus of mm-λ paraboloidal reflectors, a solution is proposed that requires the surface to be terraced in steps that are small in comparison with the required wavelengths and for the unwanted wavelengths. In general, the shorter wavelengths will be reflected from the plane surface of a single step, while the longer wavelengths will be focused as before.

621.396.677.83 2898
Unique Feed System Improves Space Antennas—P. D. Potter. (*Electronics*, vol. 35, pp. 36–40; June, 1962.) A simple beam-shaping modification reduces spill-over in a two-reflector Cassegrain system by a factor of two and increases aperture efficiency.

621.396.677.833 2899
28-Ft Liquid-Spun Radio Reflector for Millimetre Wavelengths—J. W. Dawson. (PROC. IRE, vol. 50, pt. 1, p. 1541; June, 1962.) Description of the manufacture of the paraboloid, general construction of the mount and beam pattern tests at $\lambda=8.5$ cm.

AUTOMATIC COMPUTERS

681.142 2900
Digital Computers—K. Mierzowski. (*VDI Z.*, vol. 104, pp. 797–803; June, 1962.) Annual review of developments in computer design with descriptions of recent installations and tabulated performance data. 46 references. For previous review see 3648 of 1961.

681.142:537.312.62 2901
Data-Addressed Memory using Thin-Film Cryotrons—V. L. Newhouse and R. E. Fruin. (*Electronics*, vol. 35, pp. 31–36; May, 1962.) Superconducting components are used throughout the system. The complete contents, and selected bits, of all words stored can be compared simultaneously.

681.142:538.221:539.23 2902
A Resonance Technique for Nondestructive Read-Out of Thin Magnetic Films—H. D. Toombs and T. E. Hasty. (PROC. IRE, vol. 50, pt. 1, p. 1526; June, 1962.) The technique is

based on UHF ferromagnetic resonance. The theory and experimental procedure are outlined.

CIRCUITS AND CIRCUIT ELEMENTS

621.3.049.7 2903
Modern Electronic Components—M. F. Tomaino. (*Electronics*, vol. 35, pp. 51–74; May, 1962.) A review covering present and future applications of solid-state physics and material technology in components production.

621.3.049.75:539.23 2904
Latest Thin-Film Circuit Techniques—E. E. Eberhard. (*Electronics*, vol. 35, pp. 37–39; June, 1962.) The design and construction of a 64-step staircase voltage generator are described.

621.3.049.75:621.382 2905
Investigation of Resistor and Capacitor Elements in a Germanium Solid Circuit—Y. Tarui and S. Denda. (*Solid-State Electronics*, vol. 5, pp. 49–53; January/February, 1962.) The resistor is a reverse-biased *p-n* junction including Sn as an alloy material; the capacitor element is formed by vapour deposition of ZnS and Al on Ge. A multivibrator is described.

621.372.413:537.312.62 2906
Nonlinear Properties of Superconducting Cavities—W. C. Eppers, E. M. Friar, and L. M. Politzer. (*J. Appl. Phys.*, vol. 33, p. 1608; April, 1962.) The nonlinear behavior induced by an RF field is investigated using a dc magnetic field as a bias.

621.372.44:621.382.3 2907
Semiconductor Inductive Elements: Parts 1 & 2—Dill. (See 3191.)

621.372.45 2908
The Geometry of Representations for Active Network—D. J. R. Stock and L. J. Kaplan. (*Arch. elekt. Übertragung*, vol. 16, pp. 223–226; May, 1962. In English.) The graphical representation of active network elements is considered. The problem of normalization of scattering matrices is discussed, and Weissfloch's theory is extended to cover the active field of elliptic transformation.

621.372.5 2909
Symmetrized Quadripoles—W. Herzog. (*Arch. Elektrotech.*, vol. 46, pp. 337–344; December, 1961.) Symmetric quadripoles for operation between different terminal resistances can be converted into unsymmetric quadripoles having the characteristics of the symmetric network. The equivalent bridge circuit of such a symmetrized quadripole is derived.

621.372.54 2910
Predistortion in Image-Parameter Filters—J. E. Colin and J. Grollemund. (*Câbles & Trans.*, vol. 16, pp. 67–69; April, 1962.) Effects similar to those produced by Darlington's predistortion method are obtained in filters with a minimum number of cutoff frequencies. Calculations are greatly simplified.

621.372.54 2911
Influence of Losses on Group Propagation Time in Ladder-Type Band-Pass Filters—J. E. Colin, B. Desdames, and P. Allemandoy. (*Câbles & Trans.*, vol. 16, pp. 70–72; April, 1962.) The effect of losses in a two-terminal element associated with an attenuation peak in a ladder filter is to reduce the group propagation time of the structure. A practical example is given.

621.372.54 2912
Reduction of the Number of Elements in Ladder-Type Filters by Use of Autotransformers—J. Gaillard. (*Câbles & Trans.*, vol. 16, pp. 123–131; April, 1962.) Several examples of a new method of simplifying filter circuits.

621.372.54:621.372.412 2913
The Introduction of Piezoelectric Crystals into Ladder-Type Pass and High-Pass Filters (with Extension to Band-Pass Filters)—J. E. Colin. (*Câbles & Trans.*, vol. 16, pp. 85–93; April, 1962.) Two methods of inserting piezoelectric crystals into previously calculated ladder structures are given.

621.372.54:621.372.57 2914
Filter Synthesis using Active *RC* Networks—M. J. Somerville and G. H. Tomlinson. (*J. Electronics and Control*, vol. 12, pp. 401–420; May, 1962.) The pass band characteristics are optimized on the basis of a minimum-mean-square-error criterion.

621.372.543.2 2915
A Theory, Free from Approximations, of Two-Circuit High-Frequency Band Fi ters with Parallel Damping and any Bandwidth—E. Peschel and K. H. R. Weber. (*Tech. Mitt. RFZ, Berlin*, vol. 6, pp. 2–10 and 45–50; March and June, 1962.) A general theory applicable irrespective of bandwidth is given for capacitively and inductively coupled tuned-circuit filters. See also 2199 of July (Weber).

621.372.553 2916
The Lossy Bridged-T Section—P. Birgels. (*Nachrtech. Z.*, vol. 15, pp. 190–195; April, 1962.) The effects of losses on the propagation coefficient and the group delay are investigated.

621.372.57 2917
Characteristic Matrices and Equivalent Circuits of Active Quadripoles—G. Marte. (*Arch. elekt. Übertragung*, vol. 16, pp. 227–229; May, 1962.) Four types of matrices are given, each representing a different relation of terminal impedance to internal impedance of a quadripole. These matrices, whose equivalent circuits indicate the quadripole characteristics, can be used in the design of cascade connections of two quadripoles.

621.372.632:621.382.333 2918
Noise in Transistor Mixer Circuits—J. S. Vogel and M. J. O. Strutt. (*Arch. elekt. Übertragung*, vol. 16, pp. 215–222; May, 1962.) An equivalent circuit is given for high-frequency noise in mixer circuits, which allows for the sources of internal noise in junction transistors. A mixer noise figure is derived whose dependence on operating point, source impedance, frequency and oscillator voltage is verified experimentally.

621.373:538.63 2919
Oscillators by means of the Magnetoresistance Effect—S. F. Sun. (PROC. IRE, vol. 50, pt. 1, pp. 1484–1493; June, 1962.) The compound InSb is the active element in the oscillator, which operates at frequencies up to 300 cps at 25°C ambient temperature. Maximum power output is 2.5 mw.

621.373.421.1 2920
Unified Treatment of Common Feedback Oscillators—V. L. Talekar. (*J. Electronics and Control*, vol. 12, pp. 433–440; May, 1962.)

621.373.43.029.42:621.383.53 2921
Glass Disk Generates Any Low-Frequency Waveform—A. H. Appleyard. (*Electronics*, vol. 35, pp. 44–45; June, 1962.) A waveform of any required shape printed on a glass disk can be generated by photoelectric scanning at disk rotation speeds from 0.005 to 50 cps.

621.373.431.1 2922
Avalanches in a Free-Running Multivibrator—A. C. Bull and J. E. Houldin. (*Electronic Engrg.*, vol. 34, pp. 455–457; July, 1962.) The abrupt changes in the triode-circuit waveform are analyzed and a method of predicting their amplitude from the static characteristic of the triode tube is given.

621.373.431.1:621.382.333 2923
Review of Bistable Flip-Flop Circuits using Complementary Transistors—F. Schreiber. (*Frequenz*, vol. 16, pp. 121–125; April, 1962.) The three fundamental types of circuit and their operation under static conditions are discussed.

621.374:[621.391+621.396.96 2924
An Auxiliary Theorem for the Theoretical and Practical Treatment of Cauer's Pulse Compression Method—D. Lohrmann. (*Frequenz*, vol. 16, pp. 156–158; April, 1962.) A simple method of calculating the time function of the output voltage of a pules compression network is given; in most cases the problem is reduced to one of calculating the transfer function of constant-group-delay filters.

621.374.32 2925
A Nondestructive and Reversible Binary Counting Circuit using Ferrite Cores and Diodes—H. B. Liem and M. J. O. Strutt. (*Arch. elekt. Übertragung*, vol. 16, pp. 189–192; 1962. Correction, vol. 16, p. 364; July, 1962. In English.) Description of the logical design of a two-digit binary counter incorporating passive elements only.

621.374.32:621.318.042.1 2926
New Idea in Counting: Incrementally Magnetized Cores—J. D. Freeman. (*Electronics*, vol. 35, pp. 40–43; June, 1962.) A multipulse magnetic counting element, the "increma," is described. It will give one output pulse for up to 16 input pulses in one stage.

621.374.32:621.387.4 2927
Counting-Rate Meter with Linear and Logarithmic Scale—G. Boucke. (*Telefunken Ztg.* vol. 35, pp. 12–19; English summary, p. 61; March, 1962.) Review of fundamental counting-rate circuits and description of a special design of two-section counting-rate meter providing a linear scale and a strictly logarithmic scale.

621.374.33:621.382.2 2928
The Use of Semiconductor Diodes in R.F. Gates—H. R. Weiss. (*Semiconductor Prod.*, vol. 5, pp. 19–23; May, 1962.) Several circuits are discussed and one is described which will give 60 db isolation at 68 Mc; it can be used for square-wave modulation up to 500 kc.

621.374.4:538.569.4 2929
Nonlinear Effects in Quantized Microwave Systems—J. R. Fontana, R. H. Pantell, and R. G. Smith. (*Electronics*, vol. 35, pp. 79–81; May, 1962.) A note on nonlinear maser operation with special reference to harmonic generation. See 2219 of July.

621.375.132.018.756 2930
The Stabilization of Pulse Amplitude in Negative-Feedback Amplifiers—K. Fränz and H. Paucksch. (*Telefunken Ztg.*, vol. 35, pp. 38–49. English summary, p. 62; March, 1962.) Design data are derived for amplifiers with a given degree of stabilization.

621.375.4 2931
A Wide-Band Transistor Logarithmic Amplifier at 45 Mc/s—R. N. Alcock. (*Electronic Engrg.*, vol. 34, pp. 444–449; July, 1962.)

621.375.9:538.569.4 2932
An Analysis of the Maser Oscillator Equations—D. M. Sinnett. (*J. Appl. Phys.*, vol. 33, pp. 1578–1581; April, 1962.) The equations

are shown not to allow periodic solutions. Previous computer solutions are found to be ambiguous, and a theory for the pulsed mode of maser operation is refuted.

621.375.9:538.569.4 2933
Microwave Generation in Ruby due to Population Inversion Produced by Optical Absorption—D. P. Devor, I. J. D'Haenens, and C. K. Asawa. (*Phys. Rev. Lett.*, vol. 8, pp. 432–435; June, 1962.) Maser action at 22 Gc has been observed in ruby using the coherent optical emission from a second ruby as a "pump" source.

621.375.9:538.569.4 2934
Filamentary Standing-Wave Pattern in a Solid-State Maser—L. Tonks. (*J. Appl. Phys.*, vol. 33, pp. 1980–1986; June, 1962.) Filamentary maser action in ruby is attributed to minute randomly distributed variations of refractive index. Thermal fluctuations at room temperature are shown to be too small to cause the filamentary characteristics.

621.375.9:538.569.4 2935
Packaged Electronically Tunable S-Band Travelling-Wave Maser System—S. Okwit and J. G. Smith. (PROC. IRE, vol. 50, pt. 1, pp. 1470–1483; June, 1962.) A ruby maser with 22-Mc bandwidth is described, tunable over an 11 per cent range of frequencies around 2250 Mc. The gain is 30 db and noise temperature 10°K; performance data are given.

621.375.9:621.372.44 2936
Parametric Effects in a Two-Level Electric Dipole System—J. R. Fontana, R. H. Pantell, and R. G. Smith. (*J. Appl. Phys.*, vol. 33, pp. 2085–2088; June, 1962.) The possibility of parametric amplification at mm λ in a two-level electric dipole system is discussed.

621.375.9:621.372.44 2937
Linear Networks with Periodically Variable Reactances: Part 1—K. Pöschl. (*Arch. elekt. Übertragung*, vol. 16, pp. 198–204; April, 1962.) The location and width of the instability regions are calculated for a loss-free two-stage parametric network. Lossy networks amenable to transformation into loss-free networks are also considered.

621.375.9:621.372.44 2938
Special Parametric Amplifiers with Tuned Circuits—M. Pommereit. (*Nachrtech.*, vol. 12, pp. 139–147; April, 1962.) The operation of several types of parametric amplifier under large-signal conditions is investigated. The two-stage amplifier operating in the normal and degenerate mode is discussed with reference to its special applications.

621.375.9:621.372.44 2939
Double-Pumped X-Band Parametric Amplifier with Extremely Large Gain-Bandwidth Product—G. C. Spacek. (PROC. IRE, vol. 50, pt. 1, pp. 1534–1535; June, 1962.) A product of almost 6 Gc is achieved using two pump sources at different frequencies and a GaAs varactor. With an amplifier gain of 40 db the over-all noise figure is 1.7 db.

621.375.9:621.372.44 2940
Broad-Band Hybrid-Coupled Parametric Amplifier—D. Sabih. (*Microwave J.*, vol. 5, pp. 87–94; May, 1962.) A unified and generalized analysis of noise theory and stability criteria is given, which can be extended to cover most types of hybrid-coupled amplifier.

621.375.9:621.372.44:523.164 2941
Continuous Operation of a Parametric Amplifier for Radio Astronomy—R. F. Clark, B. J. Dean, and A. E. Covington. (PROC. IRE, vol. 50, pt. 1, pp. 1541–1542; June, 1962.) Description and performance details of a low-noise parametric preamplifier. Precautions to ensure long-term stability are noted.

621.375.9:621.372.44:537.56 2942
Theory of a Plasma Negative-Resistance Amplifier—J. Y. Wada and Z. A. Kaprielian. (*J. Appl. Phys.*, vol. 33, pp. 1551–1562; April, 1962.) Parametric amplification in a plasma column in a modified quadrupole cavity is considered. The amplifier power gain and critical pump power are derived.

621.375.9:621.372.44:621.372.632 2943
A New Approach to Simplified Parametric-Amplifier Tuning—J. A. Luksch and V. Stachejko. (PROC. IRE, vol. 50, pt. 1, pp. 1540–1541; June, 1962.) An experimental four-frequency up-converter is described which can be rapidly tuned over the range 1230 to 1350 Mc; instantaneous bandwidth is 8 Mc and noise figure <2 db.

621.375.9:621.372.44:621.382.23 2944
Low-Noise Reactance Amplification using Refrigerated Epitaxial Silicon Varactor Diodes—K. M. Eisele and R. L. Rulison. (PROC. IRE, vol. 50, pt. 1, p. 1523; June, 1962.) Means of improving the excess noise temperature of a parametric amplifier by refrigeration are described.

621.375.9:621.372.44:621.382.3 2945
Pushing Transistors above their Frequency Limits with Parametric Conversion Operation. Rohde. (See 3192.)

621.375.9:621.382.23 2946
Tunnel-Diode Amplifiers—Chih-Ho Chen. (*Semiconductor Prod.*, vol. 5, pp. 19–25; April, 1962.) Design procedures and factors affecting design and operation of parallel-tuned tunnel-diode amplifiers are described. The stability and selection of the gain or gain-bandwidth product of the amplifier are strictly dependent upon the choice of the proper source and load resistances to match the diode negative resistance.

621.375.9:621.382.23 2947
Broad-Band Hybrid-Coupled Tunnel-Diode Amplifier in the UHF Region—H. Boyet, D. Fleri and R. M. Kurzrok. (PROC. IRE, vol. 50, pt. 1, p. 1527; June, 1962.) Performance is discussed taking account of the parasitic circuit elements.

621.375.9:621.382.23 2948
The Transfer and Noise Characteristics of a Tunnel Diode Low-Pass Amplifier—H. Rieck and K. H. Löcherer. (*Arch. elekt. Übertragung*, vol. 16, pp. 165–172; April, 1962.) Stability conditions are derived and the transfer gain and noise figure are calculated as a function of signal frequency and circuit parameters. Experimental investigations largely confirm the calculated gain figures for operation in the range 200–600 Mc.

621.376.223 2949
Crystal-Diode Modulators—G. J. Phillips and R. G. Manton. (*Electronic Tech.*, vol. 39, pp. 252–260; July, 1962.) A description is given of the development of low-power balanced modulators suitable for amplitude modulation of carriers in the range 10–1000 Mc. The modulation frequency response is substantially uniform from dc to 10 Mc. Applications include the measurement of AM suppression in FM receivers, and the generation of modulated sound and vision signals for testing television receivers.

GENERAL PHYSICS

535.2 2950
Light Modulation by the Raman Effect—R. G. Brewer. (*J. Appl. Phys.*, vol. 33, pp. 1606–1607; April, 1962.) The possibility of modulating light by the Raman effect is considered. The modulation frequencies would cover the range of 1 to 10^5 kMc.

537.312.8 2951
Theory of Tunnelling and its Dependence on a Longitudinal Magnetic Field—P. N. Argyres. (*Phys. Rev.*, vol. 126, pp. 1386–1393; May, 1962.) A quantitative theory of magneto-tunneling is presented for substances whose energy-band structure in the presence of a magnetic field can be adequately described within the approximations discussed by Yafet (1175 of 1960).

537.322:621.56 2952
Theory of Minority-Carrier Thermoelectric Cooling—M. A. Melehy. (*J. Appl. Phys.*, vol. 33, pp. 2100–2106; June, 1962.) A new thermoelectric cooling effect due to backward injection of minority carriers in a forward-biased *p-n* junction is discussed.

537.5 2953
Electron Current Pulsations in Multipacting Discharges—A. Miller and H. B. Williams. (*J. Appl. Phys.*, vol. 33, p. 1621; April, 1962.) Recent observations of RF discharges between plane parallel electrodes in a vacuum and subjected to a large axial magnetic field have shown very large periodic pulsations in the electron current which have frequencies of a few hundred kc and are simultaneous with the light pulsations.

537.525.2 2954
Temperature Dependence of Field Emission at Very Low Current Densities—M. Drechsler. (*Z. Phys.*, vol. 167, pp. 558–562; April, 1962.) Field emission between 1 and 10^{-6} a/cm² was measured for clean *W* as a function of temperature between 300 and 1350°K. A strong increase in temperature dependence was found for decreasing current density.

537.533 2955
Landau Damping of Space-Charge Waves—J. Berghammer. (*J. Appl. Phys.*, vol. 33, pp. 1499–1504; April, 1962.) Current modulation of an electron beam with a Maxwellian distribution of velocities is treated on the basis of the linearized Boltzmann equation without collisions.

537.533 2956
Reduction of Beam Noisiness in a Drifting Multivelocity Beam—D. L. Bobroff. (*J. Appl. Phys.*, vol. 33, p. 2142; June, 1962.) The analysis of multivelocity electron beams by Siegman (748 of 1958) is checked by an alternative method.

537.56 2957
Fluctuations in a Plasma: Part 2—The Numerical Factors in the Coulomb Logarithm—O. Aono. (*J. Phys. Soc. Japan*, vol. 17, pp. 853–864; May, 1962.) Calculations are made of the energy loss of a fast charged particle moving through a plasma and the relaxation rate of ion and electron temperatures. Part 1: 1208 of April.

537.56 2958
Collision Damping of Plasma Oscillations—D. F. DuBois, V. Gilinsky, and M. G. Kivelson. (*Phys. Rev. Lett.*, vol. 8, pp. 419–421; June, 1962.) The rate of damping to a two-particle final state (collision damping) has been calculated; this higher-order process is the main damping mechanism for all but very short wavelengths.

537.56:538.56 2959
Plasma Electron Oscillation and Concentration Patterns—K. G. Emeleus and A. Garscadden. (*J. Electronics and Control*, vol. 12, pp.

393–394; May, 1962.) Apparent depressions in electron concentration found during probe investigations are discussed.

537.56:538.566 2960
Oscillations of a Three-Component Assembly with or without Magnetic Field—C. Devanathan. (*Z. Phys.*, vol. 167, pp. 337–348; March, 1962. In English.) Small-amplitude wave propagation is considered in a plasma composed of singly ionized molecules, free electrons and neutral molecules, in the presence and absence of a constant homogeneous magnetic field.

537.56:621.372.8 2961
Plasma Waveguides as Low-Loss Structures—A. J. Lichtenberg and J. R. Woodyard. (*J. Appl. Phys.*, vol. 33, pp. 1976–1979; June, 1962.) The dispersion relation for EM wave propagation between two semi-infinite plasma slabs is found under the assumption of an infinite magnetic field along the slabs.

537.56.08 2962
Probe Measurements in Quiescent Plasmas—K. Takayama. (*Rev. Elect. Comm. Lab., Japan*, vol. 10, pp. 100–116; January/February, 1962.) In discharge tubes equipped with two movable grids, quiescent positive columns were obtained for a discharge current of 2–500 ma. A method of measuring the electron drift velocity in the column was developed based on the Hall effect.

537.56.08 2963
Probes for Plasma Research with Wide-Band Integrators—S. Berglund and S. Westerlund. (*Electronics*, vol. 35, pp. 44–45; June, 1962.) A compensated *RC* circuit similar to the footstrap integrator is described. With a voltage gain of three, it has a bandwidth of 18 Mc and no overshoot is caused by pulse inputs.

538.56+534.2 2964
Propagation of Randomly Perturbed Fields—R. C. Bourret. (*Canad. J. Phys.*, vol. 40, pp. 782–799; June, 1962.) A preliminary account is given of an approximation method useful in the calculation of the ensemble mean value of a field subject to perturbation by, or interaction with, a stochastic field, where the stochastic field is characterized by its two-point (bilocal) autocorrelation function. Two examples are given: wave propagation in a medium with random scalar inhomogeneities, and turbulent diffusion.

538.566:535.42 2965
The Diffraction of Electromagnetic Waves around a Finite, Perfectly Conducting Cone—F. H. Northover. (*Quart. J. Mech. Appl. Math.*, vol. 15, pt. 1, pp. 1–9; February, 1962.)

538.566:535.42 2966
Diffraction of Spherical Scalar and Vector Waves at Axial Points of a Circular Aperture and Disk—K. L. McDonald. (*J. Appl. Phys.*, vol. 33, pp. 1944–1949; June, 1962.) Analysis of EM wave diffraction in a homogeneous isotropic dielectric medium for emission from a Hertzian oscillator source and from a point source. An exact solution is given, in the vector case, at axial points of observation a) behind a circular aperture in a "black" screen and b) behind its complementary "black" disk. Scalar theory predicts considerably larger values than EM theory.

538.566:535.42 2967
Diffraction of Electromagnetic Centimetre Waves at Square Diaphragms—H. Kleineberg. (*Hochfrequenz und Elektroak.*, vol. 71, pp. 65–74; April, 1962.) The applicability of approximation methods under near-field conditions is discussed with reference to the results obtained by Koch (509 of 1961). The field distribution at $\lambda = 3.2$ cm was measured for diaphragm apertures from 1.5×1.5 to 15×15 cm.

538.569.4:538.221:621.318.134 2968
Saturation of Ferrimagnetic Resonance with Parallel Pumping—P. Gottlieb and H. Suhl. (*J. Appl. Phys.*, vol. 33, pp. 1508–1514; April, 1962.) The microwave susceptibility for subsidiary ferrimagnetic resonance with parallel pumping is calculated from the saturation amplitude of the unstable spin waves.

538.569.4:621.375.9:535.61-1 2969
Continuous Operation of a Solid-State Optical Maser—L. F. Johnson, G. D. Boyd, K. Nassau, and R. R. Soden. (*Phys. Rev.*, vol. 126, pp. 1406–1409; May, 1962.) A continuous coherent wave at 1.065 μ is obtained at 85°K, and 1.058 μ at room temperature, using Nd^{3+} in $CaWO_4$. A power output of ~10 mw is observed near threshold using silver coatings only.

538.569.4:621.375.9:535.61-1 2970
Optical Maser Characteristics $BaF_2:U^{3+}$—S. P. S. Porto and A. Yariv. (PROC. IRE, vol. 50, pt. 1, pp. 1542–1543; June, 1962.)

538.569.4:621.375.9:535.61-1 2971
Excitation, Relaxation and Optical Maser Action at 2.407 Microns in $SrF_2:U^{3+}$—S. P. S. Porto and A. Yariv. (PROC. IRE, vol. 50, pt. 1, pp. 1543–1544; June, 1962.)

538.569.4:621.375.9:535.61-2 2972
Laser = Optical Maser—H. Awender. (*Radio Mentor*, vol. 28, pp. 300–308; April, 1962.) A review of maser and laser developments with 67 references.

538.569.4:621.375.9:535.61-1 2973
Quantum Mechanical Effects in Stimulated Optical Emission—R. C. Williams. (*Phys. Rev.*, vol. 126, pp. 1011–1014; May, 1962.) Time-dependent perturbation theory is applied to 3-level systems for one-photon and two-photon emission processes. Under certain conditions frequency modulation of the maser line may be possible.

538.569.4:621.375.9:535.61-2 2974
Control of Population Inversion in Pulsed Optical Masers by Feedback Modulation—R. J. Collins and P. Kisliuk. (*J. Appl. Phys.*, vol. 33, pp. 2009–2011; June, 1962.) A shutter disk is used to modulate optical feedback, which results in an enhancement of the output power level of a pulsed ruby optical maser. An explanation of the effect is given.

538.569.4:621.375.9:535.61-2 2975
Ultrasonic Feedback Modulation of an Optical Maser Oscillator—A. J. De Maria and R. Gagosz. (PROC. IRE, vol. 50, pt. 1, p. 1522; June, 1962.) By insertion of an ultrasonic cell between the end plates of a ruby optical maser, the optical oscillation can be synchronized with the ultrasonic frequency.

538.569.4:621.375.9:535.61-2 2976
Alignment Characteristics of a Helium-Optical Maser—J. Killpatrick, H. Gustafson and L. Wold. (PROC. IRE, vol. 50, pt. 1, p. 1521; June, 1962.) Structural details of the instrument and performance data are given.

538.569.4:621.375.9:535.61-2 2977
8-Inch Ruby Amplifier—J. L. Wentz. (PROC. IRE, vol. 50, pt. 1, pp. 1528–1529; June, 1962.) A device is described using two elliptical cavities with a mutual focal line for exciting a ruby laser amplifier.

538.569.4:621.375.9:535.61-2 2978
Optical Maser Action in $CaWO_4:Er^{3+}$—Z. J. Kiss and R. C. Duncan, Jr. (PROC. IRE, vol. 50, pt. 1, p. 1531; June, 1962.)

538.569.4:621.375.9:535.61-2 2979
Pulsed and Continuous Optical Maser Action in $Ca\text{-}F_2:Dy^{2+}$—Z. J. Kiss and R. C. Duncan, Jr. (PROC. IRE, vol. 50, pt. 1, pp. 1531–1532; June, 1962.)

538.569.4:621.375.9:535.61-2 2980
Optical Maser Action in $CaF_2:Tm^{2+}$—Z. K. Kiss and R. C. Duncan, Jr. (PROC. IRE, vol. 50, pt. 1, pp. 1532–1533; June, 1962.)

538.569.4:621.375.9:535.61-2 2981
Dissociative Excitation Transfer and Optical Maser Oscillation in $Ne\text{-}O_2$ and $Ar\text{-}O_2$ RF Discharges—W. R. Bennett, Jr., W. L. Faust, R. A. McFarlane, and C. K. N. Patel. (*Phys. Rev. Lett.*, vol. 8, pp. 470–473; June, 1962.)

538.569.4:621.375.9:535.61-2 2982
Comments on "Noise in Photoelectric Mixing"—H. A. Haus and C. H. Townes: B. M. Oliver. (PROC. IRE, vol. 50, pt. 1, pp. 1544–1546; June, 1962.) Further discussion of signal-to-noise ratios of photoelectric heterodyne and homodyne receivers, with emphasis on phase fluctuations. See 820 of March.

539.124 2983
An Extensive Model of the Electron—P. A. M. Dirac. (*Proc. Roy. Soc. (London), A*, vol. 268, pp. 57–67; June, 1962.) The model consists of a charged conducting surface with surface tension. The equations of motion which relate to change of shape are deduced. The energy of the first excited state having spherical symmetry is 53 times the rest-energy and this may not represent a muon. The model does not account for spin.

GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

523.164 2984
A Survey of the Background Radiation at a Frequency of 404 Mc/s—I. I. K. Pauliny-Toth and J. R. Shakeshaft. (*Monthly Notices Roy. Astron. Soc.*, vol. 124, pp. 61–77; May, 1962.) The brightness temperature of the sky was measured with a beamwidth of 7.5° between half-power points between declinations −20° and +90°. The absolute error is about ±2.2° and the relative error for two points about ±0.8°. The results are shown in tables and graphs.

523.164:551.507.362.2 2985
Interference to Radio Astronomy from Belts of Orbiting Dipoles (Needles)—A. C. B. Lovell and M. Ryle. (*Quart. J. R. Astron. Soc.*, vol. 3, pp. 100–108; June, 1962.) Possible interference effects with radio astronomy due to powerful radio and radar transmitters operating with space needles such as those of the West Ford Project are discussed.

523.164:551.510.535 2986
An Experimental Investigation of the Scintillation of Radio Stars Observed at Frequencies of 223 and 456 Megacycles per Second from a Location Close to the Auroral Zone—C. G. Little, G. C. Reid, E. Stiltner, and R. P. Merritt. (*J. Geophys. Res.*, vol. 67, pp. 1763–1784; May, 1962.) Both phase-switch and phase-sweep interferometers were used. Scintillations during the IGY were much more intense in the auroral than in the temperate zone, and were under geomagnetic and not principally under solar control. The effects of the finite angular size and the wide band spectrum of the source are considered.

523.164:621.391.82 2987
A Note on Radio Interference at a Frequency of 408 Mc/s—Pauliny-Toth and Shakeshaft. (See 3146.)

523.164.3:523.42 2988
A Consideration of Microwave Radiation associated with Particles in the Atmosphere of Venus—C. W. Tolbert and A. W. Straiton. (*J. Geophys. Res.*, vol. 67, pp. 1741–1744; May, 1962.) Fluctuation of electrical charges on particles in the atmosphere may give rise to radiation at wavelengths between 10 and 0.86 cm.

523.165 2989
Energy Spectrum of Electrons in the Outer Radiation Belt—W. N. Hess and J. A. Poirier. (*J. Geophys. Res.*, vol. 67, pp. 1699–1709; May, 1962.) A comparison of numerical calculations and recent measurements of the outer-belt spectrum suggests that neutron decays produce a reasonable fraction of the outer-belt electrons.

523.165 2990
Rocket Observations of Solar Protons during the November 1960 Events: Part 2—L. R. Davis and K. W. Ogilvie. (*J. Geophys. Res.*, vol. 67, pp. 1711–1716; May, 1962.) Part 1: 2288 of July (Ogilvie, *et al.*).

523.165:523.75 2991
High-Energy Electrons of Solar Origin—P. Meyer and R. Vogt. (*Phys. Rev. Lett.*, vol. 8, pp. 387–389; May, 1962.) Electrons with energies >100 Mev were detected following the solar flare of July 14, 1961.

523.165:537.56 2992
Electromagnetic Properties of an Expanding Stream of Magnetized Plasma—C. G. Fälthammar. (*J. Geophys. Res.*, vol. 67, pp. 1791–1794; May, 1962.) The EM properties of magnetized plasma streams from the sun are discussed on the basis of a model in which the plasma expands radially.

523.165:550.38 2993
Propagation of Solar Corpuscles and Interplanetary Magnetic Fields—T. Obayashi. (*J. Geophys. Res.*, vol. 67, pp. 1717–1729; May, 1962.) The characteristics of interplanetary magnetic fields are inferred from the terrestrial effects of solar particles.

523.165:550.38:551.507.362.1 2994
Some Notes on Cosmic Radiations and Magnetic Fields Measured by Pioneer V—T. Obayashi. (*J. Geophys. Res.*, vol. 67, pp. 2039–2044; May, 1962.) A discussion of the interaction between solar cosmic rays and magnetic plasma clouds, with reference to Pioneer V data.

523.3:550.38 2995
The Cislunar Magnetic Field—J. H. Piddington. (*Planet. Space Sci.*, vol. 9, pp. 305–318; June, 1962.) Measurements made by Pioneer V, Lunik II and Explorer X of magnetic field and ion flux are reviewed and interpreted in terms of geomagnetic and interplanetary field models.

523.3:621.391.63 2996
Optical Echoes from the Moon—L. D. Smullin and G. Fiocco. (*Nature*, vol. 194, p. 1267; June, 1962.) Moon echoes were detected using 0.5-msec pulses of about 50 joules energy at 6934 Å, obtained from a ruby optical maser.

550.37:550.385 2997
Position and Movement of the Equatorial Electroject over Ghana—D. G. Osborne. (*J. Atmos. Terr. Phys.*, vol. 24, pp. 491–502; June, 1962.) Report on measurements of absolute vertical field and diurnal variation at two stations during February and March, 1961.

550.38 2998
The Earth's Distant Magnetic Field—E. R. Harrison. (*Geophys. J. R. Astron. Soc.*, vol. 6, pp. 479–492; June, 1962.) The theory refers to the tear-drop shape of the magnetosphere, the properties of the gegenschein, and mechanisms for accelerating electrons and protons on the night-side of the earth.

550.380 2999
An Experiment to Determine the Nature of the Earth's Distant Magnetic Field—E. R. Harrison. (*Geophys. J. R. Astron. Soc.*, vol. 6, pp. 462–467; June, 1962.) See 1252 of April.

550.385:621.391.812 3000
Correlation of Short-Period Fluctuations of the Earth's Magnetic Field and Instantaneous Frequency Measurements—K. L. Chan, D. P. Kanellakos and O. G. Villard, Jr. (*J. Geophys. Res.*, vol. 67, pp. 2066–2072; May, 1962.) Frequency fluctuations of HF CW transmissions are correlated with geomagnetic field variations during sudden commencements, geomagnetic storms and geomagnetic micropulsations. See 3018 below.

550.385.3 3001
The Influence of Continents and Oceans on Geomagnetic Variations—W. D. Parkinson. (*Geophys. J. R. Astron. Soc.*, vol. 6, pp. 441–449; June, 1962.) The vector representing changes in the field during bays and similar events tends to lie in a plane whose orientation at coastal observatories exhibits systematic behavior.

550.385.37 3002
Propagation of Hydromagnetic Waves in the Lower Exosphere and the Origin of Short-Period Geomagnetic Pulsations—J. A. Jacobs and T. Watanabe. (*J. Atmos. Terr. Phys.*, vol. 24, pp. 413–434; June, 1962.) A theory is proposed regarding the origin of micropulsations of the order of one to several seconds duration (pearls).

550.385.37:551.510.536 3003
Micropulsations and Hydromagnetic Waves in the Exosphere—F. L. Scarf. (*J. Geophys. Res.*, vol. 67, pp. 1751–1761; May, 1962.) The theory of hydromagnetic waves in the exosphere is re-examined in view of recent evidence which indicates that the steady-state hydromagnetic-wave spectrum in the lower exosphere is sharply cut off for frequencies above 1.5–3.0 cps.

550.385.4 3004
A Note on Partial Criticisms of Dynamo Theory of Magnetic Disturbance—K. D. Cole. (*J. Atmos. Terr. Phys.*, vol. 24, pp. 541–543; June, 1962.) Comment on earlier criticisms [*e.g.*, 198 of 1961 (Akasofu)] of the author's theory.

550.385.4 3005
Origin of Sudden Commencements—T. Saemundsson. (*Nature*, vol. 194, p. 1268; June, 1962.) Different mechanisms for recurrent and sporadic geomagnetic storms are described which explain the general absence of a sudden commencement in the former type of storm. Satellite measurements of the interplanetary magnetic field are used to deduce a gas density of ≥ 1 particle/cm^3 under normal conditions near the earth's orbit.

550.385.4:523.75 3006
Sudden-Commencement Hydromagnetic Waves and the Enhanced Solar Wind Direction—C. R. Wilson. (*J. Geophys. Res.*, vol. 67, pp. 2054–2056; May, 1962.) An analysis of sudden commencements indicates that the solar plasma causing the sc may be guided by the extended solar magnetic field.

551.507.362.1:550.38 3007
A Comparison of Explorer VI and Explorer X Magnetometer Data—E. J. Smith. (*J. Geophys. Res.*, vol. 67, pp. 2045–2049; May, 1962.) A comparison of data does not confirm the existence of a ring current, but suggests that a magnetic tail exists.

551.507.362.2 3008
Radio and Photographic Observations of Artificial Satellites—D. H. Shinn and N. R. Phelp. (*J. Brit. IRE*, vol. 23, pp. 465–475; June, 1962.) Observation methods and means of analyzing results are discussed. 51 references.

551.507.362.2:551.510.61 3009
Satellite Measurements of Atmospheric Structure by Refraction—L. M. Jones, F. F. Fischback and J. W. Peterson. (*Planet-Space Sci.*, vol. 9, pp. 351–352; June, 1962.) Measurements of upper air structures to heights of 25 km (25 mb) by use of a satellite-borne automatic star tracker is suggested.

551.507.362.2:621.391.812.3 3010
A Method for the Analysis of Combined Faraday and Differential Doppler Recordings in the Presence of Horizontal Gradients and Vertical Satellite Motion—E. Golton. (*J. Atmos. Terr. Phys.*, vol. 24, pp. 554–558; June, 1962.) The method is described in its exact form and approximations suitable for practical application are given.

551.507.362.2:621.391.812.33 3011
The Effect of the Earth's Magnetic Field on Measurements of the Doppler Shift of Satellite Radio Transmissions—F. de Mendonça and O. K. Garriott. (*J. Geophys. Res.*, vol. 67, pp. 2062–2065; May, 1962.) The difference between Doppler shifts of ordinary and extraordinary waves is expressed in terms of the Faraday rotation rate. Experimentally, two-frequency differential Doppler recordings may show a beat at the Faraday rotation rate.

551.510.53 3012
Upper-Atmosphere Temperatures Derived from Charged-Particle Observations—S. J. Bauer and R. E. Bourdeau. (*J. Atmos. Sci.*, vol. 19, pp. 218–225; May, 1962.) "Charged-particle temperature results obtained by three different methods (rocket and satellite-borne Langmuir probes, rocket measurements of the altitude dependence of charged-particle densities above the F2 peak, and radar incoherent back-scatter) are compared with recent models of kinetic gas temperature."

551.510.535 3013
Apparent Ionospheric Motion in the Night-Time D Region—G. S. Sales and S. A. Bowhill. (*J. Atmos. Terr. Phys.*, vol. 24, pp. 451–465; June, 1962.) Analysis of the results of vertical-incidence observations at Pennsylvania State University by means of pulse radio reflections at 60 kc. A prevailing drift velocity <20 msec, mainly towards the east, was found, but the predominant part of the drift velocity was a random component of about 22 msec with no preferred direction.

551.510.535 3014
Simplified Analysis of Ionospheric $h'(f)$ Records, using Wave Refractive Index—H. Unz. (*J. Atmos. Terr. Phys.*, vol. 24, pp. 475–481; June, 1962.) The method of analysis discussed is applicable to both the ordinary and extraordinary waves. It is based on the solution of the integral equation given in 2966 of 1961. See also 2964 of 1961 (Titheridge).

551.510.535 3015
The Anomalous Ionospheric Stratification F1·5—L. H. Heisler. (*J. Atmos. Terr. Phys.*, vol. 24, pp. 483–489; June, 1962.) The F1·5 cusp on ionograms is not a true stratification but is the result of distortion of the F layer during the passage of a traveling disturbance. The

interpretation of ionograms containing F1·5 is discussed.

551.510.535 3016
The Formation of the Daytime Peak of the Ionospheric F2 Layer—S. A. Bowhill. (*J. Atmos. Terr. Phys.*, vol. 24, pp. 503–519; June, 1962.) A simplified model, in which the loss region for the ionization has a sharp upper boundary, is shown to give a layer with defined maximum, due solely to production and diffusive transport. Comparison with the complete solution indicates that the simplified model is sufficiently accurate for heights at and above the layer maximum.

551.510.535 3017
A Spread-F Index—R. Penndorf. (*J. Atmos. Terr. Phys.*, vol. 24, pp. 543–545; June, 1962.) An index is proposed, based on the frequency spread near the critical frequencies, to indicate the degree of spread-F and its diurnal variation. Examples are given for six Arctic stations.

551.510.535 3018
On the Altitude at which some Solar-Flare-Induced Ionization is Released—D. P. Kanellakos, K. L. Chan, and O. G. Villard, Jr. (*J. Geophys. Res.*, vol. 67, pp. 1795–1804; May, 1962.) During about 20–25 per cent of optically observed flares the instantaneous frequency of an ionospherically propagated CW HF signal is changed by a few cps. In about one third of these cases the azimuthal angle of arrival is also altered. It is suggested that these effects result from the introduction of new ionization, probably in the height region 120–200 km.

551.510.535:523.746 3019
Variations in the Relation between Sunspot Number and I_{F2}—L. W. Barclay. (*J. Atmos. Terr. Phys.*, vol. 24, pp. 547–549; June, 1962.) A comparison of the annual mean values of sunspot number (R) and the F2-layer index I_{F2} suggests that, over the period 1938–1957, there has been a decrease in the amount of ionization in the F2 layer for a given sunspot number. There is evidence for a reversal of the trend since 1957.

551.510.535:550.385.4 3020
Effect of Magnetic Disturbances on the Equatorial Ionospheric Jet Current—R. W. Wright. (*Nature*, vol. 194, pp. 1169–1170; June, 1962.) The diurnal range of the vertical component of the earth's magnetic field at Ibadan, 3° south of the magnetic equator, is much reduced on magnetically disturbed days. Possible changes in the jet current which could account for this are discussed.

551.510.535:551.507.362 3021
Ionosphere Perturbations caused by a Moving Body—A. V. Gurevich. (*Isk. Sput. Zemli*, no. 7, pp. 101–124; 1961. English translation, *Planet-Space Sci.*, vol. 9, pp. 321–344; June, 1962.) A calculation of the effect of a moving body on a partially ionized gas when the body has a velocity greater than the thermal velocity of the molecules and ions and less than the velocity of the electrons.

551.510.535:551.507.362.1 3022
Local Ionospheric Disturbance Created by a Burning Rocket—J. E. Jackson, H. A. Whale, and S. J. Bauer. (*J. Geophys. Res.*, vol. 67, pp. 2059–2061; May, 1962.) After an initial increase on ignition a reduction in local electron density caused by a burning rocket was observed by Faraday rotation.

551.510.535:551.507.362.2 3023
A Simple Model for Atmospheric Density Variations from 200–800 km—A. D. Anderson. (*J. Atmos. Sci.*, vol. 19, pp. 207–217; May, 1962.) The model is based on orbital decay data from four satellites assuming that there is direct proportionality between the 10.7-cm solar RF flux and the density.

551.510.535:551.507.362.2 3024
Total Ionospheric Electron Content by Doppler Integral of the Satellite Signal—J. Mass. (*J. Atmos. Terr. Phys.*, vol. 24, pp. 549–553; June, 1962.) A method is proposed for use with satellites transmitting on a single frequency, preferably between 20 and 50 Mc. The integral under the Doppler curve is measured and compared with the theoretical integral without refraction.

551.510.535:551.594.21 3025
Thunderstorms and Sporadic-E Layer Ionization over Ottawa, Canada—A. G. Rastogi. (*J. Atmos. Terr. Phys.*, vol. 24, pp. 533–540, June, 1962.) Vertical-incidence ionospheric soundings taken during the passage of isolated thunderstorms in 1959 are studied. An increase in fE_s and in multiple radio reflections from the E_s layer indicate an association between thunderstorm occurrence and E_s-layer ionization.

551.510.535:621.391.812.63 3026
Fluctuations of Ionization in the Ionosphere at Middle Latitudes and Variation of the Usable Frequencies for Short-Wave Long-Range Communications Caused by Lunar Tides—G. Lange-Hesse and E. Schott. (*Arch. elekt. Übertragung*, vol. 16, pp. 205–212; April, 1962.) An analysis of observational data obtained at Lindau, Germany and Slough, England, reveals a distinct seasonal effect in the tidal variations of critical frequencies and MUF 3000.

551.510.535:621.391.812.63.029.51 3027
The Rate of Decrease of the Reflection Height of Long Waves at the Lower Edge of the E Layer at Sunrise—H. Koch and R. Schminder. (*J. Atmos. Terr. Phys.*, vol. 24, pp. 545–547; 1962. In German.) On the basis of field-strength recordings at 155 kc in East Germany, the sunrise effect is plotted together with the time of sunrise at 80 km for 1961. The vertical velocity of the reflection level is calculated to be 4.8 msec.

551.510.535(98/99) 3028
Universal-Time Control of the Arctic and Antarctic F Region—R. A. Duncan. (*J. Geophys. Res.*, vol. 67, pp. 1823–1830; May, 1962.) A study of f_0F2 data shows that universal-time control is very marked over the Antarctic and less marked over the Arctic. In both hemispheres maximum electron densities occur close to the time of local noon at the geomagnetic pole.

551.510.535(98/99):621.391.812.631 3029
Some Observations of Polar-Cap Absorption in the Northern and Southern Hemispheres—D. H. Jelly and C. Collins. (*Canad. J. Phys.*, vol. 40, pp. 706–718; June, 1962.) The pca events of the IGY occurred simultaneously in the Arctic and Antarctic regions.

551.510.536 3030
On the Structure of the Topside Ionosphere—S. J. Bauer. (*J. Atmos. Sci.*, vol. 19, pp. 276–278; May, 1962.) The expression previously given for the electron-density distributions above the F2 peak (1566 of May) is expanded to the more general case of a ternary-ion mixture.

551.510.536:551.507.362.2 3031
On the Evaluation of Electron Density Distribution in the Outer Ionosphere by Satellite Radio Signals—H. Uunz. (*J. Atmos. Terr. Phys.*, vol. 24, pp. 521–531; June, 1962.) In the method proposed a satellite carries a frequency-sweep transmitter for frequencies above the maximum plasma frequency of the F2 layer. The analysis of results is based on an extension of the method given in 3014 above.

551.594.5:621.391.812.63.029.62 3032
Long-Distance Propagation of Ultra-Short Waves in the Central European Area by Auroral Back-Scatter—Lange-Hesse. (See 3144.)

551.594.5:621.396.96 3033
The Effect of Absorption and Multiple Scattering on Radio-Auroral Observations—G. F. Lyon. (*Canad. J. Phys.*, vol. 40, pp. 739–748; June, 1962.) Simultaneous observations of back-scattered signals on four frequencies show an anomalous amplitude/frequency dependence which is explained by the effects of auroral-type absorption, or multiple scattering, or both.

551.594.5:621.396.96 3034
Radio-Auroral Reflection Mechanisms—G. F. Lyon and P. A. Forsyth. (*Canad. J. Phys.*, vol. 40, pp. 749–760; June 1962.) Amplitude measurements of back-scattered signals at 42, 60, 82 and 104 Mc are described. The results indicate the occurrence of both weak scattering and critical reflection, with critical frequencies up to 104 Mc.

551.594.6 3035
Doppler-Shifted Cyclotron Radiation from Electrons: a Theory of Very-Low-Frequency Emissions from the Exosphere—R. L. Dowden. (*J. Geophys. Res.*, vol. 67, pp. 1745–1750; May, 1962.) Calculations based on this theory, considering propagation conditions and various particle speeds, give frequency/time spectra similar to those observed. Further tests of the theory are discussed.

551.594.6 3036
On an Integral Equation in the Theory of Dispersion Curves of Whistlers—J. A. Faucher. (*Canad. J. Phys.*, vol. 40, pp. 794–796; June, 1962.) Laplace and Hankel transforms are used to solve exactly an equation arising in whistler theory.

551.594.6:550.386 3037
Whistlers and Magnetic Activity—R. E. Barrington and W. E. Thompson. (*Canad. J. Phys.*, vol. 40, pp. 775–781; June, 1962.) An analysis of IGY data from eight stations shows an increase of whistler occurrence with k_p up to k_p values 1–3, followed by a decrease as k_p is further increased.

551.594.6:551.510.536 3038
Exospheric Electron Temperatures from Nose Whistler Attentuation—H. B. Liemohn and F. L. Scarf. (*J. Geophys. Res.*, vol. 67, pp. 1785–1789; May, 1962.) A theory in which observed whistler cutoff arises from thermal broadening of the cyclotron resonance has been applied to Pope's new whistler data. At 4 to 4.5 earth radii an electron density of 200 to 700/cm³ and temperature of $2 \cdot 5.10^5$°K are obtained.

LOCATION AND AIDS TO NAVIGATION

621.396.9:621.391.64 3039
Infrared Applications in Navigation—C. M. Cade and C. J. Hart. (*J. Brit. IRE*, vol. 23, pp. 477–480; June, 1962.) The application of modern infrared sources including lasers makes the systems superior to radar for short-range operation.

621.396.96:629.13.052 3040
Gross Errors in Height Indication from Pulsed Radar Altimeters Operating over Thick Ice or Snow—A. H. Waite and S. J. Schmidt. (PROC. IRE, vol. 50, pt. 1, pp. 1515–1520; June, 1962.) Field measurements of electrical proper-

ties of ice and snow are used for discussion of serious reading errors which can occur in standard 440-Mc pulsed radar altimeters. See also 2612 of 1961 (Piggott).

621.396.965 3041
Array Radars—a Survey of their Potential and their Limitations—J. L. Allen. (*Microwave J.*, vol. 5, pp. 67–79; May, 1962.) The first of a series of articles covering phased arrays and electronic scanning, with emphasis on ground-based, long-range radar applications. Typical array configurations and beam-steering techniques are described and design factors are discussed.

MATERIALS AND SUBSIDIARY TECHNIQUES

535.215:546.48'221 3042
The Electrostatic Charging of Cds Single Crystals following the Passage of Current—K. W. Böer, Y. Kümmel, and K. E. Schroeter. (*Z. Phys.*, vol. 167, pp. 403–415; April, 1962.) The results of measurements on CdS crystals with In and Au contacts are analyzed statistically and a general formula for charging is derived. The influence of crystal inhomogeneities, boundary layers and field excitation is discussed.

535.215:546.48'221:537.311.4 3043
The Contact Problem on CdS Single Crystals—K. W. Böer and K. Lubitz. (*Z. Naturforsch.*, vol. 17a, pp. 397–405; May, 1962.) Au, In, Ga, Al and Cd contacts vapor-deposited on CdS single crystals at various temperatures were investigated on the basis of potential-probe measurements, I/V characteristics, electro-optical effects and noise measurements. The preparation of thermally stable Al contacts without barrier layer is described.

535.215:546.48'241 3044
On the Electrical and Optical Properties of *n*-Type Cadmium Telluride Crystals—S. Yamada. (*J. Phys. Soc. Japan*, vol. 17, pp. 645–653; April, 1962.) Ionization energies, Hall mobilities and magnetoresistive effects have been measured.

535.37:546.47'221 3045
Infrared Sensitivity of the ZnS:Cu:Co Phosphor—B. Goldstein and J. J. Dropkin. (*Phys. Rev.*, vol. 126, pp. 966–970; May, 1962.) An investigation of the effects of infrared irradiation in the temperature range −178° to +250°C on luminescence and photoconductivity.

535.37:546.47'241 3046
Mirror Absorption and Fluorescence in ZnTe—R. E. Dietz, D. G. Thomas, and J. J. Hopfield. (*Phys. Rev. Lett.*, vol. 8, pp. 391–393; May, 1962.) The optical properties of a new type of "center" found in ZnTe are described.

535.376:546.47'221 3047
Recombination Probability and Drift Velocity in Zinc Sulphide—F. Matossi. (*Z. Phys.*, vol. 168, pp. 1–5; May, 1962.) Discussion of electroluminescence phenomena observed earlier [2341 of July (Matossi & Schmid)].

537.227 3048
Evidence for a New Phase Boundary in the Forreoelectric Lead Zirconate-Lead Titanate System—H. M. Barnett. (*J. Appl. Phys.*, vol. 33, p. 1606; April, 1962.) Anomalies in the variations of dissipation factor, charge release and strain with temperature are attributed to a possible transition between two ferroelectric states.

537.227 3049
Ferroelectricity in the Compound $Bi_4Ti_3O_{12}$—P. H. Fang, C. R. Robbins, and B. Aurivillius. (*Phys. Rev.*, vol. 126, p. 892; May, 1962.) A note on dielectric properties and identification of a ferroelectric axis.

537.227 3050
Ferroelectricity in the Compound $Ba_2Bi_4Ti_5O_{18}$—B. Aurivillius and P. H. Fang. (*Phys. Rev.*, vol. 126, pp. 893–896; May, 1962.) Some ferroelectric and crystallographic properties are presented.

537.227/228.1:546.431'824–31 3051
Asymmetric Piezoelectric Hysteresis Loops in $BaTiO_3$—H. L. Stadler. (*J. Appl. Phys.*, vol. 33, pp. 2141–2142; June, 1962.) Results of piezoelectric measurements on single-crystal plates indicate that fringing effects [see 3332 of 1959 (Miller and Savage)] explain the polarization asymmetry reported by Husimi (160 of 1959).

537.227:546.431'824–31 3052
Barium Titanate Crystals Grown from Malt—F. H. Horn. (*J. Appl. Phys.*, vol. 33, pp. 1615–1616; April, 1962.) Boules from which single crystals may be cut are grown using iridium crucibles in an oxidizing atmosphere.

537.227:546.431'824–31 3053
Conductivity Injection and Extraction in Polycrystalline Barium Titanate—K. Lehovec and G. A. Shirn. (*J. Appl. Phys.*, vol. 33, pp. 2036–2044; June, 1962.) The breakdown process at temperatures between 250 and 300°C and with applied dc fields ranging from 2 to 6 kv/cm has been investigated experimentally. A detailed interpretation of the complex phenomena observed is not given.

537.227:546.431'824–31 3054
Some Phenomena of Poling Process in Reduced $BaTiO_3$ Crystal—H. Ogihara and S. Ikegami. (*J. Phys. Soc. Japan*, vol. 17, pp. 707–708; April, 1962.) The formation of complex patterns seen under a polarizing microscope is considered.

537.227:546.431'824–31 3055
Incorporation of Antimony in Barium Titanate $BaTiO_3$—O. Eberspächer. (*Naturwiss.* vol. 79, pp. 155–156; April, 1962.) Two different substitution mechanisms are identified.

537.227:546.431'824–31 3056
Electron-Microscope Display of Ferroelectric Domains in Thin $BaTiO_3$ Single-Crystal Films—H. Pfisterer, E. Fuchs, and W. Liesk. (*Naturwiss.*, vol. 49, pp. 178–179; April, 1962.) Two transmission images obtained by electron microscope are reproduced; the crystal thickness is of the order of 1000 Å.

537.227:546.431'824–31 3057
Discussion on the Crystal Growth of Barium Titanate Single Crystals in KF Flux Fusion—S. Waku. (*Rev. Elect. Comm. Lab., Japan*, vol. 10, pp. 1–19; January/February, 1962.)

537.311.33 3058
Double Injection with Negative Resistance in Semi-insulators—N. Holonyak, Jr., S. W. Ing, Jr., R. C. Thomas, and S. F. Bevacqua. (*Phys. Rev. Lett.*, vol. 8, pp. 426–428; June, 1962.) The experimental results are presented of measurements of negative resistance at room temperature in GaAs, Si and Ge *p-si-n* diodes due to double injection. The results are believed to confirm Lampert's theory (1977 of June).

537.311.33 3059
Experimental Evidence for Helical Instabilities in a Semiconductor Plasma—F. Okamoto, T. Koike, and S. Tosima. (*J. Phys. Soc. Japan*, vol. 17, pp. 804–807; May, 1962.) The magneto-oscillatory plasma effect in *n*-type Ge has been investigated using probes on the sides of an oscillistor specimen. Experimental results give strong evidence for the existence of the helical plasma instability predicted by Glicksman (1197 of April).

537.311.33 3060
Direct Lattice Absorption associated with Shallow Impurity Levels—J. R. Hardy (*Phil. Mag.*, vol. 7, pp. 953–960; June, 1962.) A detailed theory is developed of single-phonon absorption arising from the failure of the adiabatic approximation for shallow impurity levels.

537.311.33 3061
Treatment of Ionized Impurity Scattering in Degenerate Semiconductors. Application of the Variational Technique in the Partial-Wave Method—P. Csavinszky. (*Phys. Rev.*, vol. 126, pp. 1436–1440; May, 1962.) The scattering process is adequately described by considering only the phase shift of zero order. An analytical formula is derived for the resistivity due to ionized-impurity scattering alone.

537.311.33:535.215 3062
Electrical Conduction and the Photovoltaic Effect in Semiconductors with Position-Dependent Band Gaps—P. R. Emtage. (*J. Appl. Phys.*, vol. 33, pp. 1950–1960; June, 1962.) The transport equations are modified for application of an illuminated isothermal semiconductor whose band gap is a linear function of one of the position coordinates. The use of such a material for a solar-energy converter is discussed, and the maximum efficiency is calculated to be 43 per cent.

537.311.33:537.312.9 3063
Isotropically Piezoresistive Semiconductor—W. G. Pfann. (*J. Appl. Phys.*, vol. 33, pp. 1618–1619; April, 1962.) The preparation of a Ge-Si alloy with large piezoresistive effects for use in stress transducers is considered.

537.311.33:539.23:535.3–15 3064
Infrared Transmission Factor and Structure of PbS, PbSe and PbTe Films Condensed by Quenching—G. Berthold, R. Hilsch, and W. Sander. (*Z. Phys.*, vol. 167, pp. 493–500; April, 1962.)

537.311.33:[546.28+546.289 3065
Correlation between Maximum Solid Solubility and Distribution Coefficient for Impurities in Ge and Si—S. Fischler. (*J. Appl. Phys.*, vol. 33, p. 1615; April, 1962.) Correlation is linear over six decades of distribution coefficient.

537.311.53:[546.28+546.289 3066
The Change of Current-Voltage Characteristics by Light Illumination on Compensated Ge and Si, at Low Temperature—K. Komatsubara and H. Kurone. (*J. Phys. Soc. Japan*, vol. 17, pp. 880–881; May, 1962.) The critical voltage at which the mechanism changes from impurity conduction to normal conduction at liquid He temperatures is studied.

537.311.33:546.28 3067
Stacking Faults in Epitaxial Silicon—H. J. Queisser, R. H. Finch, and J. Washburn. (*J. Appl. Phys.*, vol. 33, pp. 1536–1537; April, 1962.) Faults have been detected with optical and electron microscopes.

537.311.33:546.28 3068
X-Ray Observations of Diffusion-Induced Dislocations in Silicon—G. H. Schwuttke and H. J. Queisser. (*J. Appl. Phys.*, vol. 33, pp. 1540–1542; April, 1962.)

537.311.33:546.28 3069
Low-Voltage Electron-Diffraction Study of the Oxidation and Reduction of Silicon—J. J. Lander and J. Morrison. (*J. Appl. Phys.*, vol. 33, pp. 2089–2092; June, 1962.)

537.311.33:546.28 3070
On the Trapping and Recombination Centres in Plastically Bent Silicon—K. Matsuura and Y. Inuishi. (*J. Phys. Soc. Japan*, vol. 17, pp. 874–875; May, 1962.) The photoconductivity decay curve before and after plastic bending of the specimen showed considerable difierences. Components due to trap formation and thermal release of trapped holes are identified.

537.311.33:546.28 3071
Surface Conductivity of Cleaved Silicon Surfaces—P. Handler. (*Phys. Rev.*, vol. 126, pp. 971–975; May, 1962.) A reinterpretation of experimental results (1910 of 1961 [Palmer, *et al.*)] in terms of a large density of states below the center of the forbidden energy gap.

537.311.33:546.28 3072
Infrared Absorption Spectrum of Sulphur-Doped Silicon—W. E. Krag and H. J. Zieger. (*Phys. Rev. Lett.*, vol. 8, pp. 485–487; June, 1962.)

537.311.33:546.28 3073
Negative Resistance of Silicon p-n Junctions at 4.2°K—G. E. Smith and D. Kahng. (*Solid-State Electronics*, vol. 5, pp. 177–178; May/June, 1962.) Impact ionization of neutral donors in p-n junctions is studied.

537.311.33:546.28:539.23 3074
X-Ray Observations of Partial Dislocations in Epitaxial Silicon Films—G. H. Schwuttke. (*J. Appl. Phys.*, vol. 33, pp. 1538–1540; April, 1962.) Dislocation and stacking faults have been detected by X-ray diffraction microscopy.

537.311.33:546.28:621.3.032 3075
Technique for Ohmic Connecting Leads to Silicon—E. Matsuura, K. Matsui, and R. R. Hasiguti. (*J. Appl. Phys.*, vol. 33, pp. 1610–1611; April, 1962.) A cold-current bond is formed using a pressure wedge and local electrical heating.

537.311.33:[546.289+546.682'86 3076
The Plastic Deformation of Germanium and InSb—P. Haasen. (*Z. Phys.*, vol. 167, pp. 461–467; April, 1962.) The initial stages of the deformation process are investigated.

537.311.33:546.289 3077
On the Low-Temperature Electrical Breakdown of Highly Compensated Germanium—K. Komatsubara and H. Kurono. (*J. Phys. Soc. Japan*, vol. 17, pp. 883–884; May, 1962.) Results of measurements at 4.2 and 2°K are illustrated.

537.311.33:546.289 3078
Magnetoresistance in Heavily Doped n-Type Germanium—Y. Furukawa. (*J. Phys. Soc. Japan*, vol. 17, pp. 630–638; April, 1962.) The dependence of the magnetoresistance on temperature and donor concentration has been investigated. The symmetry relation is obeyed at 300°K; at lower temperatures the behavior depends on whether the donor concentration lies in the impurity or metallic conduction range.

537.311.33:546.289 3079
Etch Patterns in Germanium—T. Arizumi and I. Akasaki. (*J. Phys. Soc. Japan*, vol. 17, p. 712; April, 1962.) A note on the patterns observed on etched surfaces after subsequent treatment with a KI-I_2 solution.

537.311.33:546.289 3080
Optical Absorption of Arsenic-Doped Degenerate Germanium—J. I. Pankove and P. Aigrain. (*Phys. Rev.*, vol. 126, pp. 956–962; May, 1962.) Data on infrared transmission at 4.2°K are used to calculate the energy gap.

537.311.33:546.289 3081
Method for Visualizing the Distribution of Current in Alloyed Germanium p^+-n Junctions—T. Tokuyama. (*Solid-State Electronics*, vol. 5, pp. 135–142; May/June, 1962.) Electrochemical changes ocurring on the surface of a junction are viewed with a microscope.

537.311.33:546.289 3082
Zener Breakdown in Alloyed Germanium p^+-n Junctions—T. Tokuyama. (*Solid-State Electronics*, vol. 5, pp. 161–169; May/June, 1962.) The breakdown of alloyed Ge p^+-n junctions is studied, and the conditions for and characteristics of breakdown controlled by the Zener and avalanche effects are reported.

537.311.33:546.289 3083
The Generation and Movement of Single Dislocations in Germanium Crystals—F. Meier. (*Z. Phys.*, vol. 168, pp. 29–41; May, 1962.) X-ray pictures are reproduced showing dislocation sources in plastically deformed Ge crystals.

537.311.33:546.289:538.594.4 3084
Cyclotron Resonance of Hot Electrons in Pure Germanium—E. Hanamura, T. Inui, and Y. Toyozawa. (*J. Phys. Soc. Japan*, vol. 17, pp. 666–675; April, 1962.)

537.311.33:546.289:539.12.04 3085
Radiation-Induced Recombination and Trapping Centres in Germanium: Part 2—Annealing in Gamma-Irradiated, Antimony-, and Arsenic-Doped Material—O. L. Curtis, Jr., and J. H. Crawford, Jr. (*Phys. Rev.*, vol. 126, pp. 1342–1349; May, 1962.)

537.311.33:546.289:539.23 3086
On the Structural Properties of Vapour-Deposited Germanium Layers—T. Arizumi and I. Akasaki. (*J. Phys. Soc. Japan*, vol. 17, p. 714; April, 1962.) Discusses the effect of impurities and dislocations on the layer formation.

537.311.33:546.57'72'241 3087
Note on the Constitution of the Semiconducting 'Compound' $AgFeTe_2$—P. Manca and F. Massazza. (*J. Appl. Phys.*, vol. 33, pp. 1608–1609; April, 1962.) Tests on $AgFeTe_2$ show that the two phases present are Ag_2Te and Fe_2Te_3.

537.311.33:546.681'18:535.215 3088
***p-n*-Junction Photovoltaic Effect in Zinc-Doped GaP**—H. G. Grimmeiss, W. Kischio, and H. Koelmans. (*Solid-State Electronics*, vol. 5, pp. 155–159; May/June, 1962.) Extrinsic and intrinsic photovoltaic effects are found in Zn-doped GaP.

537.311.33:546.681'19 3089
Rapid Zinc Diffusion in Gallium Arsenide—R. L. Longini. (*Solid-State Electronics*, vol. 5, pp. 127–130; May/June, 1962.) When Zn doping is very heavy, a small fraction of the Zn should go into donor interstitial positions and diffuse very rapidly.

537.311.33:546.681'19 3090
Behaviour of Manganese in GaAs—L. J. Vieland. (*J. Appl. Phys.*, vol. 33, pp. 2007–2008; June, 1962.) Report of Hall-effect and mobility measurements on Mn-doped GaAs crystals.

537.311.33:546.681'19:535.215 3091
Photoemission of Holes from Tin into Gallium Arsenide—R. Williams. (*Phys. Rev. Lett.*, vol. 8, pp. 402–404; May, 1962.) Experimental evidence is presented.

537.311.33:546.873'241 3092
Temperature Dependence of Hall Coefficient in p-Type Bi_2Te_3 Crystals—K. Shogenji and T. Sato. (*J. Phys. Soc. Japan*, vol. 17, p. 727; April, 1962.) Various band structures are considered.

537.311.33:621.382.032.27 3093
Accurate Solution of an Idealized One-Carrier Metal-Semiconductor Junction Problem—J. R. Macdonald. (*Solid-State Electronics*, vol. 5, pp. 11–37; January/February, 1962.)

537.311.33:621.387.462 3094
Doping of Crystals by Ion Bombardment to Produce Solid-State Detectors—T. Alväger and N. J. Hansen. (*Rev. Sci. Instr.*, vol. 33, p. 567; May, 1962.) A technique for the production of Si charged-particle detectors is described.

537.312.62 3095
Current Distribution in Thin Superconducting Films—E. H. Rhoderick and E. M. Wilson. (*Nature*, vol. 194, pp. 1167–1168; June, 1962.) Good agreement is obtained between experimental and theoretical distributions.

537.312.62 3096
Magnetic-Field Dependence of the Surface Impedance of Superconducting Tin at 3 kMc/s—P. L. Richards. (*Phys. Rev.*, vol. 126, pp. 912–918; May, 1962.) Measurements have been made as a function of crystal orientation and temperature for pure Sn and Sn-In alloy. Results are at variance with the Dresselhaus theory.

537.312.62 3097
Tunneling into Superconductors at Temperatures below 1°K—I. Giaever, H. R. Hart, Jr., and K. Megerle. (*Phys. Rev.*, vol. 126, pp. 941–948; May, 1962.) Results of measurements on Pb, Sn, In and AC are discussed in relation to theory.

537.312.62 3098
Solution of the Gap Equation for a Superconductor—G. J. Culler, B. D. Fried, R. W. Huff, and J. R. Schrieffer. (*Phys. Rev. Lett.*, vol. 8, pp. 399–402; May, 1962.)

537.312.62 3099
Superconductivity of Iridium—R. A. Hein, J. W. Gibson, B. T. Matthias, T. H. Geballe, and E. Corenzwit. (*Phys. Rev. Lett.*, vol. 8, pp. 408–410; May, 1962.)

537.312.62:537.311.31 3100
The Characteristics of Superconductors in Contact with Normal Conductors—P. Hilch. (*Z. Phys.*, vol. 167, pp. 511–524; April, 1962.) Measurements of transition temperature as a function of film thickness were carried out on superimposed Ca and Pb films vapor-deposited on a quartz plate.

537.312.62:538.63 3101
The Magnetic Detection of Superconductivity of Vapour-Deposited Metal Films—R. Hilsch, D. Korn, and G. U. Minnigerode. (*Z. Phys.*, vol. 167, pp. 501–510; April, 1962.) Measuring apparatus and results obtained with it are described.

537.312.62:538.63 3102
Observed Sign Reversal of a Magnetic Field Penetrating a Superconductor—K. E. Drangeid and R. Sommerhalder. (*Phys. Rev. Lett.*, vol. 8, pp. 467–469; June, 1962.) Tin films were deposited on the outside surface of a glass cylinder and cooled to liquid-He temperature. An alternating magnetic field was ap-

plied externally parallel to the axis and shown to have reverse sign inside the cylinder.

537.323:546.817'241 3103
Measurements of the Room-Temperature Thermoelectric Parameters of Lead Telluride—G. W. Johnson. (*J. Electronics and Control*, vol. 12, pp. 421–432; May, 1962.) Measurements on single-crystal *n*- and *p*-type PbTe over a range of carrier concentrations are reported.

538.221 3104
Ferromagnetic Anisotropy Induced by Cold Rolling of Dilute Ferromagnetic Cu-Co Alloys—T. Mitui, M. Satô, and T. Sambongi. (*J. Phys. Soc. Japan*, vol. 17, pp. 639–644; April, 1962.) The magnitude of the induced anisotropy has been measured. The effects of homogeneous and inhomogeneous deformation are considered.

538.221:537.312.9 3105
Elastoresistance Effect in Iron Single Crystals—M. C. Martin and D. R. Morrow. (*J. Appl. Phys.*, vol. 33, pp. 1905–1908; June, 1962.) Experimental investigations of the dependence of elastoresistance and elastoresistivity on crystal orientation and deformation temperature.

538.221:538.63 3106
Anomalous Hall Effect and Magnetoresistance of Ferromagnetic Metals—J. Kondo. (*Progr. theoret. Phys.*, vol. 27, pp. 772–792; April, 1962.) A theoretical treatment on the basis of the localized *d*- or *f*-electron model. Results are compared with experimental data.

538.221:538.632 3107
Temperature Dependence on the Hall Effect in Iron and Nickel—T. Okamoto, H. Tange, A. Nishimura, and E. Tatsumoto. (*J. Phys. Soc. Japan*, vol. 17, pp. 717–718; April, 1962.) The ordinary and extraordinary Hall coefficients are treated separately.

538.221:538.652 3108
Dynamic Characteristics of an Iron-Cobalt Alloy for Magnetostrictive Transducers—O. Henkel. (*Hochfrequenz und Elektroak.*, vol. 71, pp. 47–53; April, 1962.) Report on investigations of the dependence of magnetic and magnetostrictive properties on heat treatment conditions for a 43 per cent Co alloy.

538.221:539.23 3109
Magnetic Properties of Ferromagnetic Films—H. Rotner. (*Z. Phys.*, vol. 168, pp. 42–45 and 148–154; May, 1962.) Part 1—Stochastic Theory of the Fine Structure of Magnetization (pp. 42–54). [Part 2—Theory of Domain Splitting (pp. 148–154).]

538.221:539.23:621.318.57 3110
Noncoherent Switching in Permalloy Films—D. O. Smith and K. J. Harte. (*J. Appl. Phys.*, vol. 33, pp. 1399–1413; April, 1962.) High-resolution Bitter pattern studies of the domain structure of permalloy films, with uniaxial anisotropy and under the influence of applied fields in the film plane are reported.

538.221:539.23:681.142 3111
Silicon Monoxide Undercoating for Improvement of Magnetic Film Memory Characteristics—B. I. Bertelsen. (*J. Appl. Phys.*, vol. 33, pp. 2026–2030; June, 1962.) Experimental results show the leveling effect of thick layers of SiO on rough substrate surfaces. This makes possible the use of metallic substrates for Ni-Fe film storage elements.

538.221:621.318.134 3112
Frequency Doubling in Anisotropic Ferrites—D. D. Douthett, I. Kaufman, and A. S. Risley. (*J. Appl. Phys.*, vol. 33, pp. 1395–1396; April, 1962.) The high planar anisotropy of single-crystal Zn_2Y enhances its frequency-doubling behavior. Experimental results agree well with the predictions of the equation of motion for the magnetization with anisotropy and losses included.

538.221:621.318.134 3113
Ferrite Crystals using an Arc Image Furnace—R. P. Poplawsky. (*J. Appl. Phys.*, vol. 33, pp. 1616–1617; April, 1962.)

538.221:621.318.134 3114
Measurement on Magnetocrystalline Anisotropy of Mn-Zn-Fe Ferrites—K. Ohta, M. Hoshino, and Y. Araki. (*J. Phys. Soc. Japan*, vol. 17, pp. 879–880; May, 1962.)

538.221:621.318.134 3115
Nickel-Zinc and Manganese-Zinc Ferrites of High Initial Permeability—H. Henniger and H. Glitschka. (*Nachrtech.*, vol. 12, pp. 105–108; March, 1962.) The effect of impurites and sintering temperature on the magnetic properties of high-permeability ferrites are discussed.

538.221:621.318.134 3116
Anisotropy of Exchange in Ytterbium Iron Garnet—K. A. Wickersheim and R. L. White. (*Phys. Rev. Lett.*, vol. 8, pp. 483–485; June, 1962.)

538.221:621.318.134:534.27 3117
Acoustic Wave Rotation by Magnon-Phonon Interaction—H. Matthews and E. C. LeCraw. (*Phys. Rev. Lett.*, vol. 8, pp. 397–399; May, 1962.) The plane of polarization of a transverse elastic wave propagating parallel to a magnetic field in a ferromagnet undergoes measurable rotation. This provides an evaluation of magnetoelastic coupling.

538.221:621.318.134:538.569.4 3118
Ferrimagnetic Resonance in Polycrystalline Thulium-Iron Garnet ($Tm_3Fe_5O_{12}$)—T. Miyadai. (*J. Phys. Soc. Japan*, vol. 17, pp. 870–871; May, 1962.)

538.222:538.569.4 3119
Paramagnetic Resonance of Tb^{3+} Ions in $CaWO_4$ and CaF_2—P. A. Forrester and C. F. Hempstead. (*Phys. Rev.*, vol. 126, pp. 923–930; May, 1962.)

538.222:538.569.4:621.375.9 3120
On the Theory of Cross-Relaxation in Maser Materials: Part 2—Harmonic Spin Coupling—M. Hirono. (*J. Phys. Soc. Japan*, vol. 17, pp. 788–795; May, 1962.) The process of harmonic spin coupling of three neighboring ions in the crystalline electric field of a maser material is examined theoretically using the method of moments. Part 1: 4239 of 1961.

538.22:546.682'873 3121
The de Haas-van Alphen Effect of InBi—Y. Saito. (*J. Phys. Soc. Japan*, vol. 17, p. 716; April, 1962.) Configurations of the Fermi surface are discussed in relation to torque measurements at 1.5°K.

538.222:546.824–31 3122
Measurements of the Dielectric Constants of Rutile (TiO_2) at Microwave Frequencies between 4.2°K and 300°K—E. S. Sabisky and H. J. Gerritsen. (*J. Appl. Phys.*, vol. 33, pp. 1450–1453; April, 1962.) Rutile, because of its high dielectric constant, has important use as a maser material.

548.5 3123
Point Defect Trapping in Cyrstal Growth—W. W. Webb. (*J. Appl. Phys.*, vol. 33, pp. 1961–1971; June, 1962.) Under some conditions point defects may be trapped during growth, but the effect is mostly negligible. Growth trapping may account for the loss of dislocations from the metal whisker crystals during growth.

548.5 3124
A Process for Obtaining Single Crystals with Uniform Solute Concentration—A. V. Valčić. (*Solid-State Electronics*, vol. 5, pp. 131–134; May/June, 1962.) Crystals are pulled from a system of joined vessels. The theory of the method is given and experimental results are noted.

MEASUREMENTS AND TEST GEAR

621.3.018.41(083.74) 3125
Calibrating Frequency Standards with V.L.F. Transmissions—J. D. Echols. (*Electronics*, vol. 35, pp. 60–63; April, 1962.) A system is described with a stability within ± 1 μs and accuracy within few parts in 10^{11} in spite of signal amplitude fluctuations of the order of 30 db.

621.317.3.029.6:537.311.33 3126
Measurements of Lifetime of Carriers in Semiconductors through Microwave Reflection—S. Deb and B. R. Nag. (*J. Appl. Phys.*, vol. 33, p. 1604; April, 1962.) A microwave reflection technique for measuring the lifetime of carriers in semiconductors is described which has certain advantages over previous techniques.

621.317.42:538.632 3127
Measurements of Static and Dynamic Magnetic Fields—H. Sima. (*Elektrotech. Z., Edn B*, vol. 14, pp. 175–178; April, 1962.) Instruments based on the Hall effect are described and the errors to which they are liable are mentioned.

OTHER APPLICATIONS OF RADIO AND ELECTRONICS

536.53:621.382.2/.3 3128
Measuring Temperature with Diodes and Transistors—L. E. Barton. (*Electronics*, vol. 35, pp. 38–40; May, 1962.) Circuit arrangements are described for obtaining a direct temperature reading over wide ranges in terms of diode voltage drop or transistor base bias for constant collector current.

621.362:537.324 3129
Effects of Surface Heat Transfer on the Performance of a Thermoelectric Generator—L. J. Ybarrondo and J. E. Sunderland. (*Solid-State Electronics*, vol. 5, pp. 143–154; May/June, 1962.) A theoretical analysis relating to the efficiency of thermoelectric generators.

621.362:621.387 3130
The Effect of Staging on Thermionic Converters—G. H. Hatsopoulos. (*Commun. and Electronics*, pp. 36–40; March, 1962.) By operating several thermionic stages thermally connected in series between two heat reservoirs of given temperature a considerable improvement in the over-all efficiency is achieved.

621.362:621.387 3131
Electron Space-Charge-Limited Operation of Cesium Thermionic Converters—E. N. Carabateas. (*J. Appl. Phys.*, vol. 33, pp. 1445–1459; April, 1962.) The operation of Cs thermionic converters with low emitter temperatures and high Cs pressures may be space-charge-limited. A quantitative analysis of this mode is presented.

621.383.8:677.52 3132
Fibre Optics for Electronics Engineers—G. V. Novotny. (*Electronics*, vol. 35, pp. 37–42; June, 1962.) Applications of the technique of transmitting light through long thin fibers of transparent material are illustrated, with

particular reference to image display techniques. Data processing and laser applications are mentioned.

621.385.833 3133
On an Analytical Expression for the Axial Field of Electromagnetic Lenses—M. L. De. (*Phil. Mag.*, vol. 7, pp. 1065–1067; June, 1962.)

621.385.833:537.533.73 3134
A New Method of Preparing Electron-Diffraction Patterns of Very Small Selected Areas of Electron-Microscope Transmission Specimens—W. D. Riecke. (*Optik, Stuttgart*, vol. 19, pp. 273–286; May, 1962.)

621.387.4:539.16 3135
Detection of Nuclear Explosions in Space and Underground—I. Maddock. (*J. Brit. IRE*, vol. 23, pp. 415–427; June, 1962.) Techniques for detecting EM and particle radiation and for seismic measurement are reviewed.

PROPAGATION OF WAVES

621.391.812.34.029.53 3136
The Dependence of Soil Conductivity and Field Strength in the Medium-Wave Range—K. Müller. (*Tech. Mitt. RFZ, Berlin*, vol. 6, pp. 11–17; March, 1962.) Results of soil-conductivity and field-strength measurements are given. The field-strength distribution over a given propagation path can be expressed in terms of an equivalent numerical distance which allows for both soil conductivity and absorption due to vegetation. Further investigations of the latter absorption effects are required.

621.391.812.61:551.594.2 3137
Field-Strength Enhancements and Thunderstorms over Decimetre-Wavelength Propagation Paths beyond Optical Range—U. Kühn. (*Tech. Mitt. RFZ, Berlin*, vol. 6, pp. 23–28; March, 1962.) Report of field-strength recordings at 1.1 Gc over four paths of length 100–300 km. Increases in field strength of up to 43 db have been observed during thunderstorms. Measurements of temperature and humidity gradients near the reception point confirm the explanatory theory advanced for this effect.

621.391.812.62.029.63 3138
Measurements of Propagation at 480 Mc/s and 780 Mc/s in the Mountains of the Lower Tatra (Č.S.S.R.)—F. Kralik, U. Kühn, and V. Vočadlo. (*Tech. Mitt. BRF, Berlin*, vol. 5, pp. 174–177; December, 1961, and *Tech. Mitt. RFZ, Berlin*, vol. 6, pp. 18–22; March, 1962.) Report on field-strength measurements in very irregular terrain. The frequency—and height—dependence of signal propagation was determined.

621.391.812.624:621.396.67 3139
The Aperture-to-Medium Coupling Loss in Beyond-Horizon Propagation—Hogg: Yeh. (See 2893.)

621.391.812.63 3140
Long-Distance One-Hop Ionospheric Radio-Wave Propagation—D. B. Muldrew and R. G. Maliphant. (*J. Geophys. Res.*, vol. 67, pp. 1805–1815; May, 1962.) A ray-tracing technique has been used to determine the characteristics of one-hop propagation in both temperate and equatorial regions. The results are compared with frequency-sweep soundings over a 5600-km path.

621.391.812.63 3141
Scattering of Radiation by the-Fluctuations in a Nonequilibrium Plasma—O. Buneman. (*J. Geophys. Res.*, vol. 67, pp. 2050–2053; May, 1962.) The theory of back-scatter from plasma fluctuations is formulated using the statistical approach, in the more general case of unequal electron and ion temperatures.

621.391.812.63.029.45 3142
The Propagation of Very-Low-Frequency Radio Waves over Distances up to 2000 km—J. K. Hargreaves and R. Roberts. (*J. Atmos. Terr. Phys.*, vol. 24, pp. 435–450; June, 1962.) The amplitude of VLF radio waves was observed at distances up to 2000 km from transmitters, with particular emphasis on diurnal variations at ranges of 1000 km. A model of the ionosphere, which assumes a sharp lower boundary, is inadequate for an interpretation of these and earlier short-range observational results.

621.391.812.63.029.51/.53 3143
Ionospheric Propagation on Long and Medium Waves: Results of an Investigation Organized by the European Broadcasting Union—W. Ebert. (*E.B.U.*, pp. 3–11, 68–80 and 109–121; February, April and June, 1962.) The aims and results of the various measurement campaigns organized by E.B.U. Working Party B from 1952 to 1960 are reviewed. The analysis and correction of the statistical data are described, and propagation curves, as functions of distance and of frequency, obtained from these investigations are given together with details of their practical application.

621.391.812.63.029.62:551.594.5 3144
Long-Distance Propagation of Ultra-Short Waves in the Central European Area by Auroral Back-Scatter—G. Lange-Hesse. (*Arch. elekt. Übertragung*, vol. 16, pp. 251–261; May, 1962.) Review of the results obtained from an analysis of observational data gathered in Germany from 1957 to the beginning of 1961. Use was made of the network of amateur stations in investigating "bistatic aurora communications" at 144–146 Mc. Further work at these frequencies and at 28 and 21 Mc is proposed. 61 references.

RECEPTION

621.376.23 3145
Lock-In Amplifiers for Signals Buried in Noise—R. D. Moore. (*Electronics*, vol. 35, pp. 40–43; June, 1962.) Signals 40 db below the input noise level of a microwave receiver can be recovered by the lock-in technique in which a phase-sensitive detector is followed by a low-pass filter. Circuit details and examples of other applications are given.

621.391.82:523.164 3146
A Note on Radio Interference at a Frequency of 408 Mc/s—I. I. K. Paulin-Toth and J. R. Shakeshaft. (*Electronic Engrg.*, vol. 34, p. 488; July, 1962.) "Measurements have been made of the net level of interference due to broad-band sources at a distance of 8 km from a large town. It is concluded that at such a distance they do not constitute a serious hazard to observations with low noise systems."

STATIONS AND COMMUNICATION SYSTEMS

621.376.2/.3:621.391.832.4 3147
Nonlinear Distortion and Noise in A.M. and F.M. Systems—E. G. Woschni. (*Nachrtech.*, vol. 12, pp. 121–124; April, 1962.) General and approximate relations between modulation level, noise and distortion are derived. Selectivity in FM systems is also considered.

621.391 3148
Investigation of the Errors Arising in the Evaluation of Statistical Signals—G. Schweizer. (*Arch. elekt. Übertragung*, vol. 16, pp. 235–244; May, 1962.) The sources of error are reviewed and formulas are derived for estimating the magnitude of the errors. The application of Fourier analysis and correlation methods in the determination of signal power spectra is investigated.

621.391:621.376.5 3149
The Technique of Instantaneous-Value Companding of Pulse-Amplitude-Modulated Signals—W. Glaser. (*Nachrtech.*, vol. 12, pp. 131–138; April, 1962.) The requirements to be met by a pcm compander system are discussed with reference to the results of distortion measurements in a compander. See also 4296 of 1961.

621.396.029.65 3150
Millimetre-Wave Transmission by Non-waveguide Means—R. G. Fellers. (*Microwave J.*, vol. 5, pp. 80–86; May, 1962.) Techniques for free-space transmission at mm λ are described. Experimental results are presented indicating that the system attenuation can be much lower than that of conventional waveguide.

621.396.216 3151
A Method for Obtaining Compatible Single-Side-Band Modulation—T. J. van Kessel, F. L. H. M. Stumpers, and J. M. A. Uyen. (*E.B.U.*, pp. 12–19; February, 1962.) The system described is based on squaring full-carrier SSB signal and suppressing, by filtering, all components except those around the new carrier frequency which is twice the original. With multitone modulation the intermodulation terms in the envelope are eliminated by limiting and subsequent amplitude modulation with the original AF signal. The system is claimed to be free from distortion, and undesired components are at a lower level than in comparable and more intricate systems. [*e.g.*, 290 of January (Kakh)].

621.396.43:551.507.362.2:621.397 3152
Transcontinental Satellite Television Transmission—D. H. Hamilton, Jr., H. H. Hoover, Jr., R. V. Locke, Jr., D. C. MacLellan, W. E. Morrow, Jr., B. E. Nichols, T. F. Rogers, and P. Waldron. (See 3160.)

621.396.43:621.372.8 3153
The Problem of Long-Distance Communications in the mm-Wavelength Range by means of the H_{01} Waveguide Mode—U. Frühauf. (*Nachrtech.*, vol. 12, pp. 125–130; April, 1962.) The requirements for long-distance waveguide transmission systems are reviewed and related measurement techniques are outlined.

621.396.43:621.391.827 3154
Interference between Satellite Communication Systems and Common Carrier Surface Systems—H. E. Curtis. (*Bell Sys. Tech. J.*, vol. 41, pp. 921–943; May, 1962.) The potential interference between Telstar-type equipment and TD-2 and TH systems is examined quantitatively.

621.396.712.3:621.397.612 3155
Group of Papers on the Ernst Becker Studio Centre of the Südwestfunk in Baden-Baden—(*Rundfunktech. Mitt.*, vol. 6, pp. 53–93; April, 1962.) Nine papers giving details of the architectural design, sound and television installations and ancillary services of the new center.

621.396.946 3156
The Quest for Reliable Earth-Space Communications—H. T. Hayes. (*J. Brit. IRE*, vol. 23, pp. 481–488; June, 1962.) Problems encountered during design and test phases of several satellite communication systems are discussed. Remedial action for correcting unforeseen deficiencies is considered.

SUBSIDIARY APPARATUS

621.3.087.4:681.142 3157
Time Compression with a Solid-State Analogue Memory—J. W. Gratian. (*Electronics*, vol. 35.) pp. 88–92; May, 1962. Equipment for the spectrum analysis of speech signals is described in which information is stored in analog form using a transfluxor matrix. Analog-to-digital conversion and mechanical-wear points such as slip-ring assemblies and contact between head and storage media are eliminated. A limitation is the signal/noise ratio of 15 db.

621.311.69:621.383.5 3158
Forward Characteristics and Efficiencies of Silicon Solar Cells—J. H. Queisser. (*Solid-State Electronics*, vol. 5, pp. 1–10; January/February, 1962.) Commercial solar cells contain localized defective areas which lower the over-all efficiency.

TELEVISION AND PHOTOTELEGRAPHY

621.397:061.3 3159
The Second International Television Symposium, Montreux—C. B. Bavill. (*J. Brit. IRE*, vol. 23, pp. 489–493; June, 1962.) An account of the proceedings of the conference held in Switzerland, April 24–May 4, 1962.

621.397:621.396.43:551.507.362.2 3160
Transcontinental Satellite Television Transmission—D. H. Hamilton, Jr., H. H. Hoover, Jr., R. V. Locke, Jr., D. C. MacLellan, W. E. Morrow, Jr., B. E. Nichols, T. F. Rogers, and P. Waldron. (Proc. IRE, vol. 50, pt. 1, pp. 1522–1523; June , 1962.) An example is shown of a transmission made via Echo I satellite.

621.397.13 3161
Narrow-Band T.V. uses Pseudo-Random Scan—S. Deutsch. (*Electronics*, vol. 35, pp. 49–51; April, 1962.) A system for closed-circuit television over telephone lines with a 10-Mc bandwidth is described. It requires a long-persistence screen giving 10 per cent light output 2.5 sec after excitation; the frame period is 2.667 sec.

621.397.132 3162
F.M. Sequential Signals Kill Multicoloured Ghosts—P. Cassagne and M. Sauvanet. (*Electronics*, vol. 35, pp. 50–52; May, 1962.) A modification of the SECAM system is described (see *e.g.*, 3660 of 1960) in which intermodulation effects and transmission distortion are reduced by using FM for the color subcarrier. Circuit details are given.

621.397.132 3163
Experimental Transmissions of Colour Television over an International Long-Distance Link between Darmstadt, Berne and Rome—K. Bernath, F. Cappuccini and J. Müller. (*Tech. Mitt. PTT*, vol. 40, pp. 127–133; April, 1962. In French.) The test results indicate that color television transmission using the modified N.T.S.C. system is feasible over a 1500-km Eurovision link incorporating modern equipment from different sources. For German version see *Nachrtech Z.*, vol. 15, pp. 161–166; April, 1962. For Italian version see *Note Recensioni Notiz.*, vol. 11, pp. 27–40; January-April, 1962.

621.397.2:621.391.827 3164
Common-Channel Interference between Television Signals of the Same or Different Standards—G. F. Newell and E. W. Taylor. (*E.B.U.*, no. 72A, pp. 56–67; April, 1962.) Report on measurements of the ratios of the amplitude of wanted and interference signals which produce a given subjective grade of interference. The television standards investigated are 405, 625 and 819 lines with 50 fields/s. A comparison of the protection ratios required at close offset for 525-line, 60-fields/s and 625-line, 50-fields/s signals is included.

621.397.62:621.391.827.25 3165
Observation of Adjacent-Channel Interference in the Television Band III in Domestic Receivers—H. Brand. (*Tech. Mitt. PTT*, vol. 40, pp. 87–97; March, 1962.) Tests were carried out on several standard-type domestic receivers to assess the effect on picture quality of the reception of unwanted signals from transmitters differing by one, two or three channel numbers from the wanted signal.

621.397.621:621.382.3 3166
Transistorized Television Receivers—(*J. Brit. IRE*, vol. 23, pp. 431–458; June, 1962.)
a) Tuners for Transistorized Television Receivers—J. K. Brown (pp. 431–433.)
b) Transistorized IF Amplifiers for Television Receivers—E. G. Chardin (pp. 433–437.)
c) Line Time-Base Circuits for Transistor Operation—R. W. A. Scarr (pp. 438–439.)
d) Some Transistor Requirements for A.G.C. Operation—J. R. James (pp. 440–443.)
e) Transistor Video Amplifier and Line Time-Base Synchronization Circuits for Television Receivers—M. C. Gander and P. L. Mothersole (pp. 445–458.)

621.397.7–182.3 3167
Mobile Television Outside-Broadcast Units—B. Gensous. (*E.B.U.*, no. 71A, pp. 20–25; February, 1962.) New light-weight equipment used by the Radiodiffusion-Télévision Française is described with examples of its use on motorcycles, in helicopters and on parachutists. For previous description of outside-broadcast techniques of the R.T.F. see 2397 of 1961.

621.397.712 3168
CBS-KNXT Hollywood Television Broadcasting Centre—R. S. O'Brien, R. B. Monroe, W. B. Whalley, and A. P. Evans—(*J. Soc. Mot. Pic. Telev. Engrs.*, vol. 71, pp. 251–265; April, 1962.) An illustrated description of layout and equipment, which includes a digital computer for automatic control of program switching, at a new broadcasting center in Los Angeles, U.S.A.

TRANSMISSION

621.396.712:681.84.087.7 3169
The Transmitter End of the F.C.C. Stereophony Multiplex System for V.H.F. F.M. Broadcasting—A. Ruhrmann. (*Nachrtech. Z.*, vol. 15, pp. 113–118; March, 1962.) The operational characteristics and tolerances of the U. S. standard system for stereophony broadcasts [see, *e.g.*, 1044 of March (Wilhelm)] are examined. The effects of amplitude and phase distortion are estimated.

621.396.74 3170
Decoupling Bridge Circuits for Parallel Connection of Transmitters in the Short-Wave and V.H.F. Bands: Part 2—Increased Bandwidth—W. Buschbeck and W. Burkhardtsmaier. (*Nachrtech. Z.*, vol. 15, pp. 181–189; April, 1962.) Modifications are described, relating to the bridge circuits given in Part 1 [1723 of May (Buschbeck)], for improving bandwidth and input impedance characteristics.

TUBES AND THERMIONICS

621.382 3171
Tunnel Emission into Vacuum—J. Cohen. (*J. Appl. Phys.*, vol. 33, pp. 1999–2000; June, 1962.) Al-alumina-Al and Al-alumina-Pt thin-film sandwiches were used as tunnel emitters. An increase of emission by five orders of magnitude was obtained by virtue of a Cs film on the Pt surface. See also 2798 of 1961 (Mead).

621.382 3172
The "Sogicon"—a New Type of Semiconductor Oscillator—M. Kikuchi and Y. Abe. (*J. Phys. Soc. Japan*, vol. 17, pp. 881–882; May, 1962.) A section of a *p*-type high-resistivity bar is etched to a small cross section to constrict a current flowing in it, producing a higher electric field in this region. Coherent oscillation in the current can be obtained by applying rectangular voltage pulses along the bar. A theoretical explanation is not given.

621.382 3173
The Space-Charge-Limited Dielectric Triode—G. T. Wright. (*Solid-State Electronics*, vol. 5, pp. 117–126; May/June, 1962.) The device can be operated at much higher frequency than transistors, with lower noise figure and improved temperature characteristic.

621.382:539.23 3174
Thin-Film Triode Research—G. T. Advani, J. G. Gottling, and M. S. Osman. (Proc. IRE, vol. 50, pt. 1, pp. 1530–1531; June, 1962.) A summary of the performance of an all-evaporated thin-film triode is given. Modifications to the simple analysis are required to explain the results.

621.382:621.372.44 3175
Semiconductor Surface Varactor—R. Lindner. (*Bell Sys. Tech. J.*, vol. 41, pp. 803–831; May, 1962.) This varactor device using surface space charge is analyzed and measurements made on several experimental units are described. Theory developed from the surface charge relation agrees with experimental data over a wide range of Si resistivity.

621.382.22:621.372.622 3176
Crystal-Diode Admittance—W. J. Lucas. (*Electronic Tech.*, vol. 39, pp. 261–263; July, 1962.) The computed curve reproduces the average measured admittance curve well over the frequency range 500–18,000 Mc.

621.382.23 3177
Junction Diodes made on Plastically Deformed Germanium—P. L. Parsons. (*Brit. J. Appl. Phys.*, vol. 13, pp. 367–368; July, 1962.) "Dislocations have been introduced into *n*-type Ge bars by plastic twisting at 850°C. The Ge resistivity was little affected, but In alloy diodes made on this material showed reduced hole storage effects, in quantitative agreement with theory."

621.382.23 3178
Analysis of a Tunnel Diode—A. Shibata. (*J. Phys. Soc. Japan*, vol. 17, pp. 770–778; May, 1962.) The constant in Esaki's equation (1784 of 1958) is determined for direct tunneling current in a narrow *p-n* junction. The calculated variations of current with temperature and bias are in good agreement with experimental results.

621.382.23 3179
Extending Tunnel-Diode Operating Frequency—K. Ishii and C. C. Hoffins. (*Electronics*, vol. 35, pp. 43–45; June, 1962.) By using a special waveguide transformer and mounting, operating frequencies can be raised to 100 Gc.

621.382.23 3180
Microwave Applications of the Silver-Bonded Diode—S. Kita. (*Electronics*, vol. 35, pp. 86–87; May, 1962.) The diode has a cut-off frequency above 130 Gc and capacitance of 0.3 pF at zero bias.

621.382.23 3181
Current/Voltage Characteristics of Germanium Tunnel Diodes—M. I. Nathan. (*J. Appl. Phys.*, vol. 33, pp. 1460–1469; April,

1962.) A report of measurements on tunnel diodes with phonon-assisted or non-phonon current predominant.

621.382.23 3182
Aging Effects in GaAs Tunnel Diodes—H. J. Henkel. (*Z. Naturforsch.* vol. 17a, pp. 358–360; April, 1962.) The causes of the irreversible decrease in the tunnel current of GaAs tunnel diodes operating under load in the forward direction are investigated. Experimental results indicate that this may be due to changes in donor and acceptor concentration, in this instance Sn and Zn.

621.382.23:538.63 3183
Magnetic-Field Effects on InSb Tunnel Diodes—L. Esaki and R. R. Haering. (*J. Appl. Phys.*, vol. 33, pp. 2106–2109; June, 1962.) The ratio of longitudinal to transverse tunnel currents of InSb tunnel diodes in a magnetic field is used for the measurement of the magnitude and voltage dependence of the tunneling exponential.

621.382.23:546.289 3184
Degenerate Germanium: Part 1—Tunnel, Excess, and Thermal Current in Tunnel Diodes—D. Meyerhafer, G. A. Brown, and H. S. Sommers, Jr. (*Phys. Rev.*, vol. 126, pp. 1329–1341; May, 1962.) Experimental results are presented for a series of measurements of the forward admittance of Ge tunnel diodes of known impurity concentration and from 4.2° to 300°K. Implications of these results were studied in Part 2: 950 of March (Sommers).

621.382.23:621.372.44 3185
Characteristics of an Exponentially Retrograded Variable Capacitance Diode: Parts 1 & 2—H. C. Nathanson and A. G. Jordan. (*Semiconductor Prod.*, vol. 5, pp. 38–42 and 24–28; April and May, 1962.) The device exhibits strongly nonlinear behavior of the type $C=aA(V+Vo)^{-\alpha}$ where α is of the order of 3. Nonlinear modes of operation are possible without stringent junction breakdown limitations.

621.382.23:621.373.51 3186
The Tunnel Diode as Oscillator—E. Klein. (*Nachrtech. Z.*, vol. 15, pp. 135–143; March, 1962.) A cubical parabola is used as an approximation to the tunnel-diode characteristics, and the operation of a tunnel diode in a parallel-tuned oscillator circuit is analyzed on this basis.

621.382.3 3187
Flux Analysis of a Transistor Structure—T. P. Brody. (PROC. IRE, vol. 50, pt. 1, pp. 1524–1525; June, 1962.) The "flux analysis" method [see 2346 of July (McKelvey)] is used to investigate the added-carrier transport problem in a one-dimensional semiconductor.

621.382.3 3188
A Metal-Base Transistor—D. V. Geppert. (PROC. IRE, vol. 50, pt. 1, pp. 1527–1528; June, 1962.) A metal-base transistor is described which uses an ordinary thermionic emitter. Test results indicate that it could compete with conventional *p-n-p* and *n-p-n* structures.

621.382.3 3189
A Hot-Electron Transistor—D. Kahng. (PROC. IRE, vol. 50, pt. 1, p. 1534; June, 1962.) Results obtained using experimental transistors with Au base, Hg emitter and Si collector are consistent with a hot-electron transport mechanism.

621.382.3:539.23 3190
The T.F.T.—a New Thin-Film Transistor—P. K. Weimer. (PROC. IRE, vol. 50, pt. 1, pp. 1462–1469; June, 1962.) Operation of the device depends on the control of injected majority carriers in a wide-band-gap semiconductor by an insulated gate. Characteristics of experimental units are given.

621.382.3:621.372.44 3191
Semiconductor Inductive Elements: Parts 1 & 2—H. G. Dill. (*Semiconductor Prod.*, vol. 5, pp. 30–33 and 28–31; April and May, 1962.) A transistor with earthed collector is operated in the α cutoff region; it has a moderate temperature sensitivity which could be partly compensated. A simple band-pass circuit is described.

621.382.3:621.375.9:621.372.44 3192
Pushing Transistors above their Frequency Limits with Parametric Conversion Operation—U. L. Rohde. (*Electronics*, vol. 35, pp. 46–49; June, 1962.) Circuits and techniques for using available transistors at UHF are described. Test figures for several types of transistor, and performance data for an experimental tuner for the range 400 Mc 5 Gc are given.

621.382.322 3193
The Mark II Spacistor—J. M. Lavine. (*Solid-State Electronics*, vol. 5, pp. 39–48; January/February, 1962.) Conductances of the order of 500 and 2500 $\mu\Omega^{-1}$, and power gains of 100 and 500 have been measured for electron flow and hole flow, respectively.

621.382.323 3194
Circuit Applications of the Field-Effect Transistor Parts 1 and 2—T. B. Martin. (*Semiconductor Prod.*, vol. 5, pp. 33–39 and 30–38; February and March, 1962.) The device equations are presented in simple form using only externally measurable parameters and the static characteristics are discussed. Applications of the transistor as an amplifier, an electronically variable resistance and an analog multiplier are described.

621.382.333 3195
Power Gain of Transistors at High Frequencies—J. Lindmayer. (*Solid-State Electronics*, vol. 5, pp. 171–175; May/June, 1962.) Calculations show that in general for junction transistors the power gain will fall by 6 db/octave and that f_{max} can be related to the common-emitter current-gain cutoff frequency.

621.383.032.217.2 3196
Enhanced Photoemission and Photovoltaic Effects in Semitransparent Cs_3Pb Photocathodes—F. Wooten. (*J. Appl. Phys.*, vol. 33, pp. 2110–2112; June, 1962.)

621.383.29 3197
Photomultiplier Detection of Microwave Modulated Light—O. L. Gaddy and D. F. Holshouser. (PROC. IRE, vol. 50, pt. 1, p. 1525; June, 1962.) A description of the application of a photomultiplier based on the dynamic crossed-field electron multiplication principle (1750 of May).

621.383.5 3198
Photoeffect in Au-CdS and Cu-CdS Photovoltaic Cells—E. D. Fabricius. (*J. Appl. Phys.*, vol. 33, pp. 1597–1603; April, 1962.) Experimental results are presented which indicate the origin of the photovoltaic current in Cu-CdS and Au-CdS photocells.

621.385:621.382.2 3199
Reverse-Biased *p-n* Junctions in Electron Tubes—F. A. White and J. C. Sheffield. (PROC. IRE, vol. 50, pt. 1, pp. 1523–1524; June, 1962.) A note on suggested "semitron" devices based on charge multiplication in a *p-n* junction incorporated as an anode or target in an electron tube.

621.385.032.21:537.29:621.374.4 3200
Microwave Devices with Field-Emission Cathodes—J. R. Fontana and H. J. Shaw. (*Commun. and Electronics*, no. 59, pp. 43–48; March, 1962.) The advantages and the design of field-emission diodes for use at microwave frequencies are discussed with particular reference to harmonic generators. See 3297 of 1958.

621.385.032.213 3201
Activation of Thermionic and Photoelectric Cathodes—D. F. Klemperer. (*J. Appl. Phys.*, vol. 33, pp. 1532–1536; April, 1962.) Clean surfaces are produced when certain oxidizer metals and semiconductors are heated in a vacuum, suggesting that the oxygen penetrates to the subsurface. This process leads to enhancement of thermionic and photoelectric emission from the surface.

621.385.032.213 3202
Emission Characteristics of a Philips Impregnated Cathode—R. F. Hill and S. R. Rouze. (*J. Appl. Phys.*, vol. 33, pp. 1607–1608; April, 1962.) A description is given of the changes in surface appearance as a function of temperature.

621.385.032.213 3203
Diffusion of Magnesium in Base Nickels for Oxide-Coated Cathodes—D. A. Campbell. (*J. Appl. Phys.*, vol. 33, pp. 1609–1610; April, 1962.) An expression is derived which shows how Mg in the form of an intermetallic compound affects the diffusion of Mg in the Mg-Ni base into the oxide coating.

621.385.032.24 3204
Synthesis of Control-Grid Structure for Remote-Cut Off Tubes from the Transfer Characteristic—K. Venkataramanan. (*J. Inst. Telecomm. Engrs., India*, vol. 8, pp. 42–49; January, 1962.) Instead of using a conventional grid of var able pitch the tube is made of three electrode assemblies in one envelope. Each assembly has a constant grid pitch and its characteristic conforms to part of the desired characteristics curve for the complete tube.

621.385.1.004.6 3205
The Behaviour of the Emission-Dependent Characteristics of Thermionic Valves during an Extended Period of Operation—W. Kirschke. (*Nachrtech.*, vol. 12, pp. 87–90; March, 1962.) Discussion of the results of measurements on long-life tubes with particular regard to the formation of cathode interface layers.

621.385.2 3206
The Calculation of the Admittance of Space-Change-Limited Diodes in Accordance with Isothermal Theory—K. Richter and G. Sagl. (*Arch. elekt. Übertragung*, vol. 16, pp. 230–234; May, 1962.) A simple method of calculation is described which is based on the theory discussed by Pötzl and Richter (1681 of 1961). A comparison with experimental results indicates closer agreement than can be achieved on the basis of the uniform-velocity theory.

621.385.6 3207
R.F. Current Behaviour in Electron Beams with D.C. Velocity Spread—T. G. Mihran. (*J. Appl. Phys.*, vol. 33, pp. 1582–1590; April, 1962.) The observed damping rate of the RF current is not explained by a distribution function analysis, which therefore may not be valid for the problem of noise in microwave tubes.

621.385.6.029.65 3208
New Approaches to Millimetre-Wavelength Devices—W. W. Teich. (*Electronics*, vol. 35, pp. 37–44; May, 1962.) A survey of

designs of electron-beam tubes for the generation and amplification of mm and sub-mm waves, covering developments in cathode construction, cooling and miniaturization techniques, and field geometry.

621.385.62 3209

High-Power Microwave Source uses Parallel Klystrons and Flywheel—L. D. Shergalis. (*Electronics*, vol. 35, pp. 46–49; June, 1962.) A power amplifier is described which operates at C-band frequencies. Single or parallel operation of up to four klystrons is possible, feeding a modified ring resonator to provide up to 100 Mw output.

621.385.623.5 3210

Long-Term Frequency Stability for a Reflex Klystron without the Use of External Cavities—G. B. Gucker. (*Bell Sys. Tech. J.*, vol. 41, pp. 945–958; May, 1962.) A description of a klystron with frequency stability within 50 parts in 10^9 over an ambient temperature range of 100°C.

621.385.623.5 3211

Honeycomb Grids for Klystrons—J. Kotowski, J. Muschaweck, and L. Fritz. (*Frequenz*, vol. 16, pp. 69–76; March, 1962.) A method of manufacturing honeycomb grids of Cu is described that avoids the structural irregularities to which grids made by existing methods are prone. From an analysis of thermal behavior and useful life under heavy load conditions special grid configurations are derived which combine a high degree of transparency with long life.

621.385.632:621.318.2 3212

A Low-Leakage Periodic Magnetic Focusing System for Travelling-Wave Tubes—W. Henne. (*Elektron. Rundschau*, vol. 16, pp. 156–158; April, 1962.) The system consists of an array of radially magnetized rings inside an iron cylinder acting as shield and magnetic return path. The effects of coupling devices and anode aperture on the field configuration, and the stray field produced by the system are discussed with reference to measurements. See also 1778 of May.

621.387:621.362 3213

The Effect of Staging on Thermionic Converters—Hatsopoulos. (See 3130.)

621.387:621.362 3214

Electron Space-Charge-Limited Operation of Cesium Thermionic Converters—Carabateas. (See 3131.)

621.387.032.213 3215

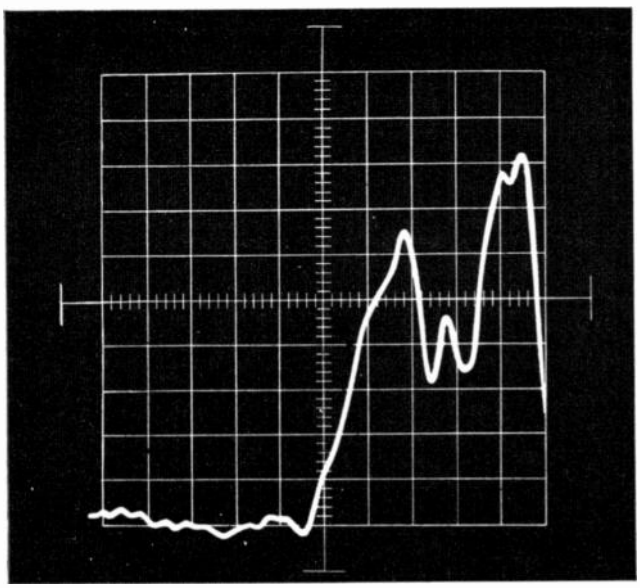
Thermionic Emission of Molybdenum in Vapours of Cesium and Cesium Fluorides—R. L. Aamodt, L. J. Brown, and B. D. Nichols. (*J. Appl. Phys.*, vol. 33, pp. 2080–2085; June, 1962.) A detailed experimental investigation of Mo cathodes, which may be useful in thermionic converters.

621.387.3 3216

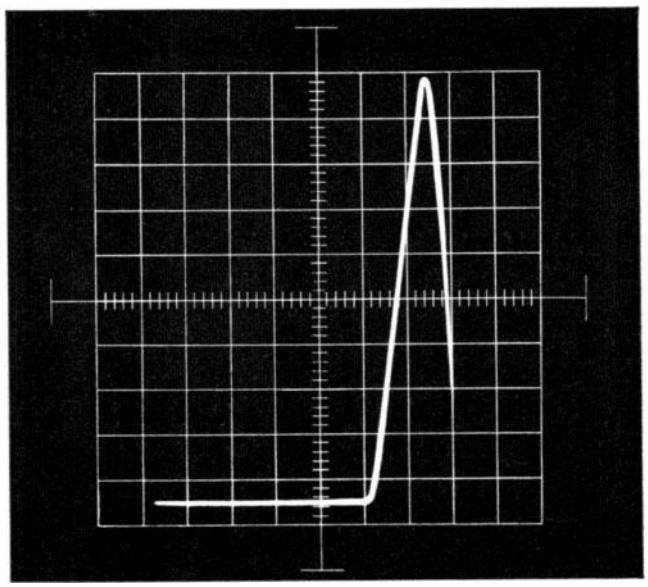
The Effect of Current and Pressure on the Impedance of Cold-Cathode Glow-Discharge Tubes—G. F. Weston. (*J. Electronics and Control*, vol. 12, pp. 379–392; May, 1962.) Simple expressions are found for the impedance in terms of "self-inductance" and "degression time."

IBM asks basic questions in mechanical analysis

What is the best design?



This is an example of the force-time relationship of impact printing as the striking head of an output printing device comes into contact with its platen.



After modifications were made in the printing mechanism, the force-time relationship reflected a cleaner impact, producing a sharper impression from the printer.

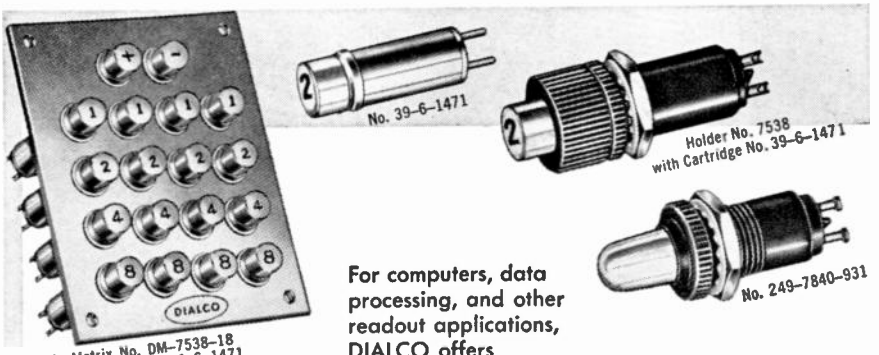
As computers go faster, so must the machines which work with them. The tremendous speed at which these mechanisms function magnifies traditional problems of wear, elasticity, and timing. To meet rigorous new standards of performance, IBM engineers are applying the techniques of mechanical analysis to the development of high-speed mechanisms.

Through a combination of experimental measurements and mathematical analyses, IBM engineers attempt to determine exactly what is taking place at any given instant during the machine cycle. For example, our engineers recently were given the assignment of increasing the operating speed of the output printing element in a data processing system by as much as twenty-five percent without major redesign. To do so, they progressively varied mass, spring forces, and the elasticity of several components, and measured resulting changes in the system's physical characteristics. By translating these experimental results into mathematical terms, they were able to arrive at a solution to the problem which might have been impossible to obtain using only trial-and-error methods.

The exact description of the way in which many variables interact during the machine cycle requires advanced measurement and analysis techniques. For example, the impact curves shown above were produced by a mechanical structure with several degrees of freedom—a system extremely complicated to calculate. IBM engineers chose to alter one structural member slightly to produce an overriding, dominant frequency. They could then analyze the structure's operation mathematically.

Mechanical analysis is helping IBM engineers to create mechanisms with fundamentally new designs. For instance, in a revolutionary new typewriter, the IBM Selectric,™ they were able to design a mechanism to compensate automatically for wear and elasticity in the print-selection system. In an important new development in disk storage, they used a hydraulic actuator to position magnetic read-write heads which float on air bearings. And they have developed a high-speed printer, controlled by a computer's program, which prints by means of engraved type suspended on a chain which moves past the paper at high speed. Mechanisms with advanced designs like these are helping to make it possible to take full advantage of the immense capabilities of the computer.

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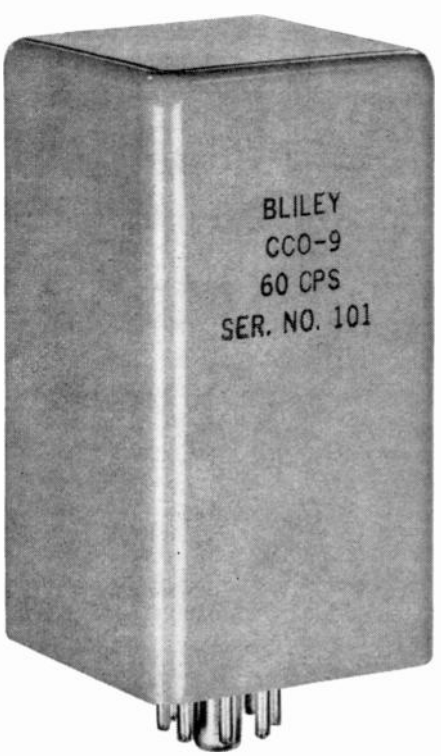
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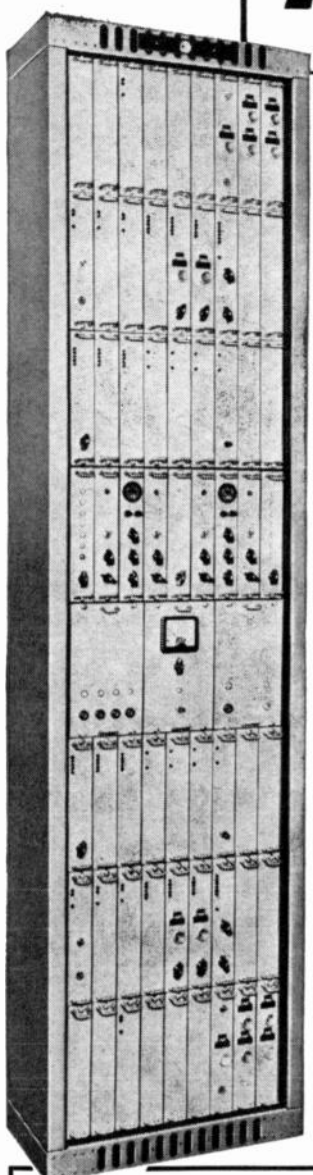
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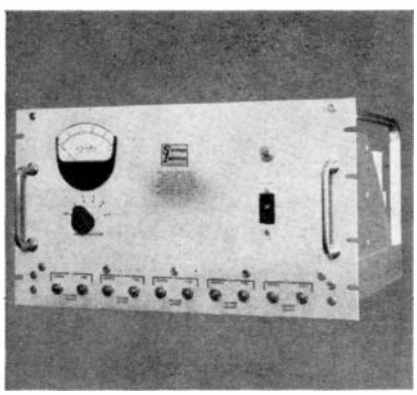
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Kravitz, R. N., Philadelphia, Pa.
Larsen, K., Los Angeles, Calif.
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Leach, W. F., Melrose, Mass.
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Manning, D. B., Manhattan Beach, Calif.
Mark, W. D., Bedford, Mass.
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McLaughlin, T. M., Dorchester, Mass.
Mertens, D. E., St. Louis, Mo.
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Mukerjee, T., North Syracuse, N. Y.
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Rhinehart, J. H., Arlington, Va.
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Robillard, R. A., Cambridge, Ohio
Rosen, J., Edison, N. J.
Rosse, T. A., Brookline, Mass.
Rothstein, A. A., Lexington, Mass.
Ruppel, R. F., Commack, L. I., N. Y.
Rusher, D. L., Anaheim, Calif.
Schiecke, K. L., Chicago, Ill.
Schneider, S. J., Overland, Mo.
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(Continued on page 90A)

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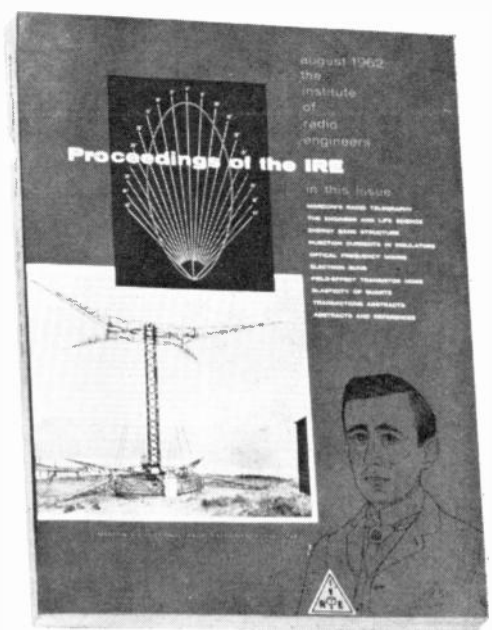
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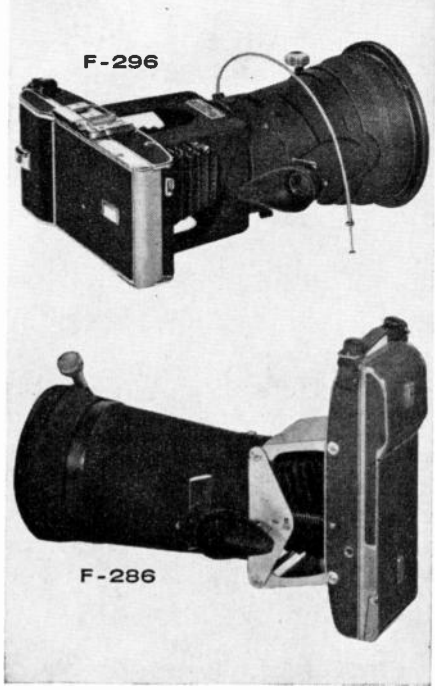
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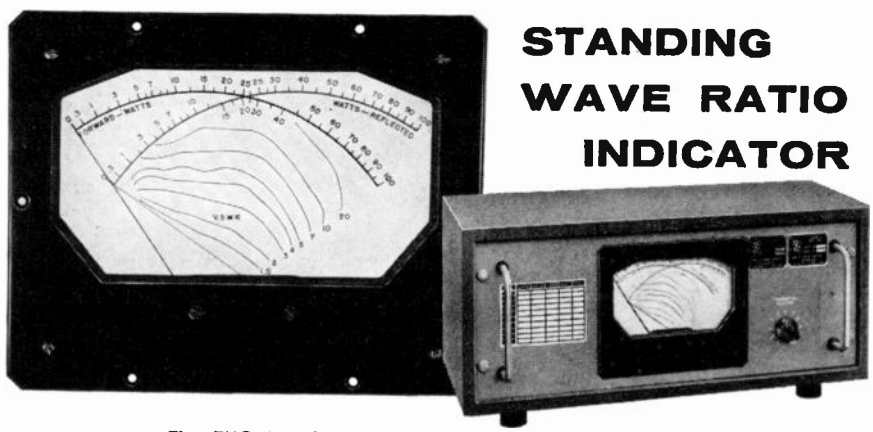
(Continued from page 88A)

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Shanley, T. J. B., Norristown, Pa.
Shelton, F. A., Jr., Dover, N. J.
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Ziegel, W. S., Los Angeles, Calif.
Ziegler, R. M., Arcadia, Calif.

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Benson, P. J., Boston, Mass.
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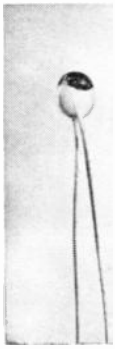
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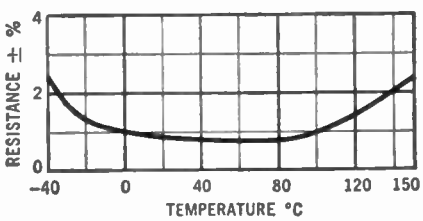
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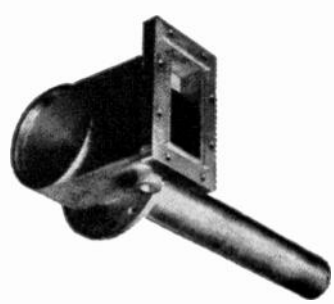
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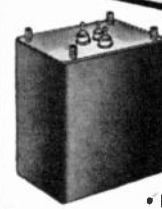
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IRE People

(Continued from page 64A)



Lee C. Barrett (A'47) has been appointed to the new post of Manager, Systems Manufacturing at Frequency Engineering Laboratories, producer of microwave equipment. He was formerly Project Engineer in the firm's Systems Engineering Group. In his new post, he will be responsible for the facilities planning, organization and production of the AN/WLR-1 Countermeasures Receiver Set and other systems equipment.

❖



Joseph Behr (A'44), Vice President of Finance, has been elected to the post of Executive Vice President at Radio Engineering Laboratories, Inc. (REL), Long Island City, N. Y. REL is the communications subsidiary of the Dynamics Corporation of America in New York.

He joined the Company to coordinate wartime Goverment Contracts in 1943, became Secretary in 1947 and a Director in 1953. He was named a Vice President in September, 1954. Prior to joining REL he was with the Asiatic Petroleum Corporation (Dutch Shell's Purchasing Subsidiary) from 1925 to 1943, and held the post of Senior Buyer there from 1931.

Mr. Behr is a graduate of Stuyvesant High School in New York. He is a member of the Radio Club of America and the Armed Forces Communications and Electronics Association.

❖





R. B. Bonney (A'41–M'44–SM'45) has been elected Executive Vice President of the Electronic Engineering Company of California, and **W. R. McQuiston** (A'48–M'51–SM'54) has been elected Vice President, Sales of the same company.

Mr. Bonney is a Director and Cofounder of the Company. He is also a Director of Siliconix, Inc., the EECO-Baldwin Piano Company joint venture for the manufacture of integrated molecular electronic circuits. He has served as Vice President since the company's incorporation in 1951. He is directly responsible for the technical management of EECO and directs the operations of the engineering department. Prior to his association with EECO he was employed in an engineering management position with the Crosley Corporation, where he supervised radar and IFF equipment design and development projects. He received the B.S. degree in electrical engineering from the University of New Mexico, and is a registered professional engineer in California and Ohio.

In his new position, Mr. McQuiston will direct all marketing operations nationally and internationally for EECO. He is a Corporation Director since 1951, and has been sales Manager since 1955. Before that he served as Division Manager of EECO's Point Mugu division at the Naval Missile Test Center. He joined the company in 1949 as a project engineer in the Point Mugu division, and later became technical staff engineer. Prior to joining EECO he was a research analyst with the Boeing Company. He is a graduate of the University of Washington.

❖

Commander John J. Dougherty, (S'52–SM'61) USN, was presented the Assistant Chief of the Bureau of Ships for Research and Development Special Honorary Award for Scientific Achievement for 1962. This award is granted annually by the Navy Department to the individual, civilian or military, making the most outstanding contribution in the field of research and development for the Bureau of Ships.

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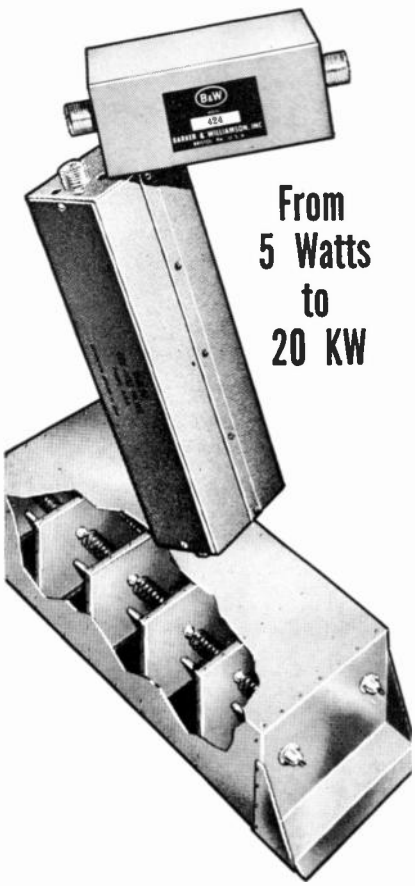
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Commander Dougherty was graduated from the United States Naval Academy in 1945, and the U. S. Naval Postgraduate School, Monterey, Calif. in 1953, with the M.S. degree in engineering electronics.

❖



Richard S. Bowers (SM'49) has been named Executive Vice President at Ortronix, Inc., Orlando, Fla. He was formerly Vice President of Engineering. In his new position he will be responsible for the overall operation of the company's three divisions.

He has specialized in installation, flight test and evaluation of various navigational radar systems, analog computers and instrumentation systems. At General Precision Laboratory, he had responsibility for the installation and flight test of prototype doppler radars for evaluation at Wright Air Development Center. Subsequent to this he had system responsibility for the integration of doppler radar into bombing and guidance systems under development in the United Kingdom. Prior to the formation of Ortronix, he had prime responsibility for an airborne FM/FM telemetry system for ICBM nose cone reentry studies at Radiation, Inc. In addition he had project responsibility for an automatic digital data recording and processing system for use on Project Vanguard.

Mr. Bowers graduated from Williams College in 1950. He is a member of the Army Ordnance Association and the Armed Forces Communications and Electronics Association.

❖



Edwin J. Bradley (A'52) has been appointed to the position of Manager of Marketing of Diodes, Inc., Canoga Park, Calif. He has previously been General Sales Manager of Hoffman Electronics Corporation, Semiconductor Division, and most recently Western Regional Sales Manager with Sperry Rand, Semiconductor Division.

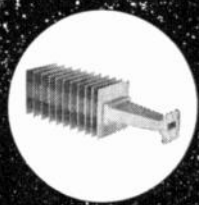
Mr. Bradley is a member of the American Radio Relay League.

❖

(Continued on page 104A)

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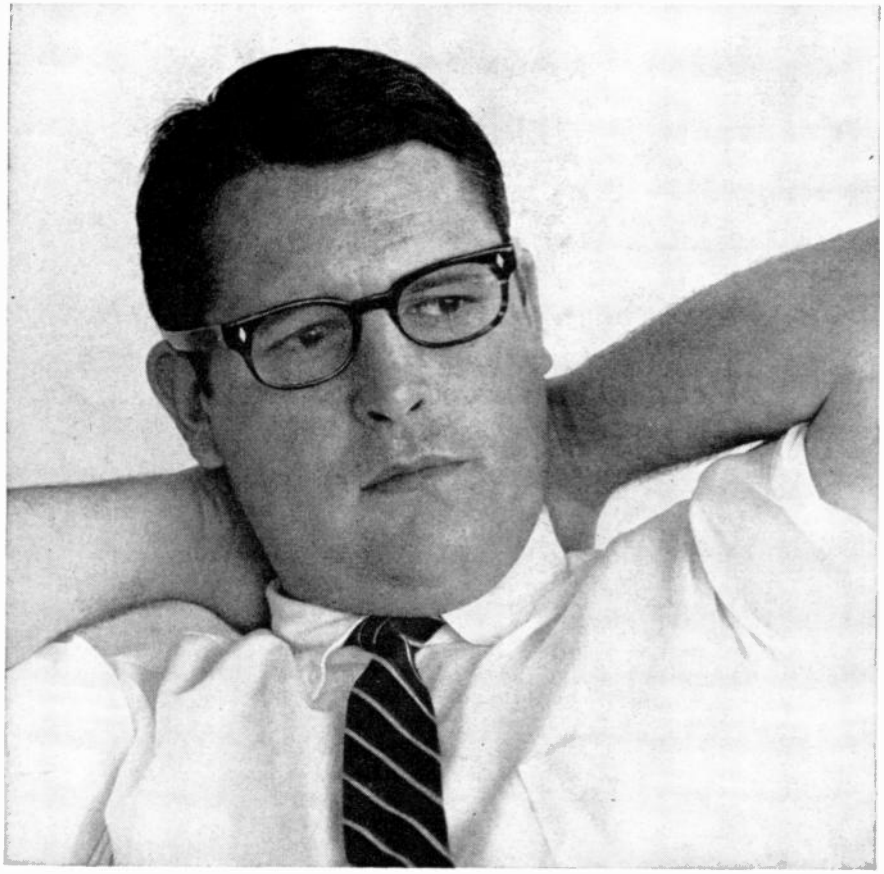




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Computer Programming
Data Processing
Celestial Mechanics

PHYSICS
Semiconductors
Cryogenics
Magnetohydrodynamics
Thermionics
Plasma
Optical Instruments
Spectroscopy
Infra-red
Molecular-Atomic
Atmospheric

JET PROPULSION LABORATORY
CALIFORNIA INSTITUTE OF TECHNOLOGY
"An equal opportunity employer."
Jet Propulsion Laboratory is operated by California Institute of Technology for the National Aeronautics & Space Administration

Positions Open

The following positions of interest to IRE members have been reported as open. Apply in writing, addressing reply to company mentioned or to Box No.

The Institute reserves the right to refuse any announcement without giving a reason for the refusal.

Proceedings of the IRE
1 East 79th St., New York 21, N.Y.

SALESMAN

To sell semi-technical instrument in eastern Pennsylvania and Delaware for manufacturers' representative on straight commission (no draw, no expenses; we handle the product on that basis). Degree not needed. Age is immaterial so long as you possess health and integrity. Write Box 2075.

ASSOCIATE PROFESSOR ELECTRICAL ENGINEERING

Ph.D. degree required. Salary $7500 to $8000 per academic year. Opportunity to develop research activity in area of interest. Address replies to: Head, Electrical Engineering Department, South Dakota School of Mines and Technology, Rapid City, South Dakota.

Ph.D. ELECTRICAL ENGINEERING

Unusual opportunity in U.S. Civil Service to recent Ph.D. in Electrical Engineering to continue his research or teaching while administering Army-sponsored basic research projects at U.S. universities, colleges and institutes. Air-conditioned offices located on the Duke University campus immediately adjacent to the Engineering College and Physics-Mathematics Departments.

Associates on the Engineering Sciences Staff active in individual research projects at Duke University campus very attractive in a progressive area of the South. Good family environment. Civil Service levels begin at GS-11 ($7560) to GS-14 ($12,210), depending on background qualifications. Write details of background or submit a Form 57 (obtainable at any post office) to J. J. Murray, Army Research Office (Durham), Box CM, Duke Station, Durham, North Carolina.

ELECTRONIC ENGINEERS

Electronic Engineers for permanent positions with Federal Communications Commission, Washington, D.C., GS-12, $8955 to GS-13 $10,635. Must be graduate engineers with good knowledge of problems involved in the area of space communications. Government will pay expenses for transportation of employee, his family, and household goods to Washington. Attractive fringe benefits include retirement, life and health insurance, automatic pay increases. Good opportunity to enter government career service. Reply to Personnel Officer, Federal Communications Commission, Washington 25, D.C.

LIFE SCIENCE ENGINEERS

The human factor as a design detail is as old as engineering; as a studied design factor it is relatively new. Engineers with training or experience in the biological or behavioral sciences who are interested in analysis and experimentation in the areas of terrain-following and other radar displays, photointerpretation, advance vehicle handling qualities and heuristic programming techniques are invited to reply to: F. P. Rentschler, Cornell Aeronautical Laboratory, Inc., Buffalo 21, New York.

RESEARCH ASSOCIATE or ASSISTANT PROFESSOR

Immediate opening for professional-level applicant as Research Associate or Assistant Professor to supervise designing, fabricating, and maintaining specialized electronic equipment for research in science and engineering. A properly qualified person will, if he desires, have the opportunity to participate in electronic research or teach. A staff member on full-time appointment may enroll for one course each semester. Salary in range of $7500 to $8500. Annual appointment with one month vacation. Application or further inquiry should be made to: Dean Virgil W. Adkisson, Research Coordinator, University of Arkansas, Fayetteville, Arkansas.

ELECTRONIC ENGINEER

Electronic Engineer, B.S., recent graduate with a few years of practical experience in electronics or communications, for design and field engineering work. Projects will involve instrumentation for studies of the auroral ionosphere, radio-wave propagation, and geomagnetism. Salary in the $900-$1000 per month range. Write: Geophysical Institute, University of Alaska, College, Alaska.

ELECTRICAL DESIGN ENGINEERS

Experienced in video-amplifiers and pulse circuitry design. Familiarity with solid-state devices and transistorized circuitry would be desirable. Send resume to: George H. Rimbach, Supervisor of Personnel, HRB-Singer, Inc., Science Park, P.O. Box 60, State College, Pa. Qualified applicants will hold at least a four-year degree in science or engineering with three to five years related experience.

MECHANICAL DESIGN ENGINEERS

Experienced in military electromechanical assemblies and high-speed rotational devices. Send resume to: George H. Rimbach, Supervisor of Personnel, HRB-Singer, Inc., Science Park, P.O. Box 60, State College, Pa. Qualified applicants will hold at least a four-year degree in science or engineering with three to five years related experience.

RELIABILITY ENGINEERS

Knowledge of reliability theory and experience in military specifications, prediction techniques, and evaluation and testing methods. Send resume to: George H. Rimbach, Supervisor of Personnel, HRB-Singer, Inc., Science Park, P.O. Box 60, State College, Pa. Qualified applicants will hold at least a four-year degree in science or engineering with three to five years related experience.

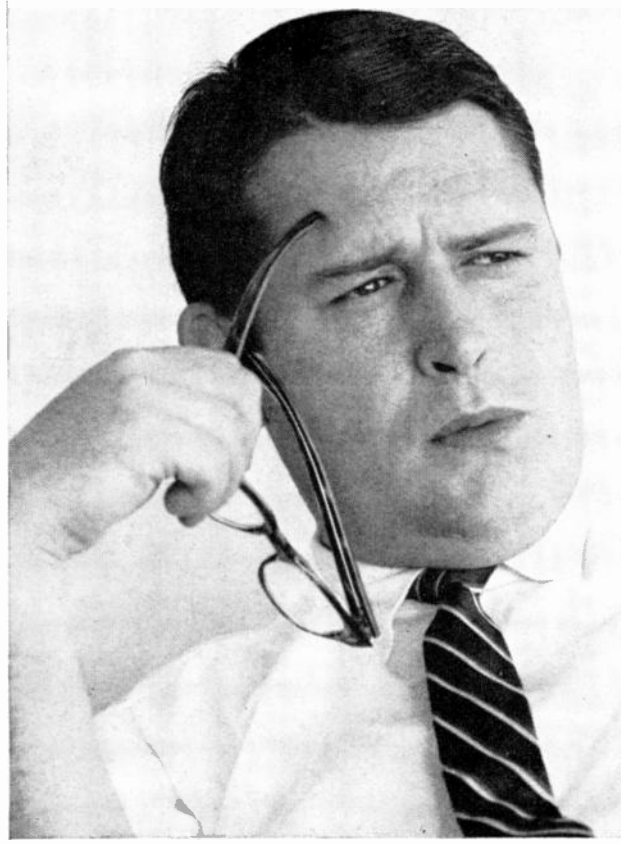
SUPERVISORY TECHNICAL PUBLICATIONS EDITOR

Incumbent will be the Assistant Technical Publications Officer for the Naval Aviation Engineering Service Unit. He will supervise the publication of the *Digest of U. S. Naval Aviation Electronics* and other technical publications. All manuscript copy will be submitted to him for review and editing. He will supervise a group of engineers engaged as writers and a clerical force and instruct a continuing changing group of Naval officers in the art of technical writing. GS-12—$8955 per annum. Send resume to: Industrial Relations Department, Naval Air Material Center, Philadelphia 12, Pa.

TEACHING POSITION

Ph.D. in Electrical Engineering needed for fall quarter. Undergraduate and master's programs. Nine months contract with summer employment readily available in area. Salary good and depends on rank and experience. New facilities. Electronics or communications fields preferable. Address replies to: Chairman, Department of Electrical Engineering, Seattle University, Seattle 22, Washington.

(Continued on page 98A)



Think harder.

Put your head to work on such as these: What's the moon made of? Is there life on other planets? What's hiding behind the clouds of Venus? What you think may be very important to other thinkers at JPL. Now's a good time to learn if JPL can use your talent. Use the short-form below, or if you've already made a long-form resume, send it today to JPL.

TELL US ABOUT YOURSELF

name_____

address_____

city_____ state_____ phone_____

education and training:

degree	subject	school	date rec'd
BS			
MS			
PhD			

currently employed_____

how long on present job_____ salary_____

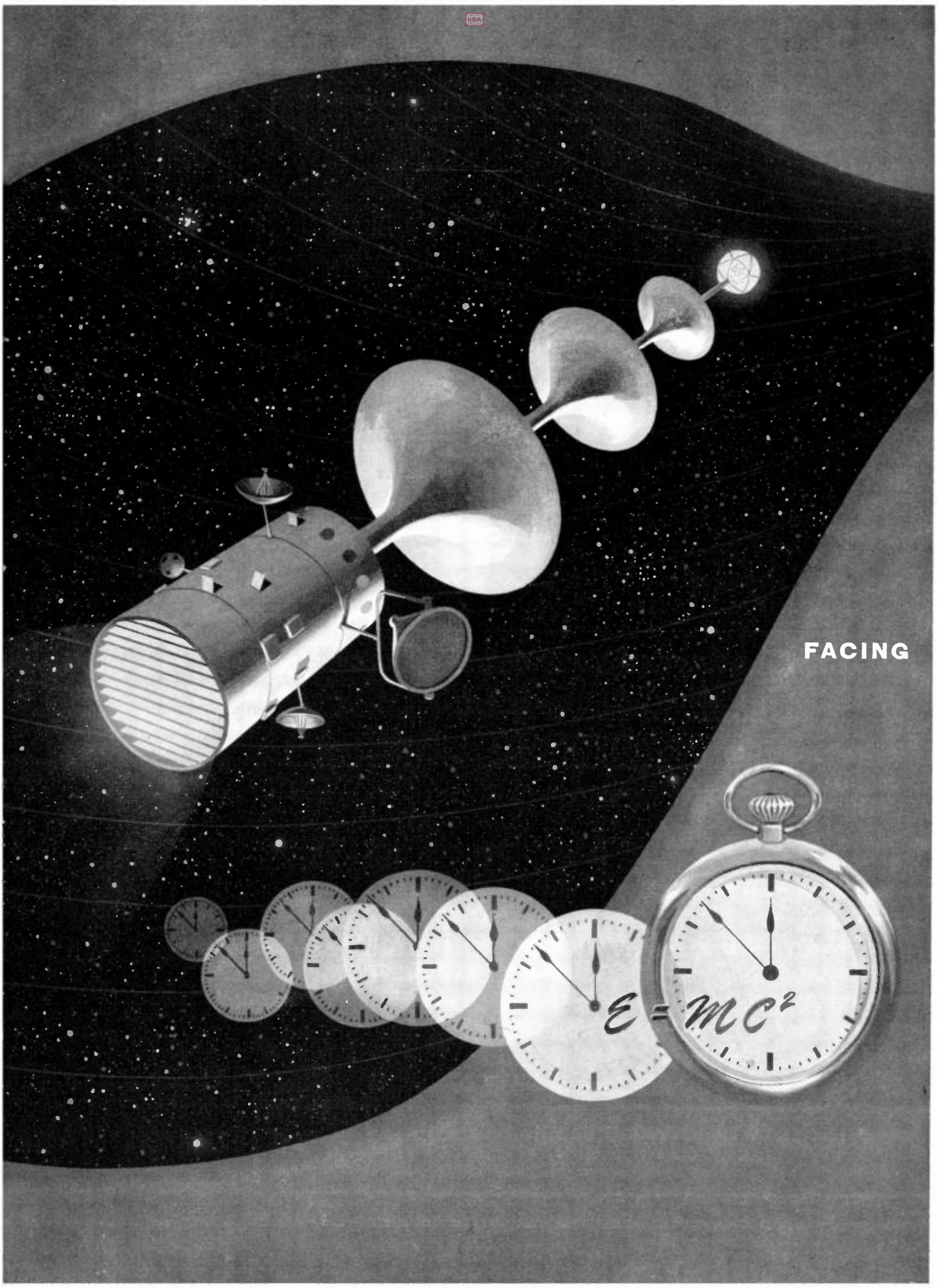
current position_____

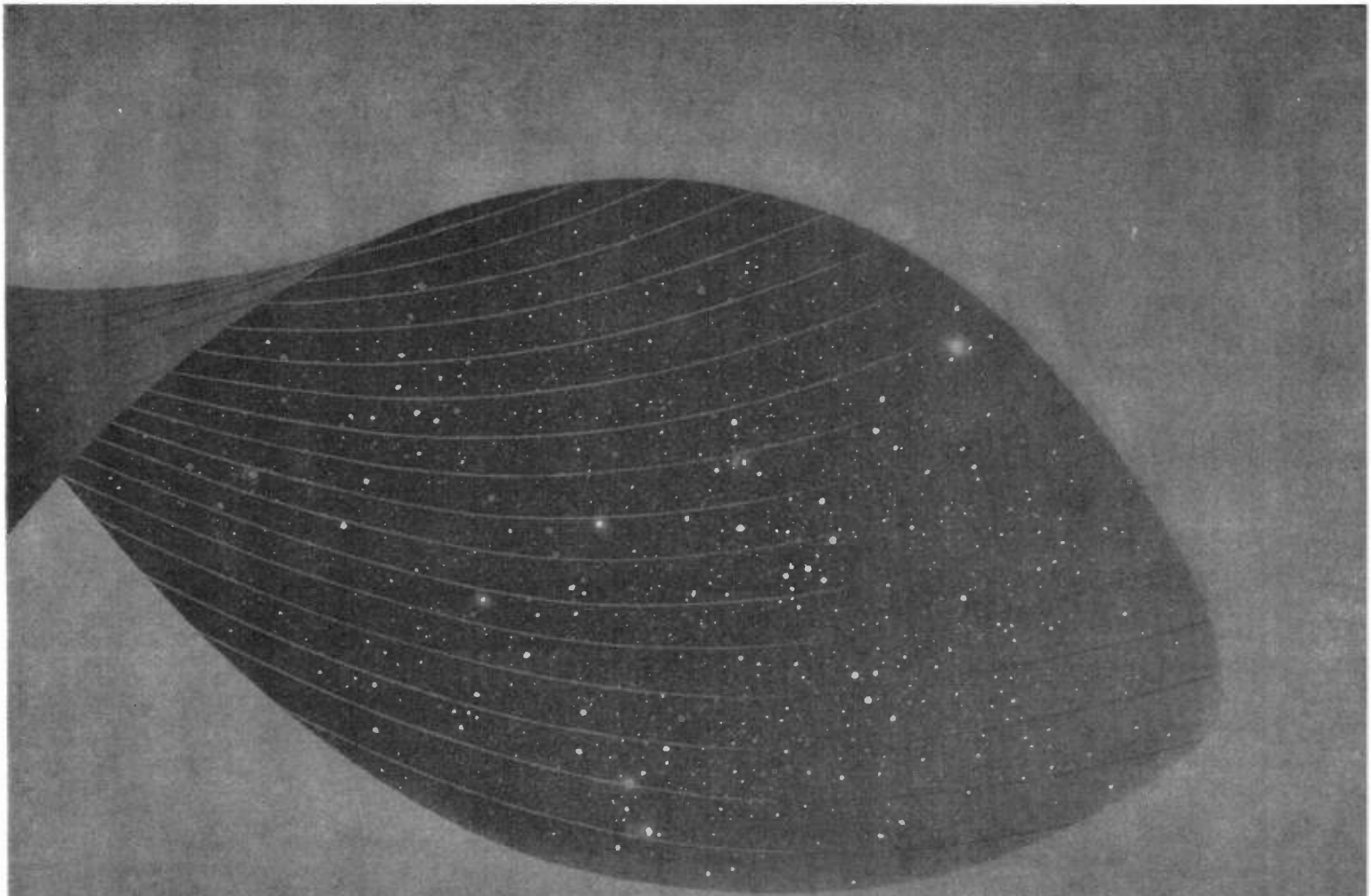
primary specialty_____

"An equal opportunity employer." • U.S. citizenship or current security clearance required.

JET PROPULSION LABORATORY
CALIFORNIA INSTITUTE OF TECHNOLOGY
4804 Oak Grove Drive, Pasadena, California
attention: Personnel Department

jpl


FACING
E = MC²



THE FOURTH DIMENSION IN PROPULSION DEVELOPMENT

Whether the universe has a "saddle shape," or any shape at all, is a matter of interesting conjecture. The matter of space travel, however, is the subject of intense experimentation. A nuclear/thermionic/ionic propulsion system, currently being studied at Lockheed Missiles & Space Company, might well become the power source for space vehicles.

Its design incorporates a nuclear reactor only one foot in diameter, generating heat at a temperature of 1850°K. This is transmitted to banks of thermionic generators, converting the heat directly into electrical energy for the ion beam motor which uses cesium vapor as a fuel. The entire system is designed without any moving parts, minimizing the possibility of failure.

Lockheed's investigation of propulsion covers a number of potential systems. They include: plasma, ionic, nuclear, unique concepts in chemical systems involving high-energy solid and liquid propellents, combined solid-liquid chemical systems. The fundamentals of magnetohydrodynamics, as they might eventually apply to propulsion systems, are also being examined. Just as thoroughly, Lockheed probes all missile and space disciplines in depth. The extensive facilities of the research and development laboratories—together with the opportunity of working with men who are acknowledged leaders in their fields—make association with Lockheed truly rewarding and satisfying.

Lockheed Missiles and Space Company in Sunnyvale and Palo Alto, on the beautiful San Francisco Peninsula, is an exciting and challenging place to work. For further information, write Research and Development Staff, Department M-24A, 599 North Mathilda Avenue, Sunnyvale, California. An Equal Opportunity Employer.

LOCKHEED *MISSILES & SPACE COMPANY*

A GROUP DIVISION OF LOCKHEED AIRCRAFT CORPORATION

Systems Manager for the Navy POLARIS FBM and the AGENA vehicle in various Air Force Satellite programs. Other current projects include such NASA programs as the OGO, ECHO, and NIMBUS.

SUNNYVALE, PALO ALTO, VAN NUYS, SANTA CRUZ, SANTA MARIA, CALIFORNIA • CAPE CANAVERAL, FLORIDA • HAWAII

WRH

PHOENIX
GOODYEAR AIRCRAFT

(Arizona Division)

offers career opportunities that will challenge your talent in the technical areas listed below.

SR. DEV. ENG'RS.

Develop wideband i-f amplifiers, i-f oscillators, i-f isolation amplifiers, and related circuits.

Develop microwave circuits and equipment including wave-guide assemblies, and related circuits.

Servomechanisms. Develop and design of antenna stabilization servo systems. Film drive control systems.

Develop radar transmitters and modulators including pulse forming networks, transformers, protection circuits, X-band amplifiers and related circuits.

Display and pulse circuits. Develop display circuits and controls. Bomb navigation cross hair circuitry.

SR. DESIGN ENG'RS.

Optical design and development. Design lenses and complete optical systems.

SR. PACKAGING ENG'RS.

Layout packaging and detailing layout of electronic sub-assemblies including i-f and r-f units.

Request Application or Send Resume to:
M. J. McColgan, Engineering Personnel
Goodyear Aircraft Corporation
Litchfield Park, Arizona

An equal opportunity employer.

Similar positions at Goodyear Aircraft Corporation, Akron, Ohio.

Positions Open

(Continued from page 95A)

ELECTRONICS OR RADIO ENGINEER

Rapidly growing Manufacturing Company, employing 160, located near Philadelphia, requires individual with minimum of 3 years experience in microwave design and measurements to head research and development in the application of coaxial cable to high frequency equipment: B.Sc. in physics or E.E. minimum requirement, Masters desirable. Send resume and salary requirements to Box 2076.

ELECTRICAL PROFESSORS

Ph.D.'s required for newly merged state university. Emphasis on graduate teaching and research. Initial opportunities in control engineering, sampled-data systems, and advanced network theory. Excellent consulting and liaison with research-minded local industry. Outstanding salaries and fringe benefits. Industry-based applicants are invited. Apply Chairman of Electrical Engineering, State University of New York at Buffalo, 3435 Main Street, Buffalo 14, New York.

SENIOR ELECTRICAL AND ELECTRONIC ENGINEERS

Challenging opportunities in expanding independent research institute located in cultural center area adjacent to universities and one of the nation's outstanding technical and scientific libraries. Surrounded by pleasant residential districts and shopping centers. Positions for B.S., M.S., and Ph.D. engineers with backgrounds and extensive experience in antennas, electromagnetic wave propagation, microwave theory and techniques, or power distribution networks and analysis. A high degree of freedom to conduct theoretical and experimental research work in an excellent technical and scientific environment. Please send your resume to: Mr. Harold L. Stout, Director of Engineering, Midwest Research Institute, 425 Volker Boulevard, Kansas City 10, Missouri. An equal opportunity employer.

CONTROLS RESEARCH ENGINEER

Electrical engineer with Ph.D. and experience in either control theory, electrical circuitry, or solid state devices needed for technical direction of active program. This is a unique attempt to apply latest concepts to commercial and industrial products on a system basis. Send resume to Box 2078.

ELECTRONIC ENGINEER

ENRICO FERMI INSTITUTE FOR NUCLEAR STUDIES has a position available for a B.S. or M.S.E.E. with a minimum of 2 years experience in the design and development of transistorized circuitry. We offer the opportunity to work with a small group of dedicated scientists engaged in basic physics research. Salary open, send resume in complete confidence. William R. Hutchins, UNIVERSITY OF CHICAGO, 956 E. 58th Street, Chicago 37, Illinois.

ASSISTANT COMMUNICATIONS ENGINEER

Large Class 1 Railroad, Opportunity for an engineer with ideas and initiative to grow with an expanding organization and assume substantial management functions. Report to head of Department and be responsible to design and specification of system communications, including development of policies, directing activities of assistant engineers, supervisors and technicians. Graduate

(Continued on page 100A)

NCR

RESEARCH ENGINEERING

PROFESSIONAL PERSONNEL

If your potential is not being recognized, consider the excellent growth atmosphere at NCR:

- Extensive Company backed R & D
- New facilities: 40% expansion in building for R & D (complete—1963)
- Advanced programs encompassing major sciences integrated for practical results
- Excellent employee benefits
- An aggressive management group
- Professional advancement encouraged through Company sponsored programs

We have openings at this time as listed below. All are for our Dayton, Ohio, R & D Center:

SENIOR CIRCUIT DESIGNER:
Familiar with solid-state circuitry; creative; desire to work at advanced development stage.*

THIN FILM:
Materials research; device development.*

DATA REDUCTION ENGINEERS/MATHEMATICIANS:
Large systems background required; project leader experience; (military).

CHEMISTS—RESEARCH:
Project leader capability; desire background and interest in physical chemistry; polymers.

TEST EQUIPMENT DESIGN ENGINEERS: BS OR MSEE

MECHANICAL DESIGN ENGINEERS:
Small mechanism experience desired.

ADVANCED DEVELOPMENT

PLANNING SPECIALISTS*

OPERATIONS RESEARCH*

INTEGRATED ELECTRONICS*

*These positions are not limited to any special level. Ph.D. background is preferred for many, and some areas of responsibility involve Management or Technical Director qualifications.

Send your personal letter to: T. F. Wade, Technical Placement
The National Cash Register Company, Dayton 9, Ohio
An equal opportunity employer



...not a bad sentiment for a job, either

In May of this year, a company-wide program was instituted by General Electric to sharpen every employee's awareness of the extra values that should be built into every G-E product — whether it's a clock-radio or a continent-spanning command and control system.

But there's no reason why this program should be limited to products. For years General Electric, and its Defense Systems Department, have placed the "Accent on Value" on professional positions as well. For example:

ACCENT ON PROGRAM DIVERSIFICATION — To briefly identify some of the program areas — **AWCS 412-L,** now in advanced development phase, will span entire continents. Its potential is global air weapons control. **MISTRAM** is a revolutionary precision trajectory measurement system with important ramifications in space vehicle terminal guidance. **NUDETS (477-L)** will be a national network of nuclear detection and measurement devices. **MISSILE AND SPACE VEHICLE GUIDANCE** is the logical extension of our engineers' pioneering work in developing the ATLAS Radio-Command Guidance System, currently used for booster guidance on the majority of America's space shots, including Project Mercury.

ACCENT ON TECHNICAL SUPPORT — The tools to get the job done. Among them, a computer facility of 7 general-purpose machines, ranging in size, speed and capability from modest to the fastest, large-scale equipment commercially available. This is augmented further by DSD's specially-designed GEESE facility — a custom mating of special-purpose digital equipment with an analog capacity of 550 amplifiers.

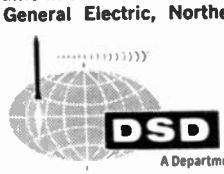
ACCENT ON PROFESSIONAL DEVELOPMENT — Full tuition refund for graduate work at Syracuse University. (27% of our engineers and scientists currently participating.) Educational leaves of absence granted to qualified individuals. We also conduct over 100 in-plant professional courses, lasting from 13 to 26 weeks. DSD works closely with the professional societies, supporting workshops and seminars, encouraging active participation and attendance by our professional people.

ACCENT ON PERSONAL REWARDS — DSD's salaries are fully competitive at every professional level. What's more, a clearly-defined salary range brackets each level of responsibility. Based on individual performance, your compensation can increase as much as one-third, within your current position description. And of course, promotion to broader responsibilities raises the upper limit of your salary accordingly.

IMMEDIATE OPENINGS — This list is somewhat long, a reflection of DSD's continuing expansion into new areas of technological promise. If you can identify yourself by title or subject, write us today. We'll reply promptly, and attempt to find the specific position that best matches your immediate interests and long-term goals.

Communication Systems / Applied Mathematics / Materials Engineering / Microminiature Electronic Packaging / Semiconductor Circuits Design / Computer Systems Applications / Electronic Liaison-Production Engineering / Microminiature Mechanical Design / Operations Analysis / Systems Equipment Analysis / Telecommunication Systems Design / Project Integration Engineering / Equipment Evaluation

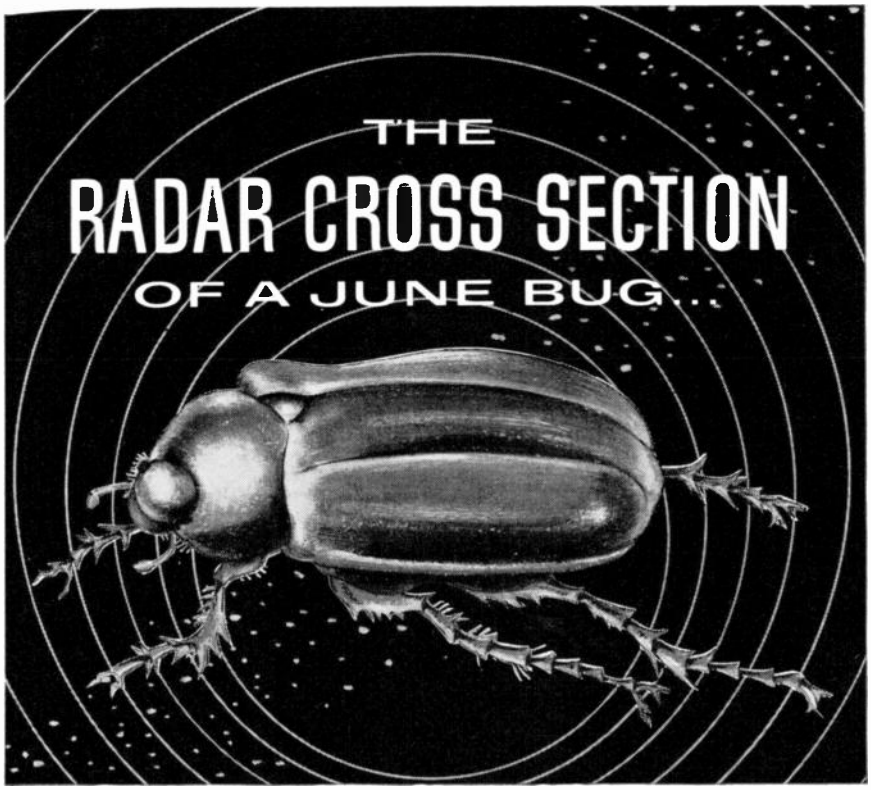
Forward your resume in strict confidence to Mr. P. W. Christos, Div. 53-MJ, Defense Systems Dept., General Electric, Northern Lights Office Bldg., Syracuse, N. Y.



DSD DEFENSE SYSTEMS DEPARTMENT
A Department of the Defense Electronics Division

GENERAL ELECTRIC

An Equal Opportunity Employer





. . . is 10^{-5} sq. meters. At least, those that sometimes interfered with our significant research in the measurement of radar cross sections were that big.

Meaningful research, however, must surmount more serious barriers than June bugs. Barriers, for example, like those of semantic differences between the tongues of electronics engineers and physicists, between aeronautical engineers and psychologists. This is particularly true in a research laboratory, where professional men of vastly different backgrounds work with and try to understand each other as they push back the frontiers in new fields.

Here, scientists and engineers with minds open to new ideas, adept at stepping over barriers between technologies, will find a broad array of problems to solve — in an environment that encourages the development of original ideas. Here, engineers conducting plasma sheath research work shoulder to shoulder with scientists doing research in ion cyclotron resonance, or in re-entry interactions with the upper atmosphere, or in gaseous electronics.

Cornell Aeronautical Laboratory welcomes atmospheric and ionospheric physicists, and engineers experienced in electronics, microwave techniques, optics, and physical chemistry as principal investigators. We invite men with *ideas* to mail the coupon today.

CORNELL AERONAUTICAL LABORATORY, INC.
of Cornell University

J. P. Rentschler J-B
CORNELL AERONAUTICAL LABORATORY, INC.
BUFFALO 21, NEW YORK

Send me a copy of your factual, illustrated employment prospectus, "A Community of Science."

Name_____

Street_____

City_____ Zone_____ State_____

☐ Please include employment information. An Equal Opportunity Employer

Positions Open

(Continued from page 98A)

to age 35. Experience in VHF or microwave radio, plus wire line carrier, telephone and data communications. Railroad communications experience desirable. Location mid-west. Salary commensurate with experience. Send resume, recent photo and salary requirements with first letter. All replies confidential. Send resume to Box 2079.

CRITICAL VACANCIES AT GRIFFISS AIR FORCE BASE, ROME, NEW YORK

Employee Development Officer	GS-11	$7560 pa
Medical Officer (Occupational, Health & Medicine)	GS-12	$8955 pa
Mechanical Engineer (2)	GS-12	$8955 pa

Wage Board Positions

#Radar Repairer	W-8	$2.41 ph
#Ground Radio Installer	W-8	$2.41 ph
#Ground Radio Installer	W-11	$2.66 ph

#—Positions require 75% travel along Eastern Seaboard & Canada.

Positions Wanted

By Armed Forces Veterans

In order to give a reasonably equal opportunity to all applicants and to avoid overcrowding of the corresponding column, the following rules have been adopted:

The IRE publishes free of charge notices of positions wanted by IRE members who are now in the Service or have received an honorable discharge. Such notices should not have more than five lines. They may be inserted only after a lapse of one month or more following a previous insertion and the maximum number of insertions is three per year. The IRE necessarily reserves the right to decline any announcement without assignment of reason.

Address replies to box number indicated, c/o IRE, 1 East 79th St., New York 21, N.Y.

FREQUENCY AND INTERFERENCE SPECIALIST

Retiring officer, 25 years radio & communication experience civil & USAF, desires full or part time position in warm climate. Extensive experience international frequency allocation, and management of operation and maintenance functions. Available Fall of 1962. Write Box 4000 W.

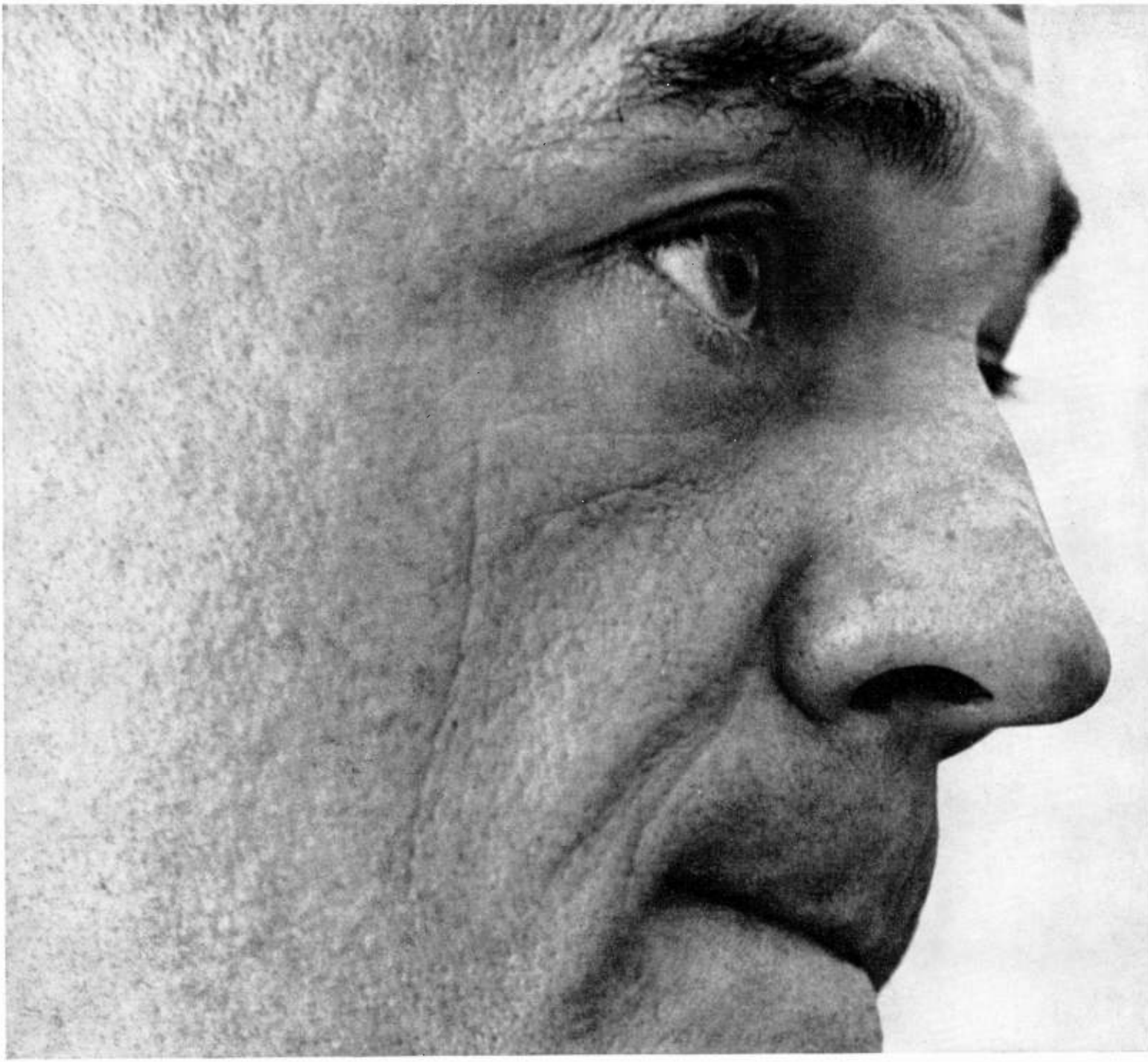
TECHNICAL PUBLICATIONS ENGINEER

Presently Director of Technical Publications and Advertising. Seeks challenging position with increased scope and responsibility. Experience includes ten years of radar and microwave technology and three years in advertising and public relations. Write Box 4001 W.

BIOLOGICAL INSTRUMENTATION

B.S.E.E.—1951. Proficient in micro volt level amplifier design, feedback system analysis, active filter design, and digital circuit design. Desires to work on problems concerned with the life process or on the development of devices for supplementing man's sensory perception. Location on West Coast desirable. Write Box 4002 W.

(Continued on page 102A)



The best men are often failures.

Their thinking doesn't stop with the tried-and-true. They like going out on technological limbs, exploring the unexplored, working in areas where a breakthrough may be preceded by many blind alleys.

Are you this breed of man? If so, you'll like Northrop. You'll especially like the unconstrained atmosphere where you're free to try the untried. We are currently at work on more than 70 active projects, and we're constantly evaluating new lines of inquiry. Projects range from space guidance and navigation to automatic checkout equipment, from computer design and world-wide communications to laminar flow control.

On the following pages you'll find some specific positions available now at Northrop Space Laboratories. Look them over. One may be just the spot for you.

But even if you don't find your specialty listed — don't go away. We simply don't have room to mention all the opportunities to be found throughout Northrop's several divisions. If you're the kind of man who likes to investigate new areas of technology, there's bound to be a place for you at Northrop. Write to Dr. Alexander Weir at Northrop Corporation, Box 1525, Beverly Hills, California, and tell us about yourself. You will receive a prompt reply. **NORTHROP** An equal opportunity employer

WRH

Upidiperous*

Melpar needs scientists and engineers who enjoy reaching the conclusions of applied research but who are not given to making upidiperous turns.*

Scientists and Engineers, who have the desire to participate in open-ended applied research programs are offered an opportunity to attack unexpected problems and originate new concepts in an atmosphere charged with the competence and enthusiasm of a growing organization and lively colleagues.

Emerging problems have led to openings for:

Applied Meteorologists and Seismologists to explore
- **New areas of instrumentation for nuclear detonation detection**
- **Extra terrestrial/terrestrial interaction Telluric currents**

Microwave Physicists for studies of
- **Microwave plasma interactions**
- **Microwave parametric amplification**

Maser-Laser Engineers and Physicists to
- **Investigate techniques of utilizing Masers and Lasers in electronic systems**

Electromagnetism Theorists for research in
- **Space reconnaissance and exploration**

Molecular Circuit Designers for developmental research on
- **Thin films** • **Polycrystals**
- **Monocrystals**

Systems Theory Specialists for studies of
- **Optimal control techniques**
- **Coding systems**
- **Statistical decision processes**
- **Network theory** • **Sampling systems**

Applied Mathematicians for Research in
- **Information theory**
- **Switching theory** • **Digital theory**
- **Analog to digital conversion processes**
- **Stochastic processes** • **Error analyses**

Scientists and engineers with PhD, MS or equivalent experience are invited to communicate with John Haverfield, Manager, Professional Placement, **Applied Science Division**

MELPAR INC

A SUBSIDIARY OF WESTINGHOUSE AIR BRAKE COMPANY

3342 Arlington Blvd., Falls Church, Va.

an equal opportunity employer

* A upidiperous turn is a 180° turnabout where the main direction or goal is temporarily forgotten or denied. The word upidiperous is seen regularly, in contracted form, on street signs: NO U TURNS.

Positions Wanted

By Armed Forces Veterans

(Continued from page 100A)

ELECTRONIC ENGINEER

B.S. (Appl. Phys.), married, age 36. Knowledge of German, Polish, Russian. U.S. Citizen with 12 years of experience in instrumentation research and development. Desires position as Field Engineer for systems installation and customer liaison. Would consider Central European assignment. Write Box 4003 W.

ENGINEER IN FRANCE

American, 36, MSEE, 8 years experience radar systems and 4 years experience professional audio, seeks position or representation full or part time in Paris, France. Fluent French. Write Box 4004 W.

ELECTRONICS TECHNICIAN

B.S. Physics, 1958, 4½ years industrial research in Solid State Physics. Veteran USMC, 2 years aviation electronics technicians. Married, family, 29 years of age. Seek foreign employment with managerial responsibility. Write Box 4005 W.

COMMUNICATION ENGINEER

B.S. '52. Age 31, married. 10 years in management and engineering of complex communication systems (radio, radar, teletype, telephone inside and outside plant). Valid FCC radio-telephone/telegraph licenses with radar and aircraft endorsements. Seeking position in sales or applied engineering western U.S., preferably Pacific coast. Write Box 4006 W.

FREE LANCE TECHNICAL WRITER

Self-made electronics engineer reads, writes, and speaks English fluently. Fifteen years experience writing military Tech Orders, Maintenance and Training Manuals, Space and Engineering proposals. Commission basis only. Write Box 4007 W.

FIELD SERVICE ENGINEER

Fifteen years of varied experience in communications and allied fileds desires meaningful employment in field service engineering. Would prefer long term, family-type overseas location where command of Russian and German would be an asset. Salary should be interesting, but is secondary to interest in work and satisfaction gained in getting the job done. Age 34. Presently employed in field service division of large electronics firm. Write Box 4011 W.

DATA PROCESSING

B.S. Math, additional grad Physics. Age 39, with 12 years experience in electronic data processing. Includes Management, supervision, systems analysis and programming of such applications as Spares Provisioning, Parts Lists, data communications and cost reduction programs. Thoroughly familiar with special and general purpose computer design. Desires responsible position in computer applications area. Write Box 4012 W.

TECHNICAL SALES REPRESENTATIVE

Technical representative for firm interested in servicing any electronic program connected with the Electronics Command of the Dept. of the Army located at Fort Monmouth, New Jersey. Have MS in EE, 14 years experience in the design and development and production of communications equipments. This includes management and supervision. Write Box 4017.

(Continued on page 104A)

ARA

ELECTRONICS ENGINEERS

Senior and intermediate openings are now available in new programs which require analytic and hardware experience in:

- Logic
- Circuit Design
- Instrumentation
- Automatic Controls
- Servo • Analog
- Surveillance Systems

ARA is a 250-man Research and Development firm active in the fields of space systems, electronics, bionics, physics, geophysics, meteorology, chemistry, aeronautics, and other areas of physical and engineering sciences.

For further information please contact
Mr. Ivan Samuels

ALLIED RESEARCH ASSOCIATES
Virginia Road, Concord, Massachusetts
"An equal opportunity employer"

Electrical, Electronic Engineers
OR
Physicists

PRINCETON UNIV.

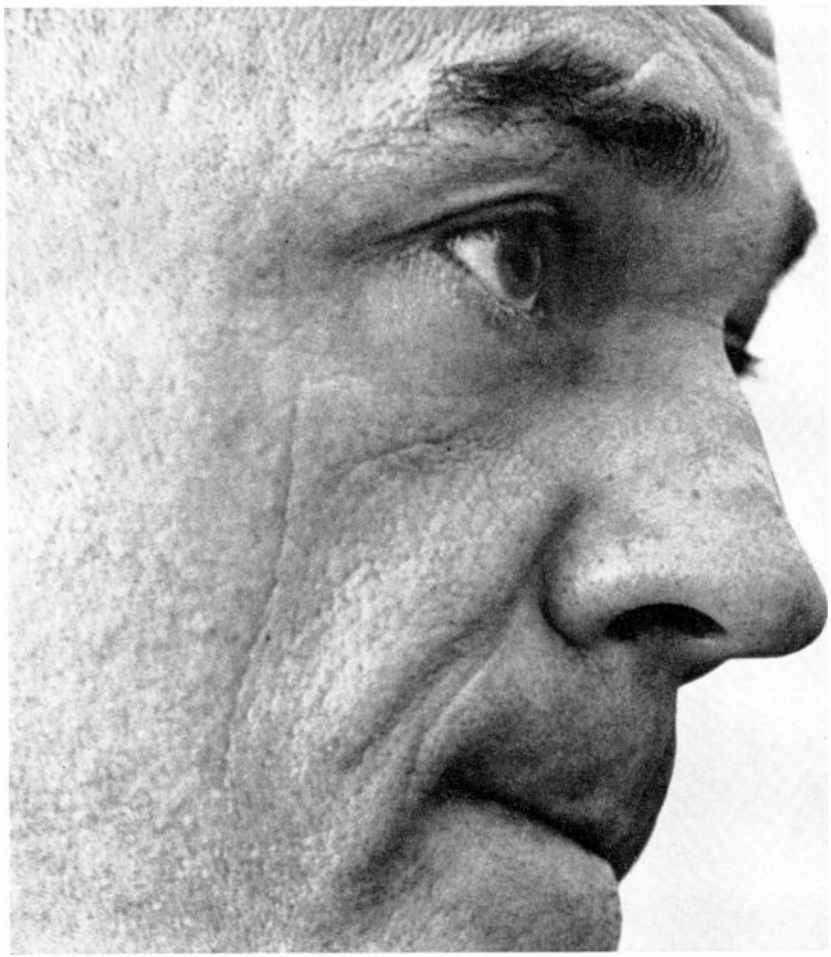
Several positions available in our design, development, operating and research areas. Bubble chamber or circuitry experience particularly helpful.

Work in suburban Princeton area. Salary comparable to industry. Unusual benefits include 4 weeks vacation and generous retirement plan.

Write to A. C. Allen

ACCELERATOR
P.O. BOX 682, PRINCETON, N.J.
An equal opportunity employer

We need men who stick their necks out.



Northrop Space Laboratories needs venturesome men. Men who aren't complacent, who constantly go beyond established thinking.

We're a new organization, and a growing one. The enthusiastic support of the entire Northrop Corporation is behind us. Those who join NSL now will grow with us, giving the directions for the years ahead. Key openings are now available for:

Research scientists, to conduct independent research on properties of surfaces with particular emphasis on stability in a simulated space environment. Also to conduct research on sealants and self-sealing and penetration resistant composite structures for the control of meteoroid damage to space vehicles.

A radiation effects physicist, to do research with emphasis on solid state devices.

A flight test engineer, who can conceive programs, estimate manpower requirements and costs.

A reliability engineer, to perform reliability analyses of space systems and subsystems in proposal and development phases.

Stress analysts, to develop fresh analytical techniques and apply them to new space structural concepts; to do stress analysis and design optimization studies on advanced space vehicle structures.

A plasma physicist, to join our growing program in the measurement of plasma properties, spectroscopy, diagnostics, accelerators, and power conversion devices.

A mathematician-physicist, to concentrate on systems analysis and operations research applied to military and non-military space systems.

Physicists experienced in electro-optical imaging devices and laser theory; engineering mathematicians interested in detection theory; reconnaissance and tracking; electronic engineers who know their way around statistical communications theory and noise phenomena; for new and original work in satellite detection systems.

For more information about these and other opportunities, write to W. E. Propst, Space Personnel Office, 1111 East Broadway, Hawthorne, California. You will receive a prompt reply.

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Responsibilities include systems engineering and advanced techniques evaluation with particular emphasis on navigation, traffic control, communications, and electromagnetic warfare. BS or MS in EE or physics plus experience.

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To work under Project Director as lead engineer supervising a group concerned with shipboard antenna problems. Duties require knowledge and experience in antenna impedance, field strength measurements, and design of matching Networks in HF and UHF frequency range.

PRINCIPAL COMMUNICATIONS ENGINEER

Will direct programs involving communication circuits or systems. Position requires substantial experience in operation, design or evaluation of communication systems, as well as background in the inter-relations of antenna, transmitter, receiver, propagation, and terminal equipment.

ELECTRONICS ENGINEER

Will work with Test Manager on design, selection, modification and supervision of activity concerning instrumentation components and systems used in testing propulsion units. BS or MS in EE or physics and experience, preferably in the rocket instrumentation field.

INSTRUMENT ENGINEER

To work with Rocket Test Manager coordinating activities of instrumentation staff with other segments in the test group and throughout ARC. Responsibilities include design, selection, modification, and operation of instrument components and systems used in static testing of propulsion units. BS or MS in EE or physics and several years' experience.

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Positions Wanted

By Armed Forces Veterans

(Continued from page 102A)

FIELD SERVICE ENGINEER

Married, age 43, retired army officer with 22 years communications experience. Since 1960 in managerial position, guided missiles range division, major airlines. Extensive background in communications facilities construction and installation. Including broadcast stations, telephone central office, microwave radio systems and missile tracking stations. Holder of FCC First Class Radiotelephone License. Western U.S. or foreign assignment preferred. Write Box 4018.

IRE People

(Continued from page 93A)



Robert E. Frese (S'51–A'53–M'57–SM'59) has been named Chief Scientist at the U. S. Army Electronic Proving Ground, Fort Huachuca, Ariz. He has been promoted to the top civilian post at the Army Signal Corps installation from his previous position of Deputy Director of the Systems Development Directorate. From 1959 to 1961, he was Deputy for Scientific Affairs of the Combat Developments Directorate.

He came to the Proving Ground in June, 1959, after more than ten years with Michigan University as student and teacher. During his last two years there, he was an Instructor and Assistant Professor in the Department of Electrical Engineering, and served as a Consultant in the field of electronic countermeasures. He was a Research Associate at the Willow Run Laboratories of the University until he joined the faculty in 1957. During this five-year period, he was engaged primarily in the evaluation of vulnerability to countermeasures of air-terminating guided missile systems.

Dr. Frese received the M.S. degree in 1952 and the Ph.D. degree in 1959, both from Michigan University. In addition to membership in five honorary scientific societies, he is active in three professional organizations. He is Chairman of the Fort Huachuca Section of the IRE, and a member of the American Institute of Electrical Engineers, and the Armed Forces Communications and Electronics Association. Since 1959 he has been a member of the Board of Directors of the Consumers' Union of the U. S., Inc., publisher of *Consumer Reports*.

❖

(Continued on page 106A)

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(2) Its listings are fundamental —the way an engineer thinks.

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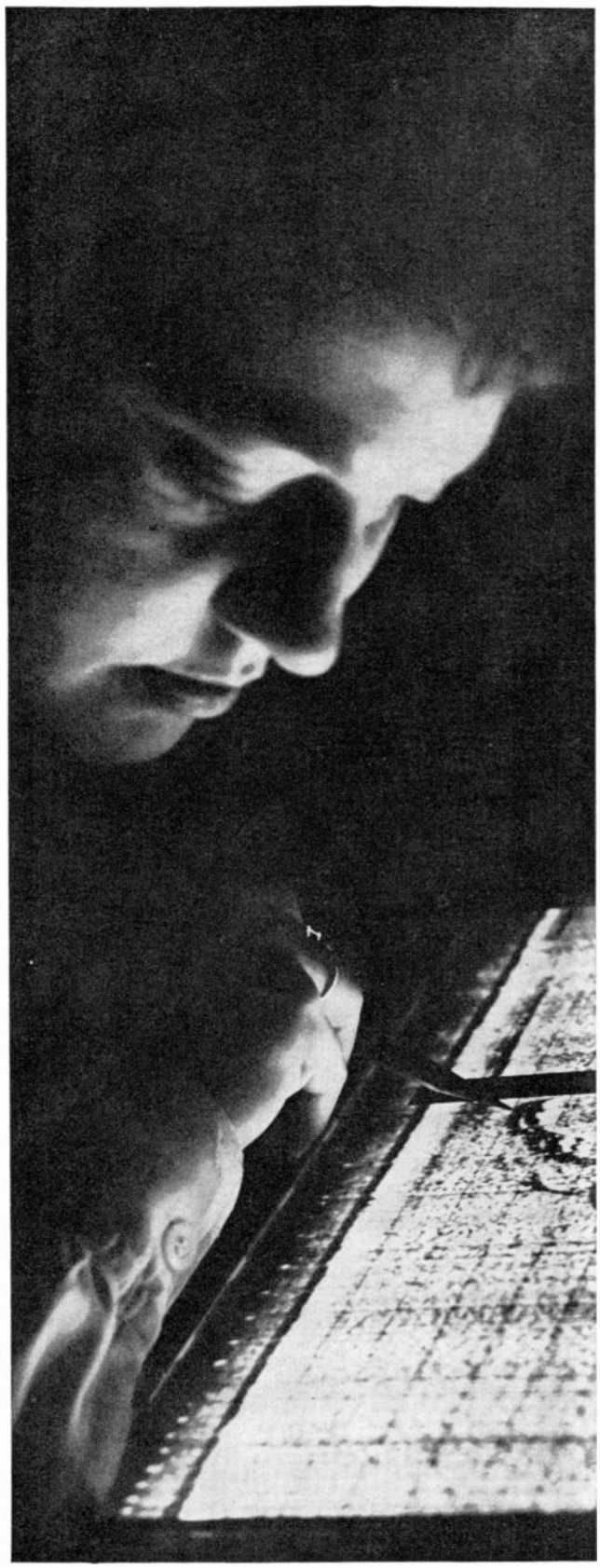
(4) Ads face listings thus helping to identify products by actual pictures.

(5) Product code numbers reduce complex and duplicate listings, saving you "searching" time and effort.

Study it . . . save it . . . work it.

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For additional information, please send your resume to:
Mr. William T. Whelan
Director of Research and Development

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(a division of the
Permanent Employment Agency)
825 San Antonio Rd.
Palo Alto, Calif.

IRE People

(Continued from page 104A)



John L. Herre (M'57), military electronics specialist, has been appointed Vice President, Government Relations, of General Instrument Corporation's Radio Receptor Division, Hicksville, N. Y. In his new post he will be responsible for all sales efforts of Radio Receptor's Marketing Group to government agencies and to original equipment manufacturers, including study, research, development, and production programs. He was most recently Government Sales Manager for the General Instrument Semiconductor Division and the Corporation's Applied Research Laboratory.

Prior to joining General Instrument in 1960, he was Sales Manager, Government and Special Accounts, for the Semiconductor Division of the Raytheon Company. Previously he had been a technical representative to the United States Armed Forces, and Manager of Government Sales and eastern industrial sales activities for Philco Corporation. During World War II, he served in various technical capacities with the United States Air Force.

Mr. Herre is a member of the American Rocket Society and the Armed Forces Communications and Electronics Association. He was educated at the University of Pennsylvania and the University of Chicago.

❖



Murray Hoffman (M'56) has joined Airtron, a division of Litton Industries, as Engineering Manager for Ferrite Devices. For the past nineteen years he has worked with the design and development of microwave components and systems, most recently with the Philco Corporation, as Supervisor of their Microwave and Antenna Group. Previously, he was with Litton Industries' Maryland Division for three years, where he directed research on high-precision automated machinery for surface-wave antennas. He was also with the Sperry Gyroscope Corporation for seven years, working with miniaturized lenses for missile guidance and tracking.

Mr. Hoffman received the B.S. degree in electrical engineering from the City College of New York in 1943, and the M.S. degree in physics from New York University in 1951. He is a member of the American Physical Society and has written numerous technical papers on microwave components and their application.

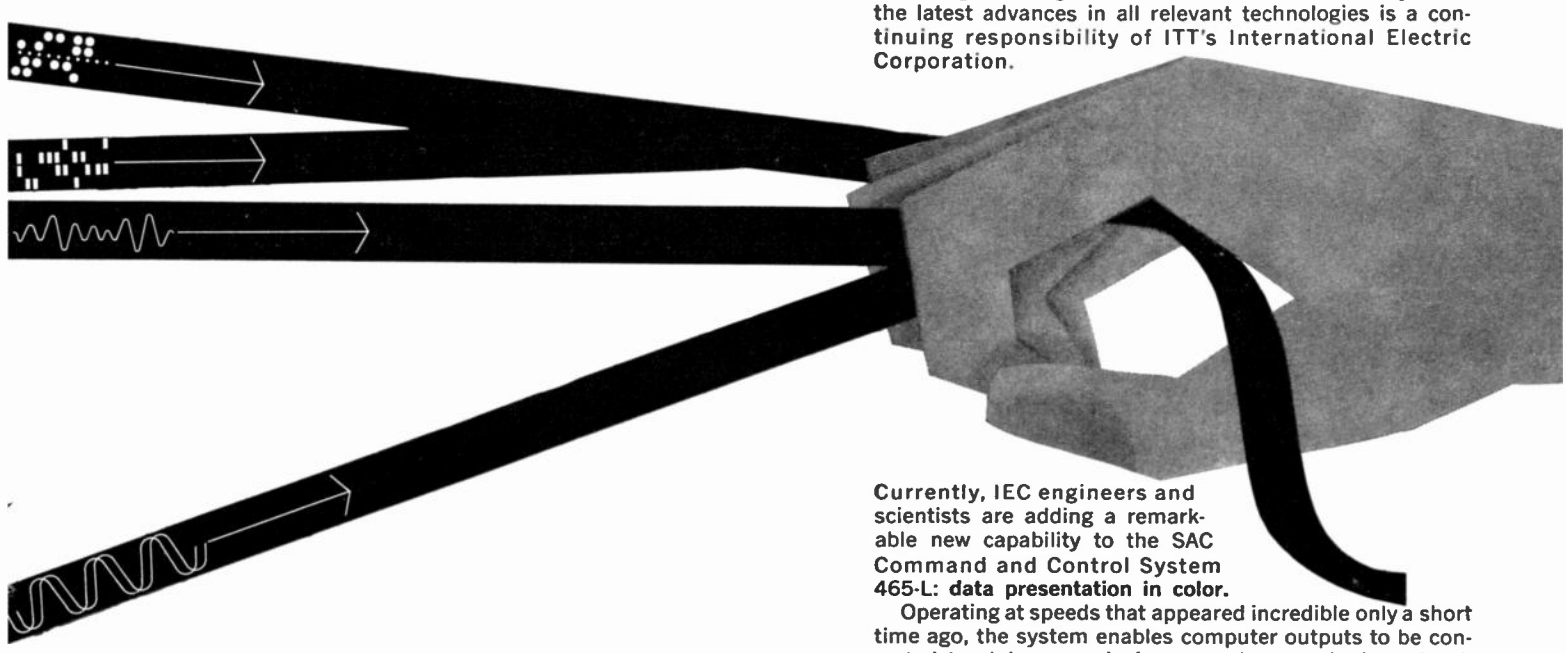
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(Continued on page 108A)

ENGINEERS, PROGRAM ANALYSTS:

securing man's control of the complex weapons he has invented

Factoring into large scale command and control systems the latest advances in all relevant technologies is a continuing responsibility of ITT's International Electric Corporation.



Currently, IEC engineers and scientists are adding a remarkable new capability to the SAC Command and Control System 465-L: **data presentation in color.**

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SYSTEMS IMPLEMENTATION ENGINEERS. Electronic engineers to develop tests for stressing and evaluating communication-display-computer systems. Recommend improvement and refinements. Also, field positions for installation and integration of digital command/control systems.

INFORMATION SYSTEMS ENGINEERS. For design of command/control and advanced communications systems. Experience in traffic, antenna and propagation theory, and mathematics as applied to communications and space technology.

DIGITAL SYSTEMS ENGINEERS. Engineers with management ability to direct sub-systems engineering effort on a global command/control system. Experience is desired in message traffic control, data processing systems, data display and multi-sequencing techniques.

Write fully in confidence to Mr. E. A. Smith, Room 86-MA, ITT-International Electric Corp., Route 17 & Garden State Parkway, Paramus, New Jersey.

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RADIO FREQUENCY INTERFERENCE CONTROL
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■ ELECTRONIC RECONNAISSANCE SYSTEMS

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BSEE with 3 to 5 years experience. Must be systems oriented in the reliability field, preferably with some operations research background. Will be responsible for designing and implementing a reliability program on a large electronic system and will make the necessary data reduction involved in such a program.

SR. DESIGN ENGINEERS
BS with 8 years experience, of which 5 must have been in design in two or more of the following: digital, RF, pulse, audio, CRT, photorecorders, magnetic recorders, pulse multiplex and frequency multiplex. To assist in evaluation of complex electronic reconnaissance systems.

SR. DESIGN ENGINEERS
BS with 3 to 5 years experience in RF and microwave receivers, digital display circuits, data handling and CRT displays including storage tube circuits. To assist in evaluation of complex electronic reconnaissance systems.

■ HYDROACOUSTICS

SENIOR SONICS ENGINEER
BS or MS in ME or Physics with at least 5 years experience in Industrial Sonics. Should have background in sonic cleaning, processing and impact drilling plus a basic knowledge of acoustics, general physics and chemistry.

■ TEST EQUIPMENT & INSTALLATIONS

PROJECT ENGINEERS
To supervise design and integration of test stations. Knowledge should include 1 or more of the following areas: flight control systems, radar indicators, HF-VHF navigation and communication equipment, microwave equipment, antenna systems and ECM. Should be familiar with all types of testing techniques and equipment associated with particular areas of interest. BSEE.

SENIOR DESIGN ENGINEERS
BSEE with thorough background in one of the following: microwave signal generators and receivers; low frequency signal generators, HF-UHF signal generators, digital and pulse circuits. AGE Systems.

■ TECHNICAL WRITING

Requires thorough background in the electronics industry in preparation of military handbooks and manuals or in engineering proposals.

For more information on the areas discussed above and how your background might be utilized in one of these programs, send a resume to Mr. M. J. Downey, Dept. 20.

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IRE People

(Continued from page 106A)



William D. Gabor (A'49–M'55) has been appointed Senior Project Engineer in the Communications Division of Manson Laboratories, Inc., Stamford, Conn.

Prior to joining Manson, he was employed by the Perkin Elmer Corporation, where he had responsibility for the concept and design of automatic camera exposure systems. He has also been associated with Trak, Inc., and Sanders Associates, in the development of sophisticated test equipment, high speed countermeasures and communications systems. In addition, he has published extensively on VHF/UHF automatic control systems and design of test equipment, and holds over 10 patents in these areas.

Mr. Gabor is a graduate of the R.C.A. Institutes.

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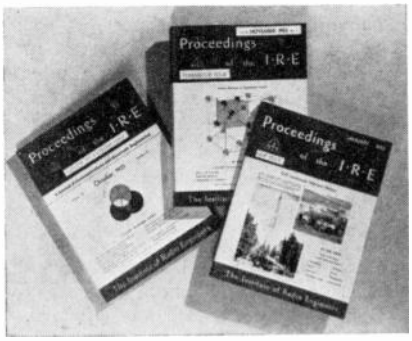


Allan A. Kunze (A'47–SM'57), of the Rome Air Development Center, Griffiss Air Force Base, Rome, N. Y., has been named General Chairman of the 8th National Communications Symposium, to be held in the Utica Memorial Auditorium, October 1–3, 1962. The Symposium, sponsored by the IRE, Professional Group on Communications Systems, has been an annual event in the area since 1955, and has become one of the nation's largest gatherings of electrical communication experts.

Dr. Kunze is Senior Scientist in the Advanced Studies Office at the Rome Air Development Center, and has both authored and delivered numerous articles and papers on communication systems and related electronic techniques at national meetings. Born in East Lansing, Michigan, he received the B.S. degree in 1934 from Michigan State College. He attended the University of Michigan on a fellowship for three years, obtaining the M.S. degree in 1935. Subsequently, he continued at the University of Michigan from 1935 to 1937, and attended New York University from 1947 to 1948, continuing his graduate work towards the Ph.D. degree. Following a year as research engineer for King-Seeley Company, Ann Arbor, Mich. he was Instructor in Electrical Engineering at the Ohio State University, from 1937 to 1941. During this same interval he was a consulting broadcast engineer for a number of Ohio broadcast stations. In 1941 he became an Assistant Professor in the Electrical Engineering Department of the Uni-

(Continued on page 110A)

BUILDING THE FUTURE IS A BIG JOB!



The radio-electronic engineers who form the membership of the Institute of Radio Engineers must remain years ahead of actual production, in order to pave the way for the products of tomorrow, through research and development today.

The special issues of "Proceedings of the IRE" help these men transform the theory of today into the production lines of tomorrow. Pictured above are the October 1951 Color Television Issue; November 1952 Transistor Issue; and January 1953 UHF Issue (all now out of print).

Special issues on special subjects are more than usually helpful, but every issue of "Proceedings of the IRE" is filled with facts and figures to keep you far ahead of the field!

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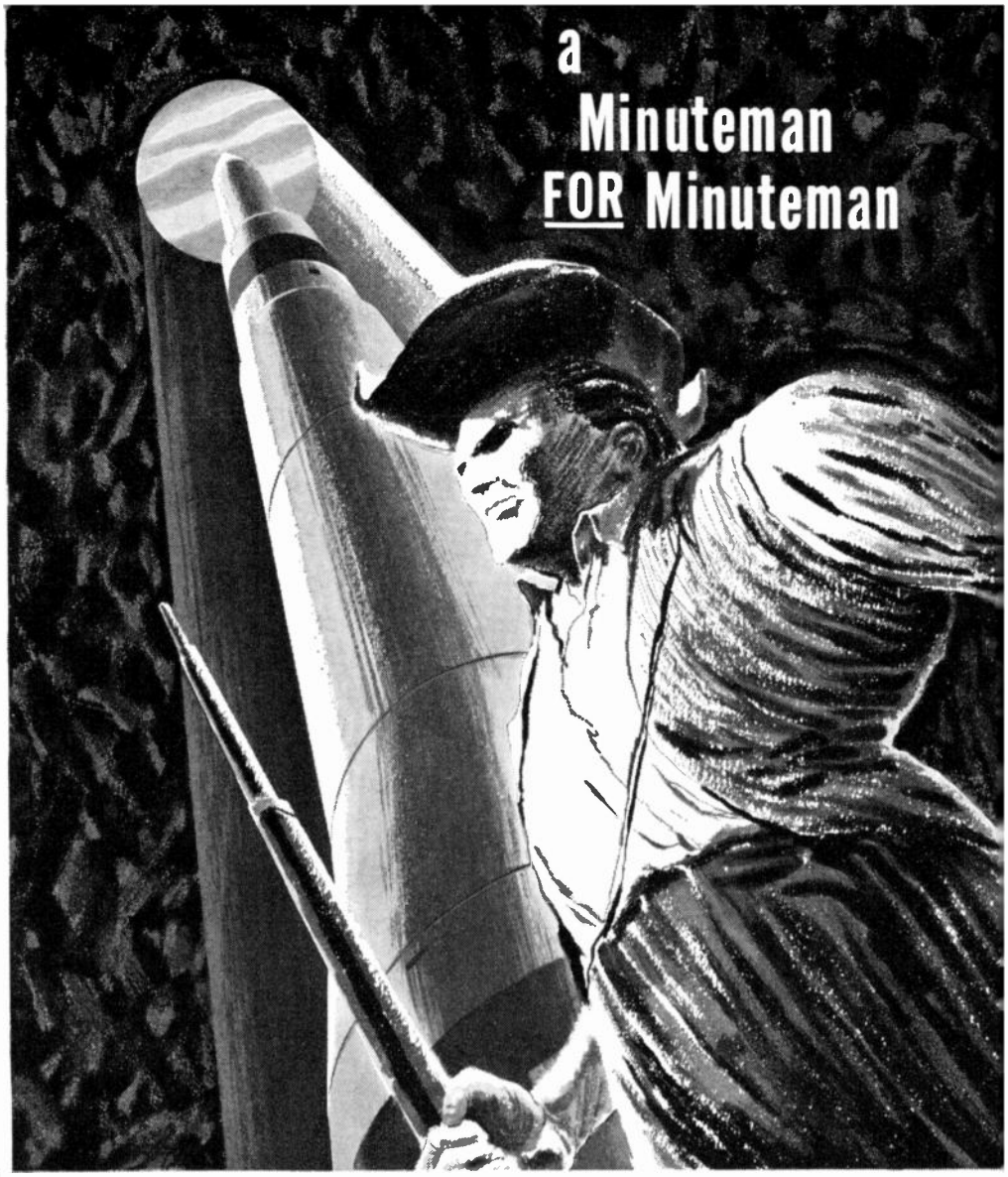
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CONTRIBUTORS URGENTLY NEEDED to man this and other recently-acquired command and control systems projects in the aerospace, surface and sub-surface communications / navigation fields. Candidates should have BSEE or equivalent with backgrounds in one or more of the following: solid state design, digital communications and / or wideband techniques. We are located ten miles northeast of Buffalo in the picturesque suburban village of Williamsville.

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Task requires a scientist with five or more years experience in research and development of advanced infrared detectors and optical components. Applicant will act in a lead capacity in present programs and in initiating programs in new proprietary areas. An advanced degree in Physics desired.

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IRE People

(Continued from page 108A)

versity of Alabama, where he instituted the course in communications.

During World War II, he served as Officer in Charge of the U.H.F. (Fighter Control) Section of the Eatontown Signal Laboratory. He was also technical advisor and later Chief of the Communications Laboratory, Watson Labs, New Jersey, responsible for the research and development of ground communications and direction finding equipment for the USAF. In 1951 he moved to the Rome Air Development Center, and in 1952 became Chief of the Equipment Development Division. Following this, he was Technical Director of the Directorate of Electronics, Senior Scientist in the Advanced Studies Office for the Center. During this period, he has been a Signal Corp member on Division 13 of the National Defense Research Council concerned with Direction Finding, the USAF member on the Research and Development Board, Panel on Communications, the USAF member on the Committee for Military Communications Systems Standards, and a member of the USAF team on a mission to England for Electronic Research Planning Coordination.

Dr. Kunze is a founding member of the IRE Rome-Utica Chapter, in which he has served as Chairman, and in other capacities, including Program Chairman and Arrangements Chairman. He was also a member of the Executive Committee of the IRE. He helped in the founding of the Rome-Utica Chapter of the Armed Forces Communications and Electronic's Association, in which he served as President and other capacities, being awarded the "Man of the Year" plaque in 1957. He was General Chairman of the 1961 Mohawk Valley Engineers Banquet Committee, and General Chairman of the 9th Mohawk Valley Management Seminar. He has also served as Program chairman, member of the advisory staff, member of the Board of Directors, and Executive Vice Chairman of previous National Communications Symposia.

❖

Edward F. Lazar (SM'55) has been appointed Marketing Manager of Sperry Rand's Electronic Tube Division facility, Great Neck, N. Y. He has had 29 years of engineering, purchasing, marketing and administrative experience with several Sperry divisions, and was active in Sperry's development and sale of early klystron tubes. His most recent responsibility was the establishment of a comprehensive cost reduction program for the Sperry Gyroscope Company as an Assistant to the Vice President for Administration. Prior to that he was Assistant to the Executive Vice President.

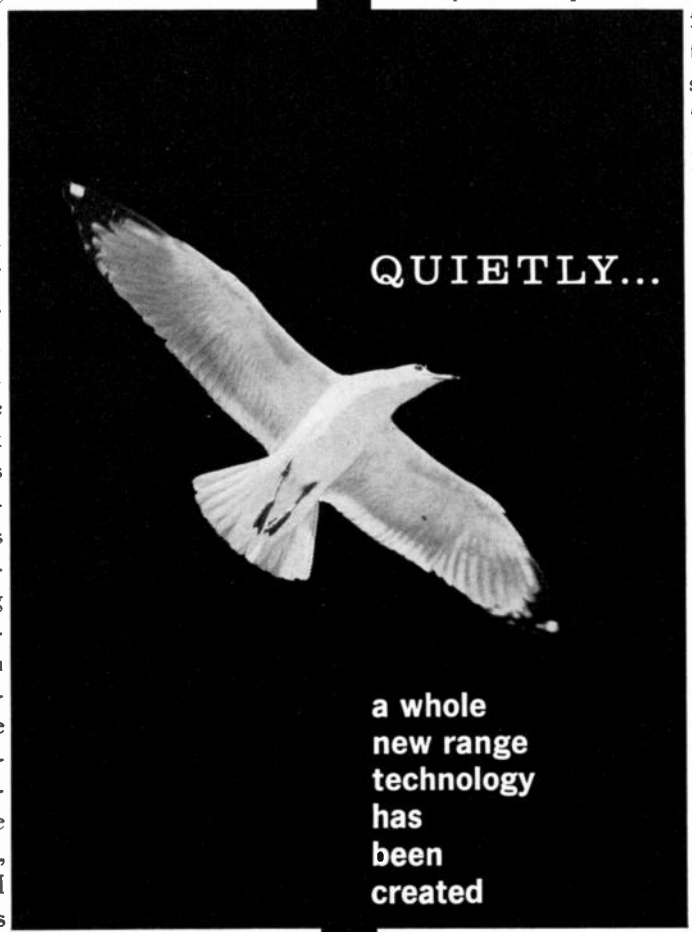
(Continued on page 112A)

Coming soon!
IRE DIRECTORY!

A Message to the Engineer/Scientist Community At Large — and a Question: there's a dynamic technological race going on at the Atlantic Missile Range, a race between the fast-increasing capabilities of new missiles and space vehicles and the capacity of range instrumentation to test their performance. ☐ We wonder how much you have heard about this. . . . and about the challenge it offers engineers and scientists with PAN AM at Cape Canaveral? ☐ You may know a small segment of the work . . . many do. But only a handful are aware of its scope. In fact, we of PAN AM'S Guided Missiles Range Division sometimes think that only the ubiquitous seagulls know the full story of the new range instrumentation technology we've created in the 9 years we've been charged with development and management responsibilities for AMR by the U.S. Air Force. ☐ The measure of the distance we've come is the measure of the technological jump between MATADOR and MARINER. ☐ In the simplest terms, this has meant acquiring ever greater funds of data, of ever higher accuracy, at ever greater distances — and converting and transmitting it at ever increasing speeds. ☐ FIRST, the existing range instrumentation and communications techniques were pushed to the utmost bounds of their capacities — THEN they were replaced with new range systems built to new concepts, as specified by PAN AM engineers and scientists backed by research groups. ☐ *Today — a new phase of range technology development is under way — staff build-up is proceeding on schedule.* ☐ To meet the demanding requirements of both today and tomorrow, much of the work of the Range is divided into three time projections:

(A) designing and implementing range instrumentation for launches programmed for this year and next;

(B) developing range technology concepts required for launches in the near future (Dyna-Soar, Gemini, Apollo test vehicles, advanced Saturn boosters and Nova);



(C) advanced planning, looking forward as much as 15 years. Includes considering such problems as how to service, launch, track and recover information from multi-million pound thrust booster systems and anticipating the problems associated with the launching and support of nuclear propelled boosters and spacecraft.

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IRE People

(Continued from page 110A)

From 1954 to 1957, he was Vice President and General Manager of Sperry Rand's Wheeler Electronic Company, Waterbury, Conn., manufacturer of control and communication equipment and electronic components. Earlier he was Director of the Federal Department, Manager of federal and special electronic sales and Chief Engineer and Manager of aircraft radio sales for the Sperry Gyroscope Company.

Mr. Lazar received the B.S. degree in electrical engineering from Cooper Union, New York, N. Y. He is a senior member of the Institute of the Aerospace Sciences and holds two patents on remote control devices.

❖



B. D. Loughlin (A'40–M'45–SM'53–F'55) one of the nation's foremost pioneers in color television technology, has been elected Vice President for Research of Hazeltine Research, Inc., a subsidiary of Hazeltine Corporation. Winner of the Vladimir K. Zworykin Television Prize of the IRE and the David Sarnoff Gold Medal Award of the Society of Motion Picture and Television Engineers, he was cited by *Fortune* Magazine in November, 1955 as "an architect of color television . . . who made outstanding contributions to theory, understanding and practice."

He joined Hazeltine upon his graduation from The Cooper Union in 1939, with a B.E.E. degree. A year earlier, he had built his own home TV set and TV test equipment, and his first task at Hazeltine was developing circuitry for TV test equipment. In 1945, he received a special E.E. degree from The Cooper Union, and in 1946 he was awarded an M.S.E.E. degree from Stevens Institute of Technology. In 1956, he was named Chief Engineer of the research division of Hazeltine Research Corporation, and the following year, he became a Research Consultant. In his new capacity, he and his engineering group will continue to concentrate their efforts on the simplification of color television receivers, with particular emphasis on three-gun tube receivers. Other projects include improved circuitry for monochrome and color television in both the VHF and UHF bands.

Mr. Loughlin is a fellow member of the Radio Club of America. He is also a member of the National Television Systems Committee (Panel 14) and American Institute of Electrical Engineers.

❖

(Continued on page 114A)

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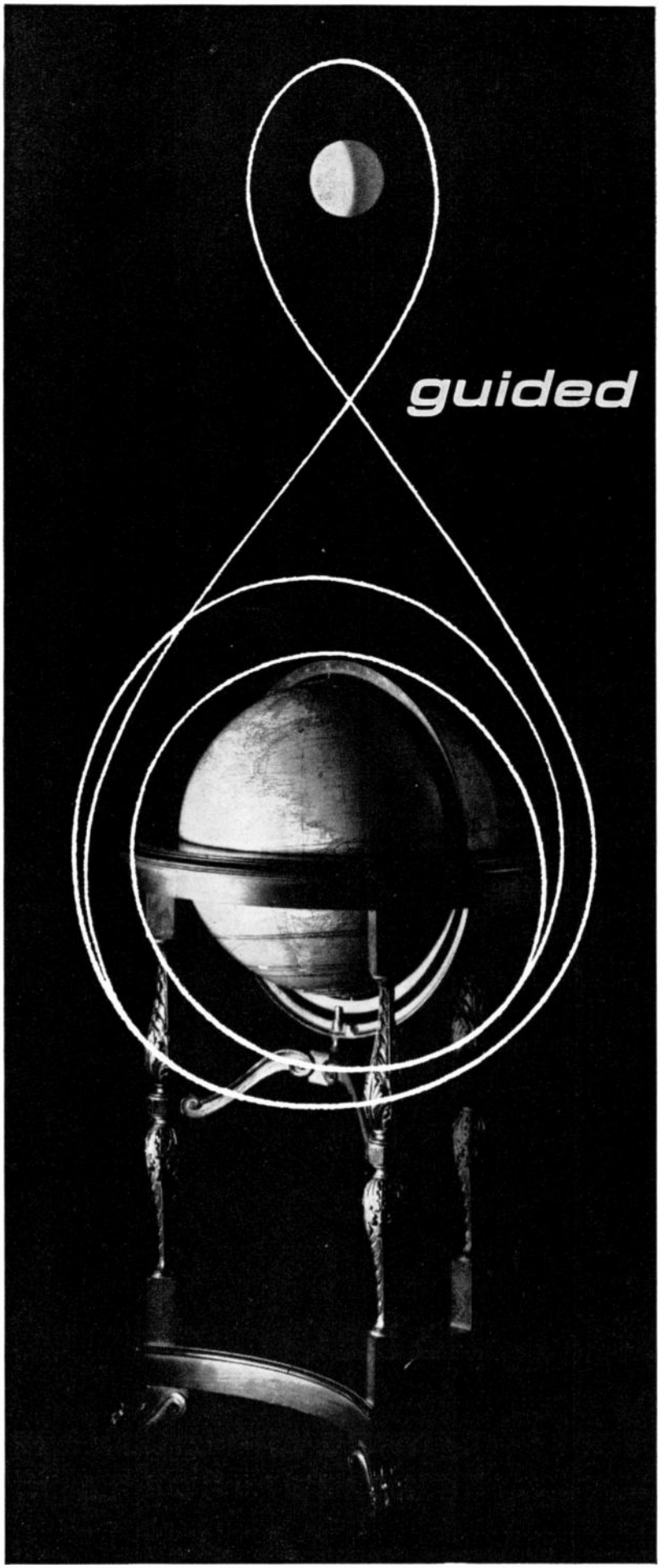
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LOS ANGELES RESEARCH AND DEVELOPMENT LABORATORY (Airborne Digital Computers & Advanced Inertial Guidance Systems) Electronic System Mechanization ■ Circuit Design Engr. ■ Re-entry Simulation ■ Aerodynamicist ■ Advanced Guidance System Analysis ■ Programmer

BOSTON RESEARCH AND DEVELOPMENT LABORATORY (Advanced Inertial Guidance Systems & Components for Future Aircraft, Ballistic Missiles & Space Vehicles) Circuit & Logic Design (Digital Computers) ■ Electronic Circuit Engrs. ■ Systems Engrs. & Mathematicians ■ Physicists ■ Instrument Engrs. ■ Electromagnetic Engrs.

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(Continued from page 112A)



Harold Marchant (M'61) has been appointed as Manager of Digital Devices for Scientific Data Systems, Inc., Santa Monica, Calif. His background includes experience designing circuits for a new line of punched card computers at Remington-Rand Univac; the development of special-purpose digital computers at Solartron, Ltd.; and the design of drum memories and digital circuit techniques for Ferranti, Ltd.

Mr. Marchant received a B.S. degree in physics from London University, and a Certificate of Advanced Study in control engineering from Cambridge University. He is a member of the British Computer Society.

❖



Basil R. Myers (S'51–A'52–M'56–SM'58) has been appointed Professor and Head of the Department of Electrical Engineering at the State University of Iowa, Iowa City, Iowa, effective from July 1, 1962.

Professor Myers has the degrees of B.Sc. (Hons.) from the University of Birmingham, England, and the M.S. and Ph.D. degrees in electrical engineering from the University of Illinois.

From 1951 to 1956 he was a member of the technical staff of Bell Telephone Laboratories, Inc., Murray Hill, N. J., working in exploratory transmission systems engineering. He was with the Department of Electrical Engineering, University of Illinois, during 1950–1951, 1956–1959 and 1961–1962, doing both teaching and research. From 1959 to 1961, he was Professor and Chairman of the Electrical Engineering Department, University of Waterloo, Waterloo, Ontario, Canada.

Dr. Myers is a member of the AIEE, the British IEE, ASEE, AAUP, SIAM, Tau Beta Pi, Sigma Xi, and Eta Kappa Nu.

❖



Helmut J. Schwarz (SM'62) is a Senior Staff Research Scientist with the United Aircraft Corporation Research Laboratories. In addition he has been recently appointed an Adjunct Professor of Physics at the Graduate Center of Rensselaer Polytechnic Institute. Until

(Continued on page 116A)



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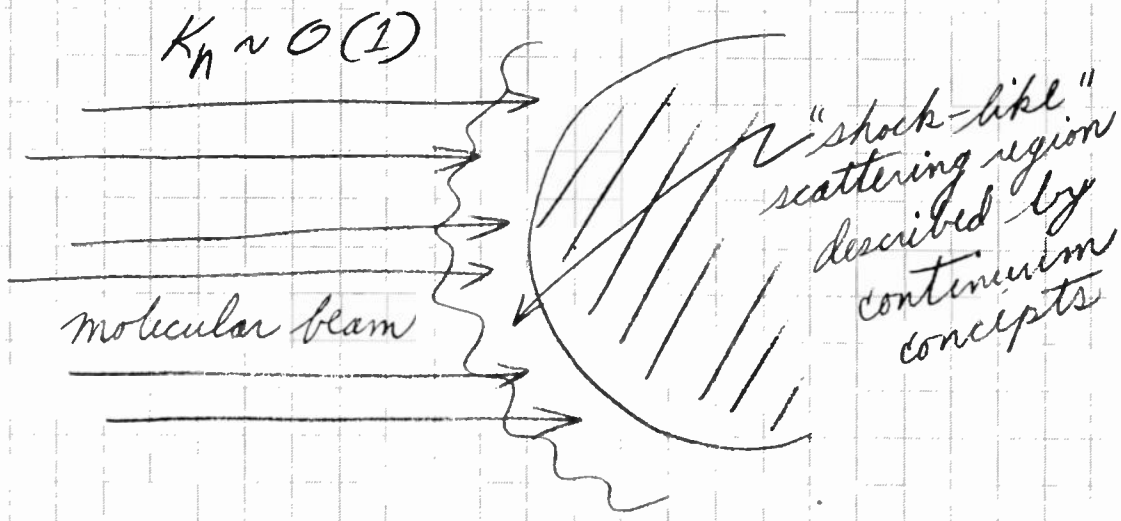
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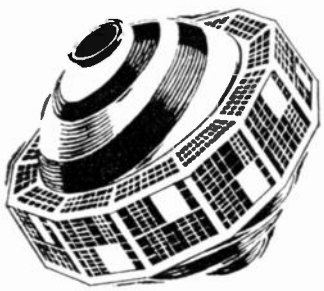
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(Continued from page 114A)

May 1962, he has been staff consultant to the Electron Beam Chief Development Engineer at Hamilton Standard Division of United Aircraft Corporation. He came to Hamilton Standard from the Radio Corporation of America, Somerville, N. J., where he was in charge of the Vacuum Deposition Laboratory. Here he was specifically concerned with research and development in micromodules.

From 1950 to 1957, he was associated with the Technical College, Brazilian Ministry for Air, and in 1904 was asked to organize the Institute for Physical Sciences, where he served as a Professor. Part of his duties included travel in Europe and America for study and evaluation of universities, institutes, and research laboratories. In addition to his other activities he taught at the University of Brazil on a part-time basis. Before going to Brazil he was an Assistant Professor at the University of Bonn, where he conducted research and development in vacuum techniques, constant acceleration tubes, ultraviolet photo cells, and high ohm resistances.

Dr. Schwarz was educated in his native Germany and received his undergraduate degree at the Reform-Real-Gymnasium. He was awarded Master's and Doctorate degrees in gaseous electronics from the University of Bonn. He is co-author of a book on high vacuum physics and has published more than 30 scientific papers. He is a member of several scientific societies, and an invited member of the Standards Committee of the American Vacuum Society. He is listed in the 10th Edition of "American Men of Science."

❖



John L. Turner (S'49–A'50–M'55–SM'59) has been appointed Manager of Tube Production of Edgerton, Germeshausen and Grier, Inc. He previously served as an engineer in the Magnetron Development Laboratory of Raytheon Company from 1949 to 1955

Mr. Turner received the B.E.E. degree from Rensselaer Polytechnic Institute in 1949, and the M.A. degree in physics from Boston University in 1953. He is a member of the Rensselaer Society of Engineers, Eta Kappa Nu, Sigma Xi, Tau Beta Pi, and Pi Delta Epsilon.

❖

Donald N. Yates (SM'57) has been elected a Vice President of Raytheon Company and appointed General Manager of its Missile and Space Division. He joined the Company last year as Director of Technical Planning.

He commanded the Air Force Missile Test Center at Cape Canaveral, Fla., from

(Continued on page 118A)

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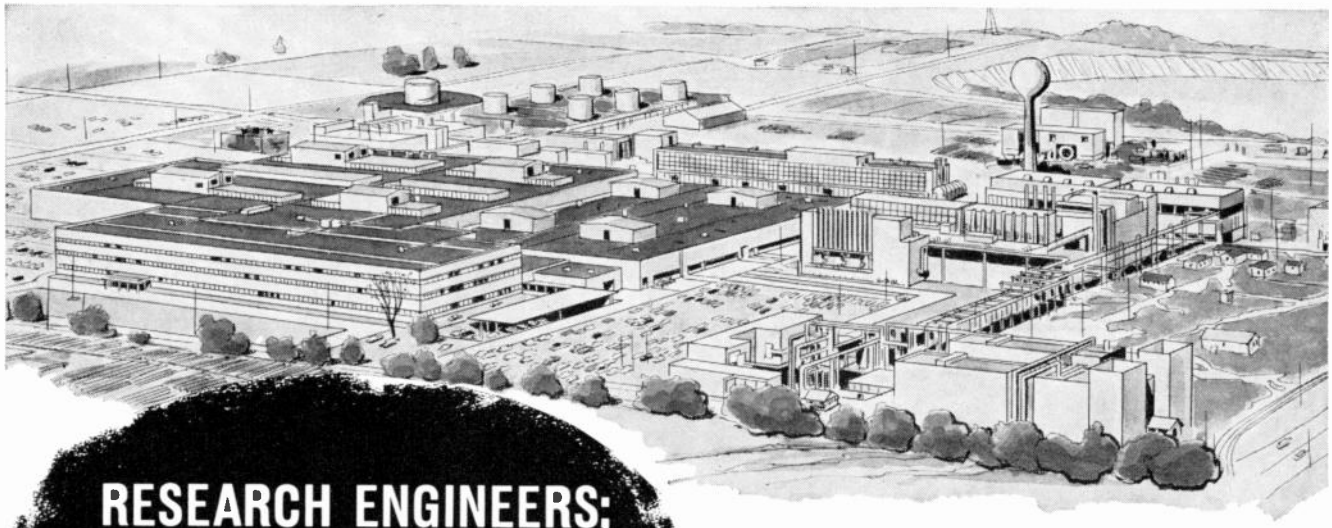
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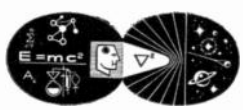
DIRECT CONVERSION DEVICE SPECIALISTS—Sc.D. or equivalent in ChE. Background in electrochemical engineering, liquid metal technology or related fields. For development of liquid metal cell devices and handling of liquid metals in the environment of an electrochemical conversion system.

Also, ME with some experience in design and fabrication of thermionic converters or related devices (microwave tubes, small ceramic-metal tubes, etc.). Requires knowledge of thermionic materials, refractory metals, high temperature ceramics, thermionic emitters, and mechanical design problems associated with high temperature electron devices.

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(Continued from page 116A)

July, 1954, to May, 1960. He retired as a Lieutenant General in 1961. At the time of his retirement, he was Deputy Director of Defense Research and Engineering in the Office of the Secretary of Defense.

A 1931 graduate of the U. S. Military Academy at West Point, he received a master's degree from California Institute of Technology. He also received an honorary LL.D. degree from Bowdoin College. He is a member and past President of the American Meteorological Society, a Fellow of the American Rocket Society, an Associate Fellow of the Institute of Aeronautical Sciences.

General Yates earned his wings at Kelly Field, Tex., and later received the rating of Command Pilot. His military decorations include two Distinguished Service Medals, and both the Army and Navy Legion of Merit Awards for outstanding services in connection with several of the country's missile and satellite programs. He has also received the French Legion of Honor, the French Croix De Guerre with Palm, the Order of the British Empire, and the Portugese Ordem Militar De Cristo.

❖



Benjamin E. Shackelford (A'23–M'26–F'38) (L) has been elected to the Board of Directors of Butler Roberts Associates. He will assume charge of the firms Development Application Group for millimeter wave generating devices.

Dr. Shackelford received his Ph.D. degree in 1916 from the University of Chicago, and has been active in the electronic industry for the last three decades. He served as President of IRE in 1948, and was a Director from 1945 to 1950 and from 1953 to 1955.

❖



Onnig Shahan (S'50–A'51–M'54) has been named Section Head of the Stripline Systems section at Micro-Radionics, Inc., Van Nuys, Calif., manufacturers of precision microwave components and test equipment.

He has gained an industry-wide reputation through his work in designing microminiature microwave components using revolutionary stripline techniques. For a number of years he was a Senior Member of the Research and Advanced Development Laboratory at Hycon Manufacturing Company. Earlier, he was a technical staff member in the Microwave Laboratory at

Hughes Aircraft Company. He has conducted pioneering efforts in higher frequency applications utilizing stripline techniques, such as developing a Y circulator operating above 17 Gc. In his new position at M-R-I, he will continue his work in the microminiaturization of systems and components using stripline.

Mr. Shahan is an electrical engineering graduate of the University of California.

❖

Sidney Shapiro (A'55–M'60) has been named Project Engineer in charge of a new inertial navigation system development program at the Guidance and Control Systems division of Litton Systems, Woodland Hills, Calif. The system, being developed under sponsorship of the Navy Bureau of Weapons, will employ a light-weight, gyro-stabilized inertial platform and micro-miniature electronic circuits to provide high navigational accuracy with a light-weight system. He has been with Litton five and one-half years. He received a B.S.E.E. degree from Pratt Institute and an M.S.E.E. degree from New York University.

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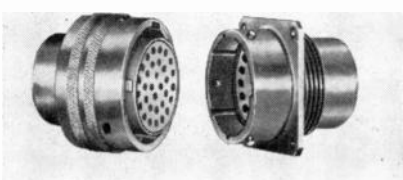
NEWS New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your IRE affiliation.

(Continued from page 76A)

Miniature Connectors

Electrical connectors providing intermateability with existing specification MIL-C-26482 connectors, plus military standard crimp contact geometry, are now available in the PT-SE series of Pygmy Electrical Connectors, from **Scintilla Div., The Bendix Corp.**, Sidney, N. Y.



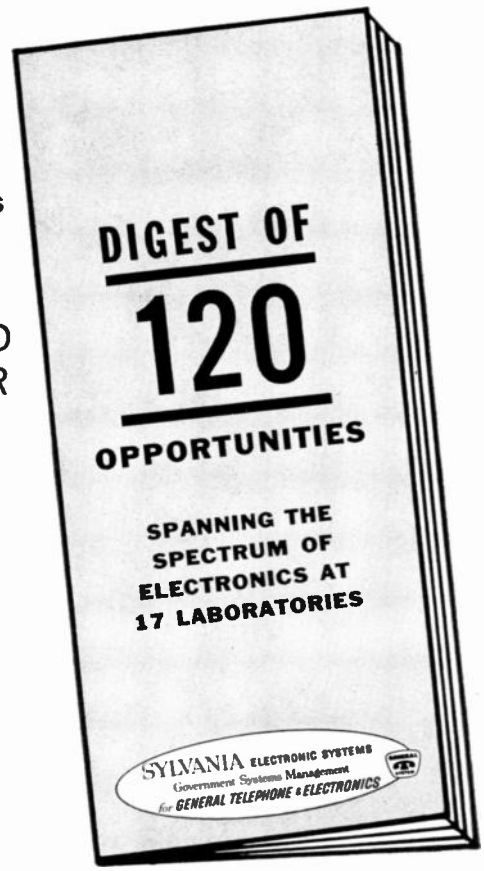
A new addition to the line is the SE series which combines proved materials and designs with MIL-C-26636 contact geometry and MS3190 wire well area. Complete intermateability of the new SE series with several million PT and SP solder type, CE crimp type, and MIL-C-26482 connectors now in the field is assured. The Bendix design features eliminate wiring diagram changes when using the SE connector, retain solder option, and provide contacts compatible with military standardized application tooling.

Complete data concerning the PT-SE connector series is presented in brochure SL-109, available on request.

(Continued on page 120A)

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(Continued from page 119A)

Silicon Controlled Rectifiers

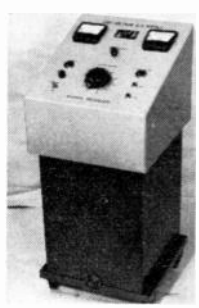
Silicon controlled rectifiers, 150 ampere (235 amp RMS), are now in production at **International Rectifier Corp.,** 233 Kansas St., El Segundo, Calif. These components are capable of replacing standard ignitron and thyratron tubes. The new devices are suitable for inverting, frequency changing, motor speed control and ac power switching, particularly in space vehicles, since they may be operated in any position and in weightless environments.



Five types of the 150 ampere devices available are designated 150RC5-150RC50, and cover the peak reverse voltage range from 50 to 500 volts. All units feature double ceramic construction (dual ceramic construction, dual ceramic insulators) to assure positive separation and high insulation between gate and cathode leads. Dual gate leads on all 5 types prevent improper or 'false' firing due to pick-up of extraneous signals and power lead transients. The new devices feature extremely high surge capabilities (3000 amperes maximum peak 1 cycle nonrecurrent) and a maximum peak forward voltage drop of 1.7 volts (@25°C @ 150 amperes average, 472 amperes peak). Typical turn-on time is 3 to 7 microseconds (@25° C for 100 amperes dc, resistive load) and typical gate current to fire is 35 ma @25° C. All units have an operating temperature range from −40 to 125°C, measure approx. 6⅞ inches over-all length, and weigh approx. 8 ounces. For more detailed data, request Bulletin SR-364. Price: $99.50 to $528.00 each—1 to 24 quantities. Delivery: from stock.

High Voltage Power Supplies

A new line of DC Power Supplies known as the TC Series has been announced by **Del Electronics Corp.,** 521 Homestead Ave., Mount Vernon, N. Y. These supplies are multi-purpose units available from 30kv up to 80kv (in steps of 10kv) and current up to 5 milliamps. They are designed for single phase, 115V 60 cps input and output is continuously variable from zero to maximum voltage. The polarity is reversible.



Typical applications include laboratory use; maintenance, such as dielectric testing; industrial, such as a power supply for electrostatic painting, electrostatic flocking, etc.

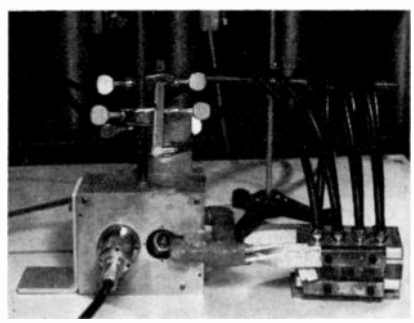
Designed into this series are complete safety and protection features—zero start interlock, preventing accidental turn-on of supply; automatic stored energy discharge switch; surge limiting resistor; fixed overcurrent relay set for 120% of rated current.

Descriptive material and prices can be obtained from the manufacturer.

New Laser Technique

A technique for concentrating the energy of a laser to multiply its instantaneous power as much as 1000 times has been developed by **General Dynamics Corp., Electronics Div.,** 1400 Goodman St., Rochester 9, N. Y.

Using the same modulation method, a laser beam has also been switched on and off at rates up to 100,000 times a second.



The new modulation technique increases the potential for such uses. A powerful non-uniform magnetic field surrounds the laser, forcing it to store up energy. When the field is removed, the laser releases energy in a concentrated pulse many times more powerful than an uncontrolled discharge.

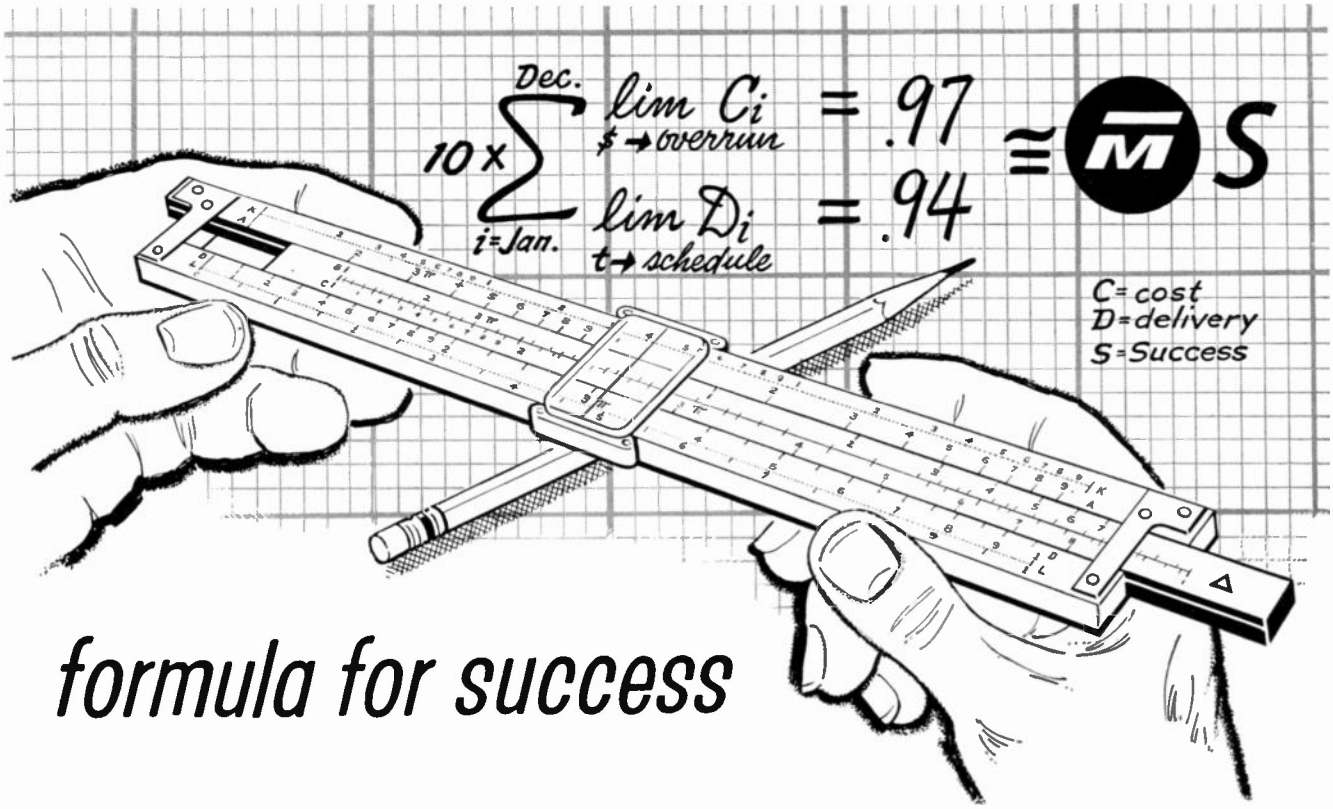
Under previous methods of control, either the laser's energy output was greatly reduced, or the intensity of the light beam damaged mechanical or optical devices employed, or both. The magnetic field control eliminates both problems.

The firm used a synthetic ruby rod about ¼ inch in diameter and 1½ inches long. The rod was cooled to the temperature of liquid nitrogen (321° below zero F.) and enclosed in a double-walled nitrogen-filled flask to maintain the rod at that temperature.

A single turn of heavy copper wire around the middle of the rod was used to create the magnetic field. A capacitor discharge of 138,000 amperes through the coil produced a magnetic field of about 60,000 gauss at the center of the rod tapering off to about 8,000 gauss at the ends. The varied distribution of the magnetic field is an important control factor.

Discharge of the capacitors was timed to coincide with the maximum intensity of the pumping lift. Normally the laser would have started to emit its light beam at about the same time in a burst gradually increasing to a peak and then even more gradually dwindling to zero.

(Continued on page 122A)



formula for success

Over a ten-year period, Martin Company, Aerospace Division of Martin Marietta Corporation, has achieved this remarkable record:

Ninety-seven per cent of our total volume of products for defense and other agencies were produced *at or below cost objectives.*

Ninety-four per cent of our projects were completed *on time or ahead of schedule.*

We think this record provides a good formula (as equated above) for *continued* Martin success.

And it is a "yardstick" that Engineers and Scientists can use to measure the excellent potential of their professional futures with Martin Company in Baltimore.

Right now, we have significant opportunities for GRADUATE ENGINEERS and SCIENTISTS in our:

ELECTRONIC SYSTEMS & PRODUCTS DIVISION

With our background of extensive and successful experience in this field, we are moving ahead rapidly on challenging, advanced electronics programs. As these programs increase—both in scope and number—we need additional top-level scientists and engineers who are skilled in a variety of technical disciplines.

Many of these openings are at the supervisory and senior technical level and require advanced degrees. There are immediate opportunities in the following areas:

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Guidance and Navigation
Microwave
Reconnaissance
Command and Control
Infrared
Human Factors
Solid State Circuitry
Modern Packaging
Data Processing

If you believe—as we do—that Martin Company's formula for continued success can be *your* key to significant professional accomplishment and attainment, we'd like to hear from you. Please write —in the manner which you believe best presents your qualifications—to Mr. J. W. Perry, Director of Executive Staffing, Dept. E-30

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The Lincoln Laboratory is a center of research and development in advanced electronics, with responsibilities in national defense and space technology. Scientists of many disciplines participate in a program directed toward extending the range and depth of scientific knowledge and solving problems fundamental to the security of the nation.

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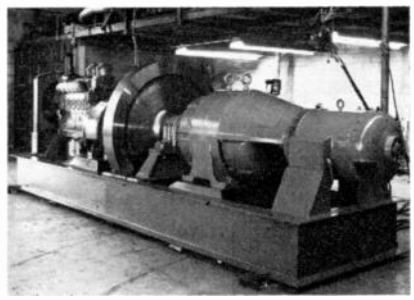
(Continued from page 120A)

Under the influence of the magnetic field, however, the ions in the ruby crystal could not radiate cooperatively to produce a characteristic laser beam. Most of the light energy was stored up in the crystal and not released until the field was removed.

The experiment has also been performed at room temperature, which requires magnetic fields ten times greater to achieve similar control.

50 to 355 kw Supply

The "no break" power system produced by **TMC Power Distribution, Inc.,** 116 Oronoco St., Alexandria, Va., provides continuous operating power for tele-communications, missile sites, computer centers; any installation where interruption of commercial power would prove detrimental to the mission of the site. The system utilizes a class A design, low slip induction motor with a Delco precise power brushless generator, providing power accuracy within 1% for both voltage and frequency outputs. Constant voltage and frequency output are provided by a low friction inertia flywheel, connected by flexible coupling to the prime source through a magnetic clutch and to the generator.



The inertia flywheel is sized sufficiently to carry the generator under full load condition for ten seconds with less than a two-cycle output frequency drop and with enough reserve power to start the emergency engine through the magnetic clutch. The low friction bearings on the inertia flywheel are engineered for 100,000 hours of operation and contain a bearing failure sensing system to provide early indication of deterioration of the bearing with adequate time to prevent complete failure.

All flexible couplings within the system are very conservatively rated for long trouble-free operation.

The emergency engine is equipped with an automatic hydraulic starting system and PSG isosynchronous governor.

A complete power panel containing appropriate current, voltage, power, power

(Continued on page 124A)

COMPUTERS AND CONTROL SYSTEMS

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- LOGIC DESIGNERS
- PROJECT ENGINEERS
- ADVANCED ELECTRONICS

Scientists and engineers needed with experience in all phases of digital computer and control system design.

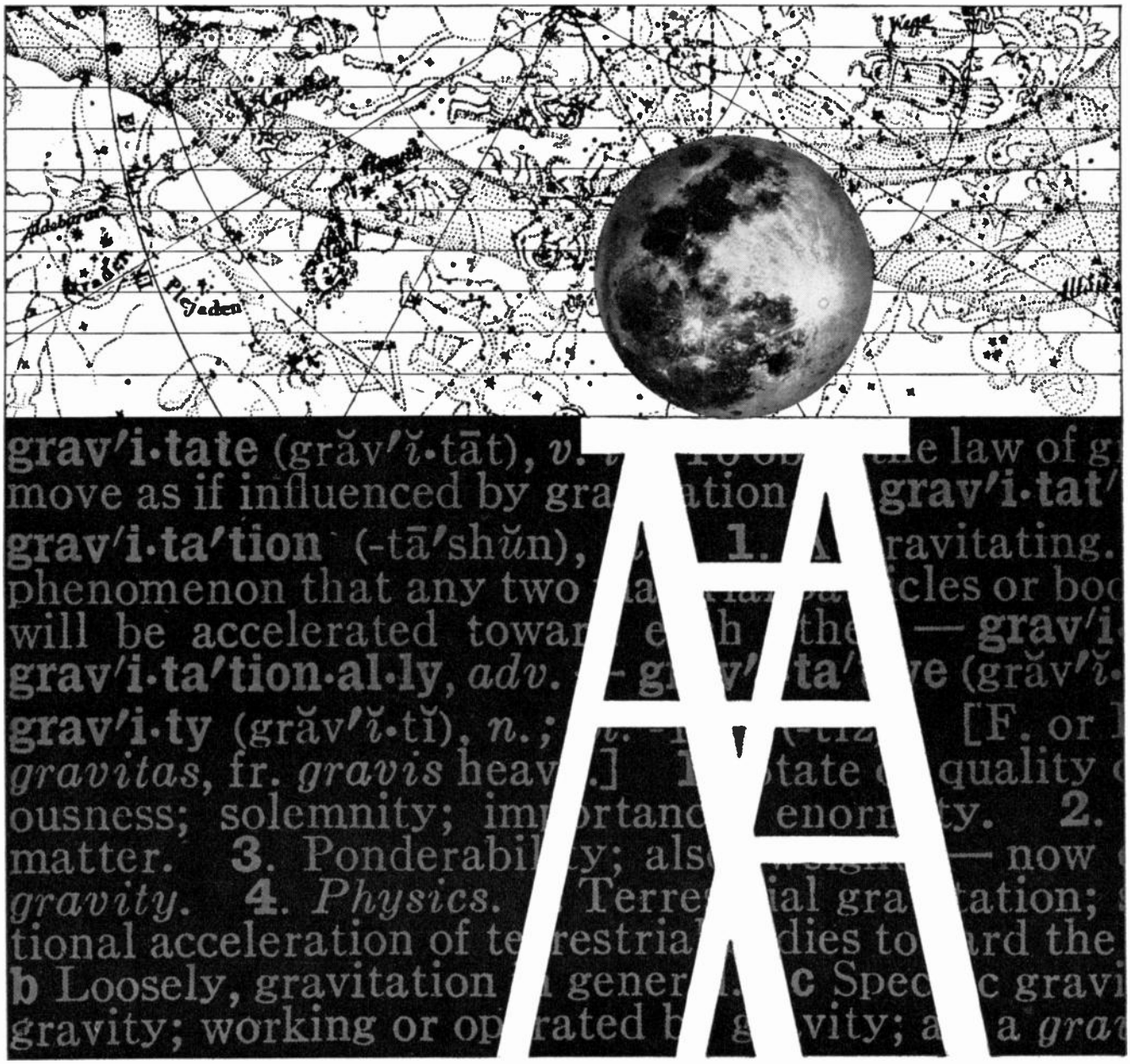
Systems organization, logic design, magnetic core and drum memories, dynamic analysis, and electro-optical correlation devices. Also advanced areas such as high-speed tunnel-diode techniques, thin films, and hybrid analog-digital techniques. Applications include airborne digital equipment, numerical machine control, photogrammetric equipment, and special-purpose control computers. Both commercial and military programs, emphasizing advanced development and research. We think you will find this work unusually stimulating and satisfying. Comfortable and pleasant surroundings in suburban Detroit.

If interested, please write or wire A. Capsalis, Research Laboratories Division, The Bendix Corporation, Southfield, Michigan.

Research Laboratories Division

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ELECTRONICS ENGINEERS
& PHYSICISTS

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Republic's Paul Moore Research & Development Center is the most sophisticated and integrated research complex in the East. Advances made here in many critical aerospace problem areas have brought Republic a diversity of new and follow-on R & D contracts leading to next-generation space and re-entry vehicles, satellites, space power and communication systems.

Electronics Engineers and Physicists are invited to consider the challenging opportunities to make important contributions on these programs in the areas of:

SPACE CRAFT COMMUNICATION — B.S., M.S.
Design and develop space vehicle communications systems including telemetry, command and on-board data handling.

DATA HANDLING (ASGSE) — B.S., M.S.
Design and develop ground station and on-board data handling systems for re-entry and space vehicle applications.

COMMUNICATION TECHNIQUES — B.S., M.S.
Develop advanced communications techniques for aerospace and space craft, includes communications theory and network synthesis.

ADVANCED SPACE RADARS — B.S., M.S.
Develop concepts and components for advanced space radar including rendezvous, mapping, acquisition and tracking applications.

RADAR INTEGRATION — B.S.
Develop specifications, install and integrate advanced radar in hypersonic and space vehicles, including antennas, transmitters, receivers, displays, power supplies, controls.

ANTENNA DESIGN — M.S.
Design and development of antennas for re-entry vehicles. Knowledge of wind effects and general re-entry radiation blackout problems.

RADAR TEST (GSE) — B.S.
Test, checkout and maintain ground radar systems. Make required circuit modifications including range gating circuits, modulators. No travel.

RE-ENTRY INSTRUMENTATION — B.S., M.S.
Design instrumentation for specific re-entry and space vehicles including telemetry systems.

MATHEMATICAL ANALYSIS CONTROLS — PhD.
Theoretical analysis of noise effects and non-linear mechanisms on automatic controls. Includes optimal control theory and generalized stability criteria.

ELECTRONIC INSTRUMENTATION — B.S., M.S.
Develop instrumentation for space vehicles. Knowledge of system integration and telemetry desirable.

SPACE GUIDANCE SYSTEMS — M.S., PhD.
Develop and analyze navigation and guidance systems using inertial and Doppler techniques and advanced nuclear gyros.

ECM REQUIREMENTS — B.S., M.S.
Mathematical analysis of ECM requirements for advanced aerospace and space craft, and specification of equipment.

FLIGHT CONTROL DESIGN — M.S.
Automatic flight controls, servo systems, nonlinear dynamic systems for space craft.

PYROTECHNIC CIRCUIT DESIGN — B.S., M.S.
Develop pyrotechnic missile circuits including safe arm, squib ignition and RFI elimination devices.

EXPERIMENTAL PHYSICIST — PhD.
Conduct experimental studies of the application of nuclear or electron resonance to gyroscopics.

ENVIRONMENTAL TESTING — B.S., M.S.
Undertake test programs to estimate component and system reliability using AGREE type methods; monitor offsite testing.

DESIGN REVIEW — B.S., M.S.
Perform mechanical or electronic design reviews, failure analyses, quantitative analyses and reports. Includes circuit analysis, component selection.

STATISTICAL ANALYSIS — B.S., M.S.
Apply statistical theory and method to prediction and analysis of aerospace component performance.

IDEP PRISM PROCEDURES — B.S., M.S.
Participate in "Interservice Data Exchange Programming" and "Program Reliability Information Systems for Management."

Interested applicants are invited to write in confidence to: Mr. George R. Hickman, Technical Employment Manager, Dept. 14K

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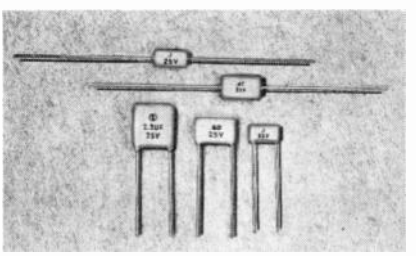
These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your IRE affiliation.

(Continued from page 122A)

factor, running time, and generator bias supply meters (all with 1% accuracy) is provided. In addition, this control panel contains automatic synchronizers, voltage and frequency sensing relays to operate on three phase, and necessary time delay relays for complete monitor and operational control of the system.

Ceramic Capacitors

High reliability and small size are the two outstanding features of **Sprague Electric Co.**, North Adams, Mass., layer-built monolythic capacitors, which utilize N030 ceramic dielectric material.



The new N030 formulation offers greater temperature stability and capacitance per unit volume than heretofore available in compact ceramic capacitors. The close tolerance on temperature coefficient to which the capacitors are held permits these units to be classed as NPO zero temperature coefficient capacitors by existing industry and military standards.

The manufacturing process employed successfully incorporates a multiplicity of layers of very thin ceramic dielectric with alternate metallic electrodes. After the successive layers have been prepared through alternate spraying of dielectric and screening of electrodes, the capacitor chip is fired as a single piece into a solid homogeneous block. Connection lead wires are attached and the unit coated for both moisture and mechanical protection with a touch insulating resin coating.

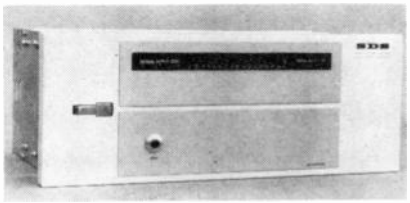
Stepping Switch

A new 36-pole 24 or 32 position removable wafer stepping switch catalog sheet, just released by **Chicago Dynamic Industries, Inc.,** Precision Products Div., 1725 Diversey Blvd., Chicago 14, Ill., contains illustration, technical data, dimensional drawings and electrical specifications of Series RS-40-MD (motor or slosyn drive) and Series RS-40-SD (rotary solenoid) 36-pole 24 or 32 position rotary stepping switch with removable wafers.

Analog-to-Digital Converters

The AD series of analog-to-digital converters produced by **Scientific Data Systems, Inc.,** 1542 Fifteenth St., Santa

Monica, Calif., are high-speed, all silicon semiconductor devices that operate at speeds in excess of 5 microseconds per bit. In addition to enhanced reliability, the use of silicon permits the AD series to operate from 0°C to 100°C.



Transformer coupling is provided for all digital control and output signals in order to eliminate system grounding problems. Compatible multiplexers and sample and hold amplifiers are available.

Significant economy results from the use of standard SDS digital modules—a 14 bit, 0.01%, 14,000 conversion per second converter is priced at $3,700.

Microwave Signal Generators

A series of four microwave signal generators, covering the frequency range of 900 to 11,000 mc, is being introduced by **Empire Devices, Inc.,** Amsterdam, N. Y.



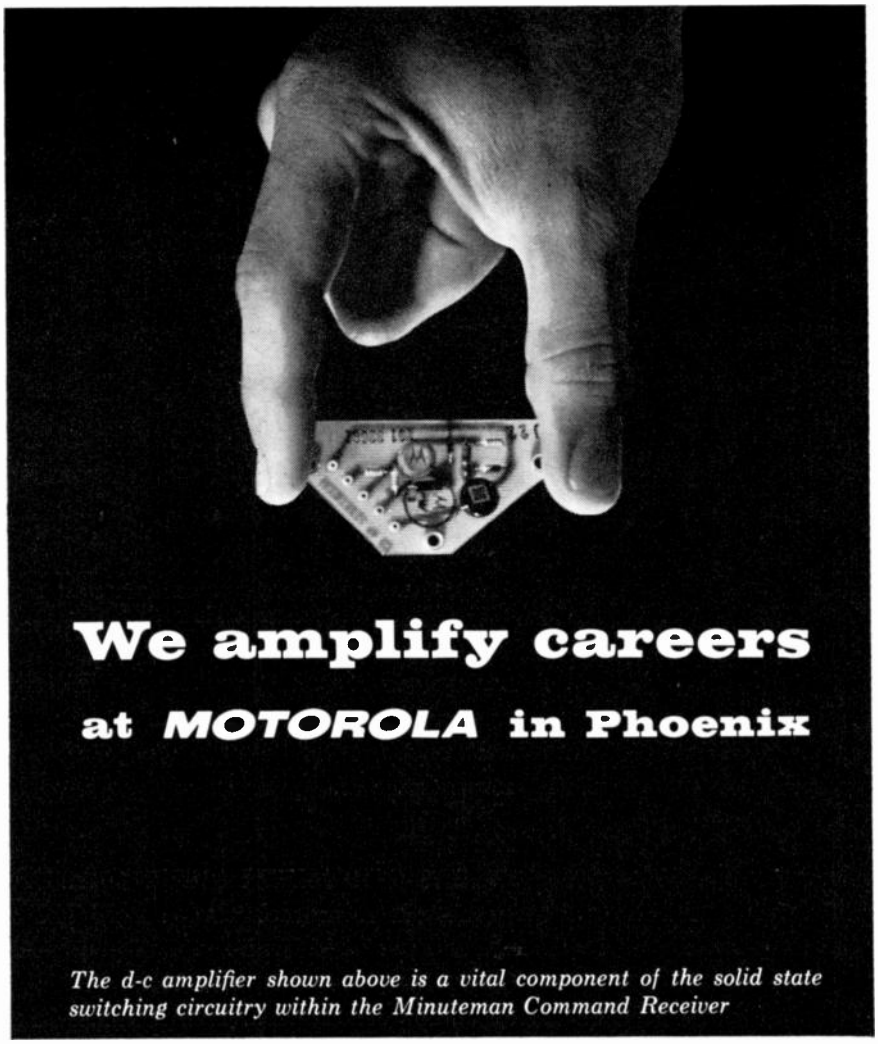
The signal generators, designated Models SG-11 (900 to 2200 mc), SG-12 (1800 to 4400 mc), SG-13 (3800 to 7600 mc) and SG-14 (7000 to 11,000 mc), are designed with complete uniformity of meter and control lay-out in order to simplify operator training and reduce human operating errors.

An outstanding major new achievement of this new line is the attainment of a degree of frequency linearity which permits frequency display in digital form with an accuracy of 0.5% or better.

All models provide internal and external pulse modulation and linear frequency modulation, as well as internal square wave modulation. Output power for each of the units is 1 milliwatt minimum, over an output range of 0 DBM to minus 127 DBM.

Applications include measurement of receiver selectivity, sensitivity, noise figure and spurious responses, determining attenuation and VSWR; measurement of conversion gain; antenna gain and pattern analysis. All construction is thoroughly ruggedized for field use. A special non-contacting short in the klystron cavity eliminates contact noise, provides long operating life and assures a high degree of reliability.

(Continued on page 126A)



We amplify careers at MOTOROLA in Phoenix

The d-c amplifier shown above is a vital component of the solid state switching circuitry within the Minuteman Command Receiver

Engineers discover, after joining the Military Electronics Division of Motorola in Phoenix, that they have a new-found enthusiasm for their work and a fresh sense of accomplishment. That's because all professional personnel are individually selected and then assigned to challenging state-of-the-art projects which fully utilize their training, experience, and creativity. We can thus provide greater career opportunities for our engineers and also broaden Motorola's capabilities as a leader in the field of advanced military electronics.

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Reliability Program Coordination
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Data Acquisition, Processing and Display
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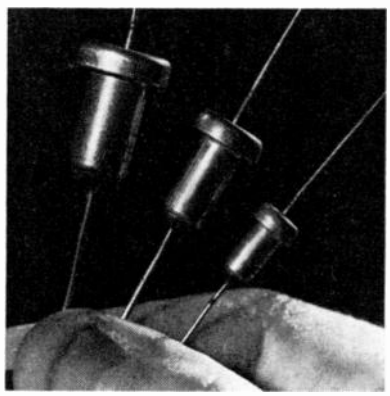
These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your IRE affiliation.

(Continued from page 125A)

Complete details are available in a new bulletin, Catalog 614, Supplement A, which may be had upon request to the firm.

MIL-C-3965 Capacitors

Types CL44 and CL45 tantalum capacitors conforming to Characteristic C of Military Specification MIL-C-3965 have been introduced by **Fansteel Metallurgical Corp.**, North Chicago, Ill.



To conform to Characteristic C, Fansteel's Rectifier-Capacitor Division has expanded its capabilities to produce capacitors conforming to all voltages, vibration grades, capacitance values, and characteristics listed for Type CL44 and CL45 capacitors.

Capacitors of this type are used for military missiles and spacecraft. Characteristic C specifies ratings of 1.7 to 560μ, 6 to 85 volts, with a temperature range of -55°C to 125°C.

Fansteel's patented leak-proof, crimped seal used in these capacitors will withstand the low atmospheric pressure of space environments.

PNP Silicon Power Transistors

PNP silicon power transistors in 75 and 85 watt ranges were shown at WESCON by **Silicon Transistor Corp.**, Carle Place, L. I., N. Y.



PNP silicon power transistors will enable design engineers to employ complementary

(Continued on page 128A)

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At ITT Federal Laboratories engineers and scientists participate not only in advancing communications theory and technology but also in the applications of these new techniques to systems of large scope and major importance. To cite a few instances: **ACE HIGH** — the command communication network of tropospheric and line-of-sight radio stations linking NATO countries ☐ **465-L** — the vast-scale Command and Control system for SAC, for which Federal Labs is doing basic R&D work in cooperation with other ITT Divisions ☐ **UNICOM** — for which ITTFL is designing the store and forward portion of a comprehensive world-wide digital communication system for the U.S. Army Signal Corps ☐ **OTHERS** — include complete development of the ground equipment for the Courier communication satellite, participation as a ground terminal for the NASA Relay satellite program ☐ Current assignments open at Federal Laboratories in Nutley, N. J., reach across the spectrum of research, development, design and test engineering; they also include systems development and program management. An unusual array of R&D facilities includes a 40-foot radio telescope used for probing space, moon-relay links and other communications tests ... and a 90-mile tropo scatter link from the Laboratories in Nutley to Southampton, Long Island ☐ **IMMEDIATE OPENINGS: Project Engineers** — project leadership, proposal preparation in the 2-, 5-, and 8-KMC bands on microwave communication equipment, solid state transmitters, receivers and modulators ☐ **Tropo Scatter Communications** — design and development engineers skilled in radio or wire line communication equipment for voice and data transmission equipment. Design experience in microwave antennas, transmission lines, filters, IF amplifiers, multipliers, parametric amplifiers and varactors. Versed in telephone transmission standards and tropospheric scatter propagation ☐ **Computer Engineers** — development of digital data recording equipment. Competence in logic designs, circuit designs, packaging and testing of A/D converters, servomechanisms, magnetic and optical recording ☐ **Space Communications** — studies and development of space craft telemetry and control equipment, RF circuits for transmitters and receivers, systems integration and instrumentation ☐ **In the broad view,** communications programs represent only a part of ITTFL activities. The entire R&D Capability is functionally organized into 17 laboratories engaged in missile and space systems, avionics, communications, electronic defense, and research in the physical sciences. At the Nutley facility, many programs are carried from R&D through design, to manufacture and installation ☐ To apply, or gain more information, forward your resume in full confidence to Mr. T. Allen, Box 86-MA, ITT Federal Laboratories, 500 Washington Avenue, Nutley, N. J. (An Equal Opportunity Employer)

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Rerouting Electronic Currents in Solids

Information processing requires the control of current paths through solids. A number of established technologies can be used to affect this control. The IBM Thomas J. Watson Research Center is taking a new look at this problem to find the physical interactions which can deliver this control with the highest possible speeds allowed by laws of nature.

This work will offer inventive people an opportunity to make really important advances. Inquiries are welcome from theoreticians as well as from those whose main concern is with problems of materials in making exceedingly small but well-controlled exploratory structures. IBM is an Equal Opportunity Employer. Applicants should possess an advanced degree or equivalent experience.

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(Continued from page 126A)

circuit push-pull output stages, eliminating an input and output transformer and several resistors and diodes. They will also make possible a direct replacement of silicon for germanium power transistors without redesign in most circuits. STC is also offering a complementary series of NPN transistors which will enable design engineers to "plug-in" a complementary pair of silicon power transistors.

Now available are 85 watt PNP devices in the square package—(2P389, 2P389A, 2P424, 2P424A). Characteristics of these units are: hFE-12 to 60 @ 1 ampere; RCE (saturation) 0.75 ohms @ 1 ampere; VCE 80 volts; IC maximum 3 amperes.

Also available are 85 watt PNP devices in the 11/16" hex package (STC5550 through STC5555) and 75 watt PNP units in the TO-3 package (STC5080 through STC5085). Characteristics of these devices are: hFE 10 to 30 @ 2 amperes; RCE (saturation) 0.5 ohms @ 2 amperes; VCE 80 volts; IC maximum 5 amperes.

Free-Gyro System

Miniaturized roll-stabilized free-gyro system developed by **Whittaker Controls and Guidance Div., Telecomputing Corp.,** 16217 Lindbergh St., Van Nuys, Calif., provides pitch, roll and yaw attitude information for telemetry or control of spinning



rocket vehicles. This system also adapts to spinning orbital vehicles and nose-cone reentry studies for either control or telemetry. Two two-degree-of-freedom gyros are mounted on a roll-attitude stabilized platform. One gyro senses and corrects roll position errors of the platform through a servo system and provides yaw attitude information. The second measures pitch attitude. Potentiometer pickoffs provide convenient data output. The system's dynamic performance characteristics are reflected in its ability to withstand accelerations of 50 g's along roll axis and simultaneous roll-spin rates in excess of 180 rpm. Operational freedom angles are: pitch 360°; roll 360°; and yaw ±85°. Maximum gyro drift does not exceed 0.5°/min. The 11-pound system contains an integral dc to ac inverter, servo electronics and caging mechanism, and is packaged in a cylinder 7 inches long and 5 inches in diameter. Power consumption is 1.5 amperes maximum at 28 vdc.

(Continued on page 130A)

Growth opportunities for communications engineers

Expanding activity at Hughes Communications Division is concerned with the definition and solution of global civilian and military communications requirements of the 1965-75 era. Immediate professional assignments exist for qualified Communications Engineers in the following project areas:

- SYNCOM I & II Studies
- 480L USAF Air Communication System Studies
- Company-sponsored Communication Satellite Ground Terminals
- Manpack Transceiver
- Air/Ground Digital Data Communication Systems
- Tactical Vocoder & Speech Processing Equipment
- Combat Area Communication Systems
- Microwave Synthesizers and Millimeter Wave Communication Equipment
- LASER Communications
- High-Power Transmitters
- Modem Developments
- Secure Data & Command Systems

If you are qualified by interest and experience, hold an accredited degree and U.S. Citizenship and would like to associate with one of Southern California's largest and fastest-growing aerospace firms, we would like to talk with you.

Please airmail your resumé or call Southern California (collect). Contact:

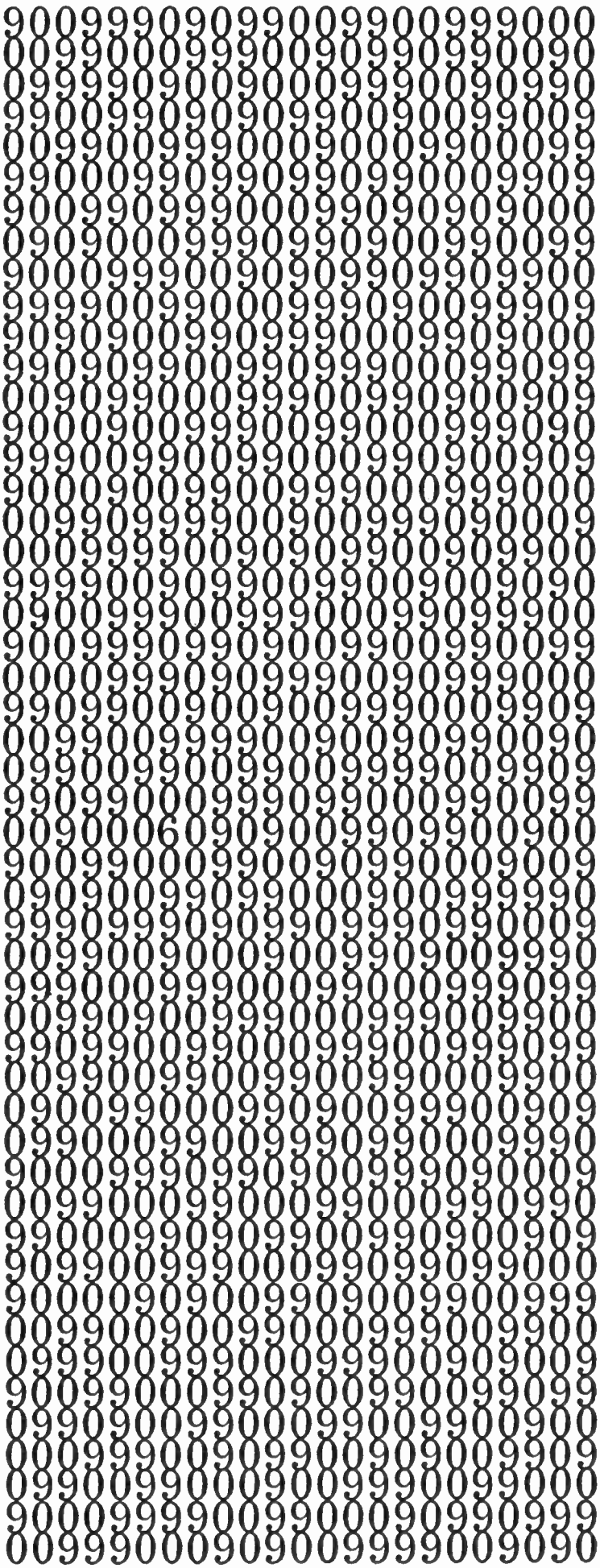
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RECONNAISSANCE SYSTEMS—*programs at Sylvania-West offer career opportunities on the San Francisco Peninsula in reconnaissance systems engineering, including feasibility studies, operational analysis, system synthesis and performance analysis, application and development of advanced techniques, specification of constituent subsystems, human factors engineering and operations research. Typical areas in which openings currently exist include the following:*

Data Analysis and Signal Processing. Carry out studies leading to the design of new, or the refinement of existing data collection systems. Studies range from the analysis required to develop and define operating requirements compatible with economic and state-of-the-art considerations, through systems conceptual design in block diagram form, including specification of hardware design approach.

Statistical Analysis. Perform statistical analyses, set up mathematical models of stochastic processes. Fields include experimental design, theory of queues, theory of mixtures, allocation of resources, search theory and general operations research.

Intercept and Detection. Direct or perform operational and technical requirements studies; electronic signal environment studies, synthesis of electronic intercept systems from conception to hardware specification and system block diagram; analysis of system performance and data related to telemetry, communications, radar and others. Direct or prepare reports and proposals and maintain technical contact with customer representatives.

Antenna and Propagation. Perform analyses of electromagnetic propagation aspects of reconnaissance and other systems; analyze direction finding problems and develop direction finding techniques; determine antenna requirements and configuration during synthesis of reconnaissance systems. Activities include report writing and customer contacts.

These openings exist at all experience levels. Advanced degrees in EE, physics, or mathematics desirable.

SYLVANIA ELECTRONIC SYSTEMS
Government Systems Management
for GENERAL TELEPHONE & ELECTRONICS

Complete information may be obtained by writing, in confidence to

Roger Harlan
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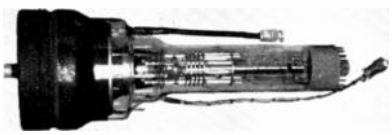
NEWS New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your IRE affiliation.

(Continued from page 128A)

Barrier-Grid Storage CRT

A barrier-grid storage cathode-ray tube, Type FW-202, having wide application in information storage, was announced by the **International Telephone and Telegraph Corp., ITT Industrial Laboratories Div.,** 3700 E. Pontiac St., Ft. Wayne, Ind.



The tube may be used as a radar moving-target indicator, as a random-access binary information store, and as an image-storage device.

The 5-inch tube contains an accurate electron gun and a sandwiched target consisting of a metallic backplate, a dielectric layer, and a metallic barrier-grid matrix of square apertures. In the scanning process, the electron beam passes through the grid and strikes the dielectric layer. The layer emits secondary electrons that pass back through the barrier-grid apertures to a collector electrode, causing the tube to emit a signal.

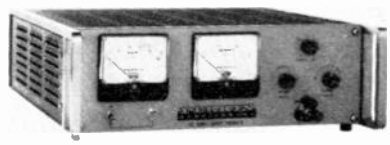
Depending on the charges left on the elemental dielectric areas and the potential on the barrier grid, the tube can reach a state of electric equilibrium. Under this condition, no secondaries pass through the barrier-grid apertures to the collector and the tube emits no signal.

When operated as a clutter cancellation device in a radar moving-target-indicator system, the radar video signal is applied to the backplate in synchronism with the scanning beam. After a few scanning cycles, the repetitive input signal from a stationary or clutter target causes the tube to reach the condition of equilibrium and outputs thereafter occur only for moving targets. A cancellation ratio of 20/1 (26 decibels) is achieved for clutter.

The tube, when used for random-access binary information storage, can store a 240-by-240 bit raster for a capacity of 58,000 bits, due to its fine spot size. The resolution is 1200 television lines per diameter, 15 writing levels are distinguishable in the output, and shading is negligible.

5 Ampere Supply

Perkin Electronics Corp., 345 Kansas St., El Segundo, Calif., has added a 5-ampere dc model to a line of dc power supplies featuring voltage regulation, remote voltage programming, full range current limiting (0–6 ampere) series and/or parallel operation, remote sensing, overload and short circuit protection and transient-free operation. This unit, model number TVRO40-5, has an output of 0–40 volts at 5 amperes and is transient-free.



The major specifications are 0.01% static regulation, 1 millivolt or less ripple, 50 microseconds or less response time and 40 pound weight. The package is a 17″ by 5¼″ by 15″ bench mount type enclosure and is furnished with adaptors for mounting in a standard 19″ wide by 5¼″ high rack space.

Complete additional data are available from the manufacturer.

Analog to Digital Converter

New smaller, simpler and faster airborne models of **Towson Laboratories, Inc.,** 200 E. Joppa Rd., Baltimore 4, Md., capacitive charge transfer analog to digital converter are now being delivered. The units are capable of wide temperature operation and other environmental conditions typical of MIL applications. The units offer high input impedance, sampling aperture of 1.3 microseconds, ±2.5 volts full scale, 10 bit resolution plus odd count parity bit, 23,000 encodings or more per second, serial binary NRZ

(Continued on page 132A)

ADVANCED SYSTEMS RESEARCH 2 ADVANCED SYSTEMS ENGINEERING

NEW DEPARTMENTS
NOW BEING ESTABLISHED AT BELL AEROSYSTEMS

Known for such "firsts" as HIPERNAS, the most accurate, pure inertial, self-compensating navigation and guidance system known, Bell is increasing its Company-sponsored R&D and further expanding its programs in Avionics. Interesting new investigations are about to start in such varied areas as controls for a lunar landing "bug" or module, learning machines, development of advanced mathematical theorems for quick, accurate measurement of reliability, and new concepts in battlefield surveillance systems and air traffic control.

POSITIONS ARE IMMEDIATELY AVAILABLE FOR:

SR. PHYSICISTS & ELECTRONIC ENGINEERS
MS or PhD with 5 years experience in physics, radar, optics and guidance. To work in company-sponsored R&D programs on microminiaturization, lasers, learning machines, new guidance and control techniques for space vehicles.

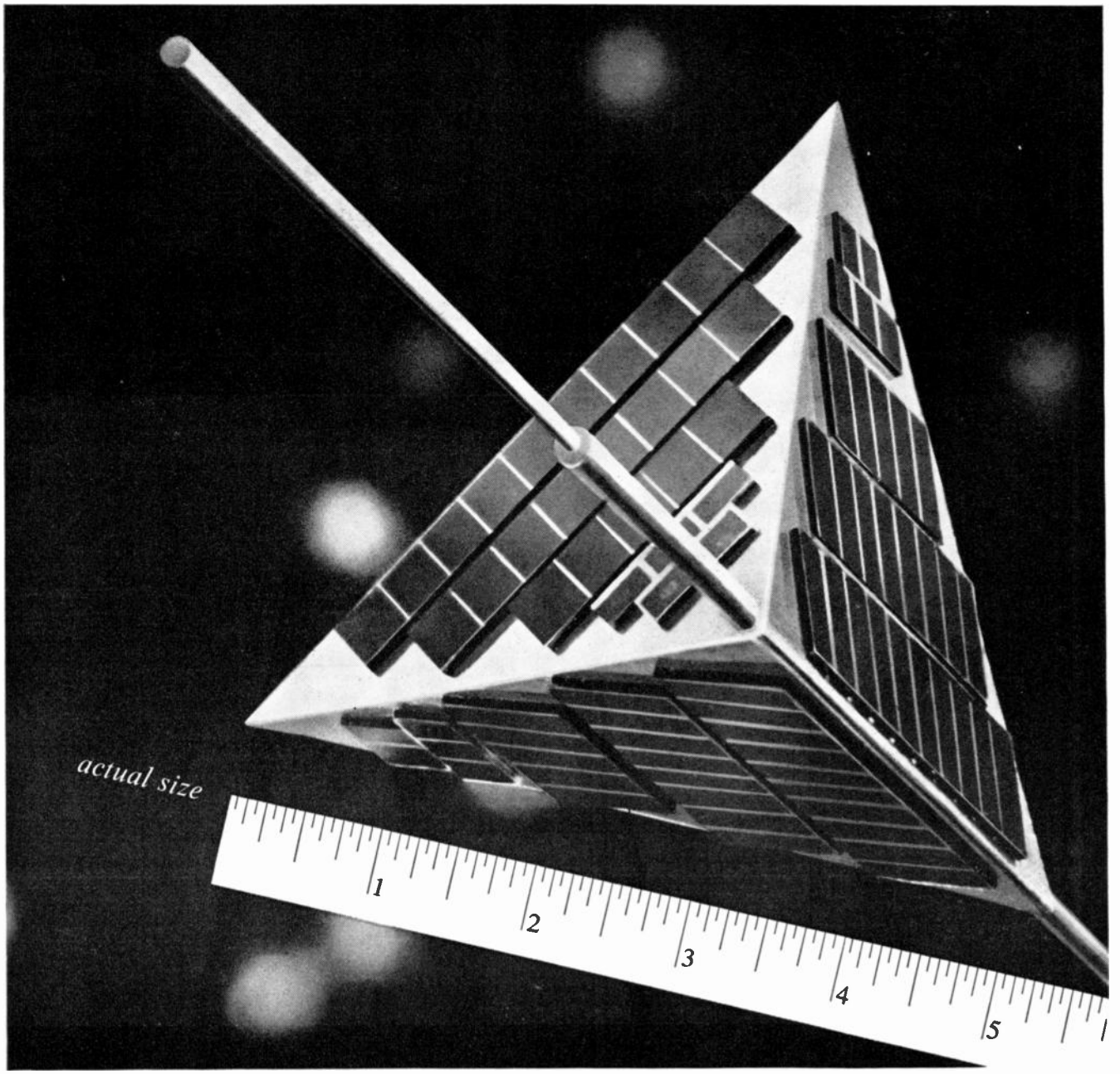
SR. ELECTRONIC SYSTEMS ENGINEERS
BS to PhD in EE with minimum 5 years experience. For synthesis and analysis work involving air traffic control, target locators, space, radar, command and control, visual simulators and feedback control systems.

CONTROL SYSTEMS & SIMULATION ENGINEERS
Supervisory and senior level positions for assignment to automatic feedback control systems, aerospace vehicles, automatic throttle systems, control systems for simulators and autopilots, closed circuit TV and circuit design. BS or advanced degree in EE plus 3 to 8 years analytical experience related to servo analysis involving hydraulic and electronic control systems, feedback theory, analog computers, simulators and autopilots.

SR. ENGINEERS & PROJECT MANAGERS
BS or advanced degree in EE plus 10 years analytical experience. Radar and antenna experience desirable. For development of pulse command, navigation and data link systems with specific assignment on such projects as nano-second pulse transmitters and receivers, precision radio distance measuring equipment, rapid frequency diversity techniques, and increasing the power, range and sensitivity of radio command and communication systems.

Your inquiries are invited. Please send resume to Mr. T. C. Fritschi, Dept. F2.

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DIVISION OF BELL AEROSPACE CORPORATION-A textron COMPANY
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The world's smallest satellite has been developed by Space Technology Laboratories. Its shape will be different from all other satellites before it. STL engineers and scientists have used a tetrahedral configuration to bring about some remarkable characteristics in a space vehicle. There will be no need for batteries nor regulators in flight. The satellite will have no hot side, no cold side. It will require no attitude control devices. No matter how it tumbles in space it will always turn one side toward the sun to absorb energy, and three sides away from the sun to cool instrumentation and telemetry equipment inside. It can perform isolated experiments in conjunction with other projects. Or it can be put into orbit by a small rocket to make studies of its own, up to five or more separate experiments on each mission it makes.

STL is active on hardware projects such as this and as prime contractor for NASA's OGO and an entirely new series of classified spacecraft for Air Force – ARPA. We continue Systems Management for the Air Force's Atlas, Titan and Minuteman programs. These activities create immediate opportunities in: Space Physics, Radar Systems, Applied Mathematics, Space Communications, Antennas and Microwaves, Analog Computers, Computer Design, Digital Computers, Guidance and Navigation, Electromechanical Devices, Engineering Mechanics, Propulsion Systems, Materials Research. For So. California or Cape Canaveral opportunities, please write Dr. R. C. Potter, Dept. I-10, One Space Park, Redondo Beach, California, or P. O. Box 4277, Patrick AFB, Florida. STL is an equal opportunity employer.

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PROFESSIONAL CAREER PROGRESS REPORT NO. 1

by JOE WATSON, Project Engineer, SPACE DEPARTMENT, APPARATUS DIVISION



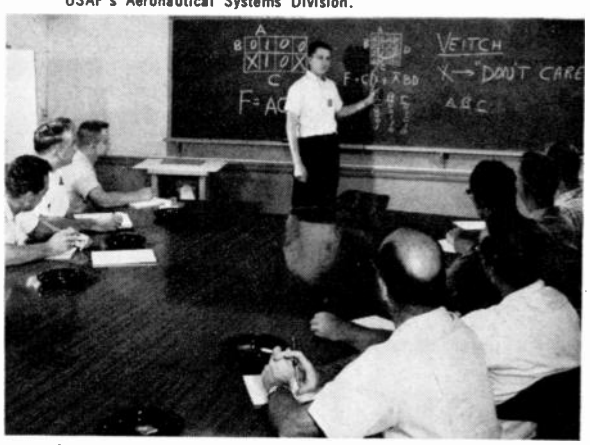
Joe Watson's integrated circuit applications work contributed materially to the design of this tiny computer (total volume only 6.3 cubic inches and weighing just 10 ounces) developed for USAF's Aeronautical Systems Division.

"My current assignment at Texas Instruments is the development of applications for integrated circuits — logic design of digital systems by mathematical techniques — and interface circuit design. In this stimulating field, TI allows me unusual **creative freedom.** And I'm working in an atmosphere of accomplishment with a great team of engineers and scientists. The solid background of experience they have in space telecommunication and control systems is a real asset to my work.

THREE ADVANCEMENTS IN TWO YEARS

"I've worked hard — and enjoyed it. This is my third major assignment since joining TI early in 1960, each an advancement in **responsibility and recognition.** My own progress, and my observations of TI associates, support TI's claim that **recognition and advancement come rapidly and in direct proportion to effort and ingenuity devoted to an assignment.**

"I earned my BS in electrical engineering at Oklahoma State in 1959 and my MS in 1960. For my thesis on digital systems design I built a small transistorized desk calculator as a practical demonstration. I found a genuine interest and an **intelligent analysis of my capabilities** when I applied at Texas Instruments. And I've found that this same attitude continues as I pursue my projects. Picking TI over all other position offers in 1960 was a really happy choice.



Three times each week, Joe Watson conducts special classes for some 30 TI engineers to keep them current with the State of the Art in logic design of digital systems and to demonstrate how circuits may be utilized in other systems being developed by TI's Apparatus division.

OPPORTUNITIES AT ALL LEVELS

"In my opinion, any engineer or scientist — junior or senior — will do well to investigate a career with TI, whether to satisfy an immediate personal need for a change, or as an exploratory step in a long-range professional advancement program.

"Some immediate openings in our SPACE DEPARTMENT are listed at left. This is just one of many TI departments where your talents might be **profitably** employed."

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Professional Openings in:

- ☐ Space Communications
- ☐ Telemetry Systems
- ☐ Ground Data Processing
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- ☐ Digital & Analog Circuit Design
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specific openings in your professional area or write for brochure giving the broad picture of TI's many-faceted activities and **challenging professional openings.**



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NEWS New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your IRE affiliation.

(Continued from page 130A)

output code format. Provision is made for external clock with command encoding. Power consumption is 5 watts or less. Reliability is enhanced by small number of components, approximately 425 or less. Units are available on a 60 day delivery basis.



Two models are presently available. The model OC-1001 is constructed to meet the requirements of MIL-E-5272C including vibration up to 2,000 cps and through the use of thermostatically controlled heaters ambient temperature range of −54°C to 71°C. It is approximately $3\frac{3}{4}''\times6''\times5''$ and weighs 3.5 pounds.

The OC-1002 is a sub-assembly model made up of plug-in printed circuit boards convenient for installation in customer's equipment. It is calibrated for operation from +10°C to +80°C. It is approximately $3''\times4\frac{3}{8}''\times4\frac{1}{4}''$ and weights 1.2 pounds.

For further information contact the firm.

Variable Filter

A solid-state, constant-percentage bandwidth variable filter with a frequency range from 8 to 2500 cps has been introduced by **Consolidated Electrodynamics Corp.,** a subsidiary of **Bell & Howell,** 360 Sierra Madre Villa, Pasadena, Calif.



Designated Type 1-159, the instrument features a 3% bandwidth at the 3 db point and an attenuation of 55 db one octave either side of the tuned frequency.

When used with CEC's Type 1-117 vibration meter, the combined units constitute an effective vibration analyzer for moni-

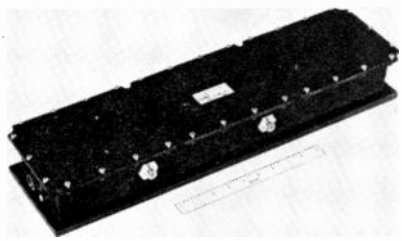
toring aircraft turbine engines, motors, generators, pumps, and so forth. A times-ten amplification permits accurate reading of low level vibrations at higher frequencies.

Sharp selectivity has been attained through the use of two cascaded twin "T" amplifiers. Supplementary feedback assures unity gain at peak response over the tuning range.

A dc battery power supply is available as an accessory. Conversion from ac to dc operation is accomplished by a change of plug-in power supplies.

Microwave Amplifier

The WJ-115, a space communications amplifier package for the 2.0–2.4 Gc band is now available from **Watkins-Johnson Co.**, 3333 Hillview Ave., Palo Alto, Calif. It employs a WJ-227 12-watt traveling-wave tube together with a solid-state supply. Enclosed in a sealed, pressurized enclosure, the unit operates under 30 g shock and 10 g vibration conditions from temperatures of −35°C.



With a drive signal of 3–6 mw, the unit delivers a nominal output power of 12 watts, with a guaranteed minimum of 10 watts over the band. DC input power required is 81 watts maximum. Probability of survival for 1 year in orbit is 90 per cent at an average baseplate temperature of 25°C and 33 per cent duty cycle. The unit may be operated at 100 per cent duty if desired. Monitoring circuits are provided for internal pressure, collector temperature, helix voltage, helix current and collector current.

Prices and delivery available on request. Inquiries may be address to Applications Engineering Dept., at the firm.

Pulse Generator



A new transistorized pulse generator, designated TPG-1A, manufactured by **Intercontinental Instruments, Inc.**, 123 Gazza Blvd., Farmingdale, L. I., N. Y.,

(Continued on page 134A)

CRITICAL PHASE
IN THE EVOLUTION OF AN OLD SCIENCE
... WAR

As the art of war evolves, the time available for command decision shrinks and the consequences of decision become greater. Modern warfare places a whole new set of technical requirements on military systems at every level . . . from weapons and sensors to the highest levels of command.

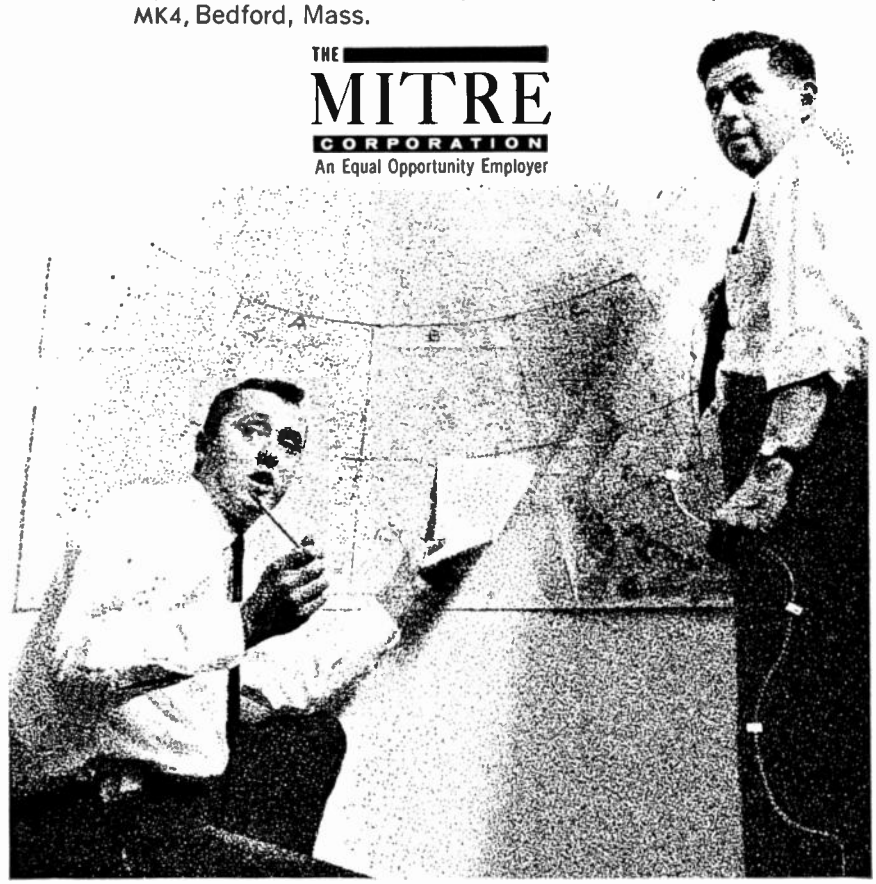
MITRE's prime assignment is to meet the challenge of these problems . . . the challenge of command and control. We are looking for talented systems men to help us.

As a systems man at MITRE, you might perform feasibility studies on new techniques in system design; or process intelligence data with large scale computers; or work on space hardware analysis. These are just a few of many work areas at MITRE. All have a direct bearing on this country's defense.

MITRE is located in pleasant, suburban Boston. Rewards are competitive. Requirements, B.S., M.S., or Ph.D. in these disciplines — electronics, physics, and mathematics.

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Invitation

to

ENGINEERS

AND

SCIENTISTS

to investigate challenging assignments in Melpar's Aerospace, Research, and Engineering Divisions. Assignments range from basic research in solids and films to the design and development of exotic scientific equipment for extraterrestrial exploration.

ADVANCED CIRCUIT DEVELOPMENT SPECIALISTS

Task requires an engineer with several years experience to design advanced circuits utilizing transistors and other solid state components. Work involves basic research on circuit synthesis and will include development of analytical techniques for applying phenomena observed in films and surfaces to circuit application. An advanced degree in Electrical Engineering desired.

CIRCUIT DESIGN SPECIALISTS

Task requires an engineer with several years experience to design and develop solid state circuits for use in new scientific equipment. This work will require close association with others working in such disciplines as optics, chemistry, automatic control theory, and mechanical design.

Write in confidence to:

John Haverfield
Manager,
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NEWS New Products

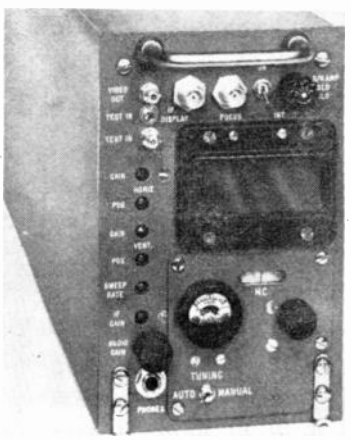
These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your IRE affiliation.

(Continued from page 133A)

provides square waves, and positive and negative pulses from 1 cps to 12 mc. Width variation is 30 nanoseconds to 0.25 seconds. Rise and fall time is 6 nanoseconds. Two outputs provide 0 to +5 v into 50 ohms and 0 to −5 v, with independent amplitude adjustment. The TPG-1A can be synchronized to 12 mc or triggered with external sine wave or pulse to 12 mc. External electronic gate can turn internal oscillator on and off to provide a variable pulse train. Unit is 8″×9″×14″. Price—$845 fob the plant.

VHF Receiver With Electronic Tuning and Display

A compact, solid-state VHF receiver featuring electronic tuning and a built-in display has been introduced by the western operation of **Sylvania Electronic Systems, a division of Sylvania Electric Products Inc.**, Mountain View, Calif.



Designated Model RC-12, the receiver can be automatically or manually tuned over a 1.5:1 frequency range anywhere from 50 to 225 mc, with a maximum noise figure of 6 db. A 60 db dynamic range will not over-load the logarithmic IF amplifier. Linear IF with AGC and FM detection can also be provided in a wide range of selectivities. The built-in 1½ by three-inch cathode-ray display has an adjustable time base for a triggered A-scan. A panoramic presentation can be obtained through an external sweep source. Separate audio and video outputs are furnished.

Extensive use of interchangeable subassemblies provides adaptability to system requirements while insuring against obsolescence. Any power frequency from 50 to 500 cps may be used, since the receiver is suitable to both aircraft and surface installation. Construction is to military specifications. More detailed information may be obtained from Sylvania Electronic Systems-West, P.O. Box 188, Mountain View, Calif.

Voltage Variable Capacitors

Improved voltage variable capacitors having a minimum Q of 25 at 50 mc are being manufactured by **Eastron Corp.**, 25 Locust St., Haverhill, Mass. Available in capacities of 7 to 100 pf measured at 4 volts, these devices provide the Q necessary for tuned circuits operating in the 50 to 250 mc range.

Manufactured in the industry standard subminiature glass package, these VVC's can directly replace conventional low Q units, provide guaranteed low leakage, with ratings to 60 volts, and feature a maximum rate of change of capacitance with voltage, resulting in tuning ratios of greater than 6 to 1.

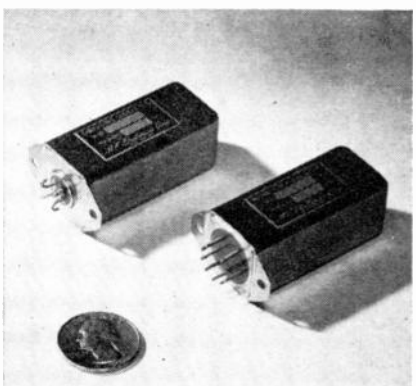
This series (VC207 through VC299) is designed for harmonic generators, remote tuned circuits, variable delay lines, spectrum analyzers, FM telemetry, and automatic frequency control.

Eastron offers special matching and selection services for critical applications. Priced at $2.95 each in 100 lot quantitites. They are available within 10 days on receipt of order.

Complete technical data and applications information are available on request.

Audio Oscillators

MF Electronics Corp., 118 E. 25th St., New York 10, N. Y., has announced the addition to their line of audio tone oscillators, two new series designed for use under shock and vibration, and designated the "R" and "S" series.

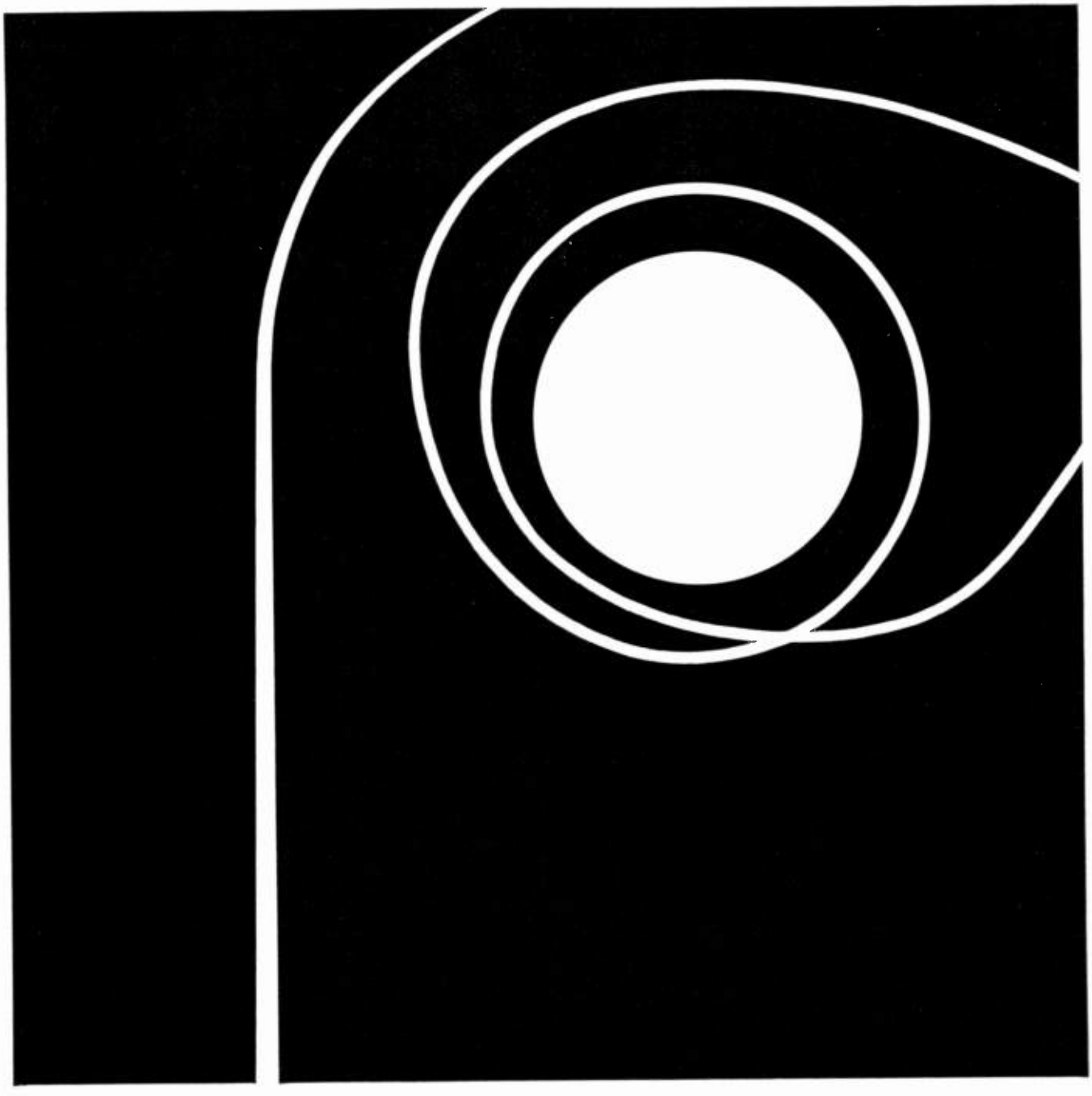


These are transistorized packages of 1½ to 2 cubic inches which generate low distortion sine waves at any specified frequency in the range of 50 cps to 80 kc. Input is 12 or 28 vdc, while output is 2 volts RMS into 600 ohms. Operating temperatures are from −25 to +55°C.

The "R" and "S" series use an AGC system for output stability, replacing the conventional thermal element usually used for this purpose. This results in a generator with ability to sustain shock and vibration, without danger of malfunction due to failure of the thermal element.

These oscillators are currently being used for test signals in telemetry and in test sets. The "R" and "S" series should be of interest to designers of military equipment, due to the shock resistant feature. Details on request to the firm.

(Continued on page 136A)



CHALLENGE FROM THE GROUND UP

At Philco Western Development Laboratories, specialists in communications sciences, space sciences and mathematical analysis begin each new space age program with analysis of mission requirements. Advance techniques in communications and control are applied to development of system demands, specifications for systems and sub-systems, and development specifications. These program-oriented efforts lead to development fabrication and test, and, finally, to installation, checkout and operation of equipment. Career opportunities for holders of B.S. or advanced degrees (electronics, mathematics, physics) offer start-to-finish challenge at Philco WDL, whose achievements include work on Discoverer, Midas, Advent and Courier programs.

Write in confidence for information on how you can find your career at Philco WDL, with the additional rewards of ideal living on the San Francisco Peninsula and professional and monetary advancement commensurate with your own ability. Requirements include U. S. Citizenship or current transferable Department of Defense clearance. Address Mr. Patrick Manning, Department R-10.

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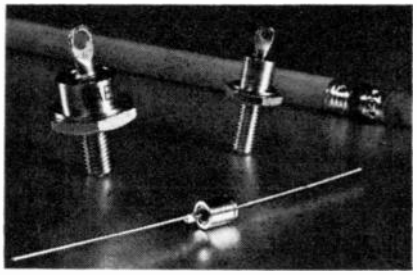
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NEWS New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your IRE affiliation.

(Continued from page 134A)

Regulated Diodes



North American Electronics, Inc., 71 Linden St., West Lynn, Mass., has introduced a new line of regulated avalanche diodes. Called High-*Q* diodes, these new NAE types are the result of an extensive development program to produce diodes and rectifiers with unsurpassed junction perfection and surface stability. Each diode is assembled and packaged in ultra-dry ambients. They are then subjected to 100 per cent thermal shocking (−65 to +200°C), leak detecting, and 200°C storage and load testing to insure the highest reliability. The line is available in 1-, 6-, 12-, and 20 ampere ranges. Write to Frank Mitchell at the firm for technical bulletin.

Differential DC Modulator

The Diehl Manufacturing Co., Finderne Plant, Somerville, N. J., has developed a new modulator for use with its line of 75 watt vacuum tube servo amplifiers. The new modulator converts differential dc signals to 60 cps ac. Gain is 1 volt rms per volt of dc unbalance at any common input voltage from −45 to +45 volts.



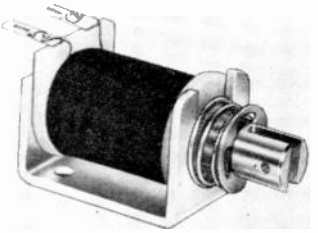
The modulator operates by applying the two dc inputs to the two grids of a differential amplifier. The differential signal is fed through a mechanical chopper and is then filtered by a capacitor loaded "bridged T" filter which is tuned to "notch" at the third harmonic.

The modulator, Model MAM-03, is contained in a module 2 in. long×2 in. wide×4⅝ in. high. It plugs directly into a socket on the amplifier chassis.

For complete information write to the firm.

Miniature Solenoid

Comar Electric Co., 3349 W. Addison St., Chicago 18, Ill., manufacturers of relays, solenoids, switches and coils, has recently announced a new patented, miniature solenoid identified as Type S1-10.



For AC operation only, the new solid frame solenoid is said to be a low cost, compact unit that permits easier installation where limited space is a factor. Available for continuous or intermittent duty, up to 220 volt ac, the electrical and physical characteristics are as follows: Size, $1\frac{7}{16}''\times\frac{3}{4}''\times\frac{13}{16}''$; Intermittent duty, 110 vac, stroke $\frac{1}{16}''$ to $\frac{1}{2}''$, pull 6 to 31 oz; 90 vac, stroke $\frac{1}{16}''$ to $\frac{1}{2}''$, pull 4 to 24 oz; Continuous duty, 110 vac; stroke $\frac{1}{32}''$ to $\frac{7}{16}''$, pull 1 to 16 oz; 90 vac, stroke $\frac{1}{32}''$ to $\frac{7}{16}''$, pull to 9 oz.

Complete details and prices are available from the firm.



Crystal Growing Arc Imaging Furnace

To open areas of crystal growing research beyond the capabilities of previously available systems, **Arthur D. Little, Inc.**, 20 Acorn Park, Cambridge 40, Mass., has introduced a new model of the ADL-Strong Arc Imaging Furnace. The crystal growing research model provides a completely non-contaminating radiant heat source for operations up to 4000°K and permits selection of inert, oxidizing or reducing atmospheres in pressures ranging from ten microns to sixty psi.



Essentially the unit consists of two high-intensity Strong Electric Corp. "Jet-arc" arc sources which are used alternately to provide continuous operation; an optical system for projecting and re-imaging the radiant energy of the arc source; a continuously adjustable Verneuil powder feed system; and a rotating platform upon which the boule is grown. The nozzle of Verneuil powder feed system and the boule platform are enclosed in a quartz chamber to provide complete environmental control.

The radiant energy of the arc source is projected by a 21″ ellipsoidal mirror and re-imaged by a 24″ ellipsoidal mirror to a spot 9–10 mm in diameter at the top of the boule. Powdered material fed from the Verneuil system passes through this high-temperature region, melted, and flows onto the boule. As the boule grows, the platform which supports it is lowered so that its position, relative to the re-imaged spot, remains constant. Vertical growth rate is adjustable from 1 to 100 mm/hr. over a total distance of 75 mm. Rotational speed of the boule is also con-

(Continued on page 138A)

Scientific Consultant $16,000-$19,000
Electronic Engineer $15,000-$18,000

The United States Air Force Security Service, San Antonio, Texas is accepting applications for two high level scientific and engineering positions in the communications/electronic field.

Scientific Consultant **has duties which include research and development, system concepts and the general guiding of projects and systems to meet approved requirements. Must be familiar with modern techniques and equipment as applied to information sensing, analysis, synthesis and evaluation. Degree and progressive practical experience preferred.**

Electronic Engineer **must be knowledgeable of modern computer/data processors and techniques and be capable of analyzing and evaluating data processing and dissemination systems for specific applications. Duties include study and recommendation for R&D data handling and data dissemination systems. Advanced degree preferred.**

Send inquiries and completed Form SF 57, Application for Federal Employment, to USAF Security Service, San Antonio, Texas, ATTN: Directorate of Civilian Personnel.



PROJECT IN POINT:

This B-58 navigator thinks he's on target at 53,000 feet!

Simulation reflects the ultimate in the *application* of science and technology. It is the electronic bridge from research to reality. At Curtiss-Wright, electronic simulation systems orient men and machines to missions for many military and industrial programs.

Project in Point: Today at Carswell and Bunker Hill Air Force Bases, B-58 navigators are being trained by the most sophisticated BOMB NAV simulators in existence. They were designed and manufactured by Curtiss-Wright under contract to General Dynamics/Convair.

The skills in systems and products developed by this and other programs are now being applied to the USAF C-141, the Lockheed turbofan freighter. Curtiss-Wright will produce fully digital simulators for flight crew training—a major step forward in this field.

These advanced activities have created immediate opportunities at Curtiss-Wright Electronics Division for solid state circuit designers, digital computer programmers and others experienced in the application of real-time digital computation to the most challenging problems in simulation.

For complete information, please write Mr. Gene H. Kelly, Manager of Professional Placement, Electronics Division. An equal opportunity employer.

ELECTRONICS DIVISION

CURTISS-WRIGHT CORPORATION

35 MARKET STREET, EAST PATERSON, N. J.

ENGINEERS • SCIENTISTS

Sylvania's Applied Research Laboratory has been pursuing investigations in the areas of speech compression, synthesis and recognition; character and pattern recognition; self-organizing systems; bionics; statistical decision theory; perception theory (color vision, in particular); pattern recognition; learning and automata theory; and contextual language analysis.

Investigations in some of these areas have progressed to a point where tangible results are being realized. For example:

■ Character recognition research has spawned systems concepts and an extensive simulation and data analysis program.

■ Orthonormal speech analysis and resynthesis has been successfully achieved via digital simulation techniques (non-real time digital simulation).

■ Statistical decision theory and perception work has borne fruit in information processing concepts in the detection, radar and radiometric fields.

■ The perception effort has led to broadening research in the applications, extension and test evaluation of color theory concepts.

a capsule report on progress in information processing research for the creative scientist and engineer



To learn more about research appointments at the Applied Research Laboratory write in confidence to Dr. Donald B. Brick, Manager, Information Processing Dept.

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NEWS New Products

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(Continued from page 137A)

tinuously variable from 0.5 to 30 r.p.m. in either direction. Temperature can be modulated throughout the entire range of radiation.

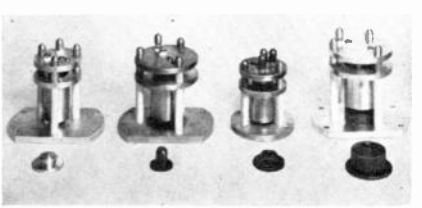
Precision Connectors Appoints Gulick

Precision Connectors Inc., an affiliate of Axel Electronics, Inc., has announced the appointment of J. M. (Jim) Gulick as sales supervisor. Mr. Gulick will operate from the plant office in Jamaica, N. Y. He will work with Precision's customer engineering and purchasing personnel supplementing the efforts of the Precision Sales Engineering Representatives.

Previous to joining Precision Connectors, Mr. Gulick was affiliated with Lionel Electronic Laboratories, Inc. (formerly Anton Electronic Laboratories, Inc.) where he was Eastern Region Sales Manager, later filling the capacity of National Sales Manager-Components Division.

Interchangeable Electron Beam Guns

A new line of interchangeable electron beam guns has been developed by **Electron Heating Corp.,** 81 Hicks Ave., Medford 55, Mass. Presently, there are four gun types available which cover the entire spectrum of materials processing applications, ranging from etching and microcomponent assembly to welding, evaporation and melting of *any known material*—metallic or non-metallic.



Gun units are triode type, and emission level is controlled with a grid bias voltage which gives control of the gun, independent of thermal inertias or filament characteristics. Temperatures in excess of 7000°F are attainable and efficiencies are greater than 90%.

Internally or externally mounted, and operating at voltages from 10 to 30 kv, EHC guns are available with widely varying power and resolution characteristics. Gun housings, focus and deflection systems, the complete power supply and control console, and/or the complete chamber and pumping system can be supplied by Electron Heating Corp.

Illustration shows the four gun types currently in production. From left to right the guns are: type 3101, a 60 watt ultra-

(Continued on page 140A)

TO ENGINEERS & SCIENTISTS
NO. 3 IN A SERIES

Noctilio leporinus, the bat that fishes, flies a search pattern 10 to 15 cm above the sea in his quest for succulent minnows, emitting frequency modulated pulses ranging from 45-60 kc at initial phases, tapering off to 25-40 kc. SO MUCH IS CLEAR. But how does he overcome his air/water interface problem?



THE PROBLEM OF AIRBORNE ECHOLOCATION OF UNDERWATER TARGETS

There's no doubt about it, Noctilio leporinus could tell human sonar engineers a thing or two about practical techniques for the utilization of the 0.1% of a train of sound waves penetrating the surface of the sea.

However, HMED sonar specialists consider N. leporinus a piker in one respect. His underwater signal propagation is restricted to depths of 3-4 cm. At HMED engineers and scientists are accustomed to deal successfully with undersea detection problems (concerning submarines, mines, natural obstructions) down to depths greater than 10,000 ft. THEY DEAL IN TENS OF MILES in their undersea-based acoustical systems for search, detection, tracking, discrimination and localization of enemy vessels of all types. (Examples: ship and submarine-borne sonars; fixed bottom mounted arrays for submarine area surveillance.)

In achieving advances in these vital defense areas, HMED engineers and scientists have developed techniques not yet paralleled by any member of the Chiroptera family (as far as they know), in very long range active and passive sonars, in very low frequency signal propagation and processing, in sonar communications and countermeasures. Right now, they are working on novel concepts in non-acoustic undersea detection that leave N. leporinus far behind in the technical stakes. (We understand he's a traditionalist—hasn't improved his methods in the last 50 million years.)

OPENINGS EXIST TODAY FOR FORWARD-LOOKING ENGINEERS open-minded to technical innovations and ready to initiate them—interested and experienced in the following areas:

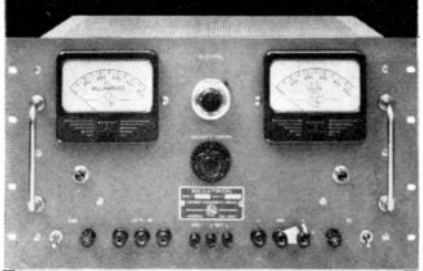
CIRCUIT TECHNIQUES; ACOUSTIC COUNTERMEASURES; UNDERSEA COMMUNICATIONS; SONAR DISPLAY TECHNIQUES; USW SYSTEMS ANALYSIS; INFORMATION PROCESSING TECHNIQUES; ACOUSTIC PROPAGATION; ELECTROACOUSTIC; TRANSDUCTION TECHNIQUES.

Additional openings exist in the fields of: TRACK-SEARCH RADAR; SHIP-BORNE RADAR; RECEIVER & SIGNAL PROCESSING; ANTENNA & MICROWAVE; DATA PROCESSING; STRUCTURAL ANALYSIS.

For more information on current opportunities write to Mr. David Carney, Div. 53-J. AN EQUAL OPPORTUNITY EMPLOYER

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ELECTRONIC MEASUREMENTS COMPANY, INCORPORATED
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NEWS New Products

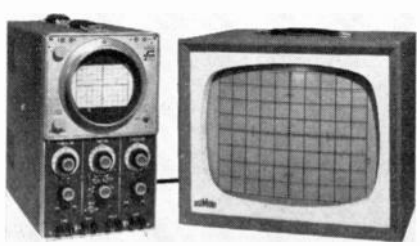
These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your IRE affiliation.

(Continued from page 138A)

high resolution etching gun; type 3102, a 3000 watt high resolution welding gun; type 3103, a 10 kv normal resolution evaporation gun and a 20,000 watt melting gun. For complete details write H. S. Berman at the firm.

Large Display Oscilloscope

An oscilloscope-monitor system to enlarge the display of laboratory precision phenomena—for remote viewing at a distance or by mass audiences—has been developed by **Allen B. DuMont Laboratories,** 750 Bloomfield Ave., Clifton, N. J., Divisions of Fairchild Camera and Instrument Corp.



The heart of the Type 436 system is a 5-inch oscilloscope which contains all functional circuitry. The scope controls and drives the deflection of the 17-inch monitor from a distance of as much as 300 feet away without significant distortion of display. Any phenomena within the systems dc-10kc frequency range is displayed many times normal size with the precision of a high quality oscilloscope.

Resolution of the large-screen indicator is 1000 lines. A 24×30 cm scale, divided into 3 cm squares, supplements accurate measurement or positioning application. The 17-inch tube is metalized and operated at 10 kv for high brilliance.

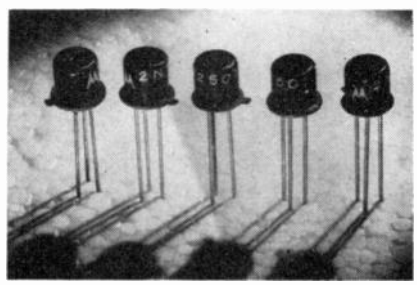
Sensitivity of the system is 5 mv/cm to 50 v/cm on the monitor and 5 mv/3 cm division to 50 v/3 cm division on the indicator. Identical X and Y axes of the scope enable equal-ordinate displays. An 8×10 cm scale on the scope permits precision adjustment of the display before enlarged presentations. Either the scope of monitor displays the phenomena by means of a switch located on the front panel of the scope.

Sweep rates from 10 sec/cm to 50 μs/cm, plus single sweeps, are available to a wide variety of presentations. Sweeps as slow as 100 seconds full scale can be obtained. Single sweeps may be reset from external contact closures.

Typical of applications for the large display oscilloscope are readout from electro-mechanical transducers, readout from analog computers, electro-medical research and monitoring, essential educational and teaching tools, production lines application and viewing, and all others where high-resolution large screen viewing is helpful or necessary.

Ultra-High-Speed Silicon Transistor

A new ultra-high-speed silicon epitaxial planar transistor, claimed to be the most completely specified silicon logic device on the market, is now available from **Motorola Semiconductor Products Inc.,** 5005 E. McDowell Rd., Phoenix 8, Ariz.



The transistor, type 2N2501, features a total pulse propagation time as low as 3 nanoseconds resulting from its extremely low capacitance, low storage time characteristics, and high gain-band-width product.

The device is characterized over its optimum usable current range, with gain (beta) specified at collector currents of 100 μa, 1 ma, 10 ma, 50 ma, and 100 ma.

In addition, the active region time constant and total control charge parameters are specified, making it possible to more closely predict circuit performance at various operating conditions (using a standard formula) than ever before.

The Motorola 2N2501 transistor, housed in a TO-18 package, is specifically designed for low-level, logic switching in the 100 μa to 100 ma region. Saturation voltage, a critical parameter in most switching applications, is specified at 10, 50, and 100 ma.

The new transistor is rated for 360 mW dissipation at 25°C ambient and has a 200°C maximum junction temperature operation.

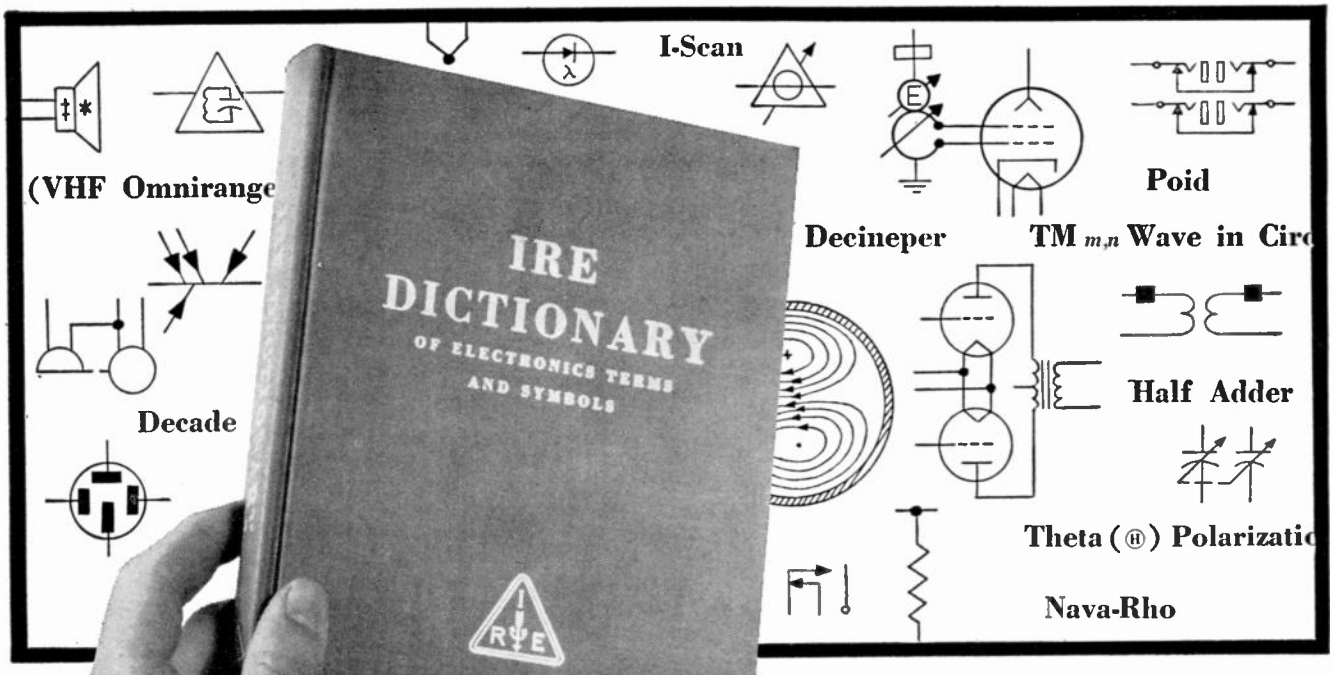
Price of the unit is $8.25 for 1 to 99 quantities and $5.50 for 100 up.

Complete electrical specifications describing the Motorola 2N2501 transistor may be obtained by writing to the firm.

Automatic Memory Core Tester

The development of the Model 2045A automatic memory core tester, designed to analyze miniature ferrite memory cores with switching speeds of up to 100 nanoseconds, has been announced by **Computer Instrumentation Corp.,** Route 38 and Lingwood Ave., Cherry Hill, N. J. The Tester incorporates a complete system of stabilized, programmed drive current pulses to excite the memory core, and a core output signal analysis system to distinguish acceptable cores from reject cores based on signal response and switching time. Front panel electro-mechanical

(Continued on page 142A)



GET THIS NEW ELECTRONICS DICTIONARY FOR YOUR BOOKSHELF!

3,700 *terms officially defined, many illustrated, in IRE's authoritative reference work*

Letter and graphical symbols also explained

CHECK YOURSELF:

From Absolute Delay to Zoning (Stepping), there are over 3,700 electronics terms you need to be aware of . . . How many do you actually know? How many can you define? Some terms have more than one meaning; are you sure you have the right, the scientifically accurate, the IRE Standard definition?

There's no need to carry around 3,700 definitions in your head. Have them handy on your bookshelf, instead, in IRE's Dictionary of Electronics Terms and Symbols. Every term accurately defined; many illustrated by simple, grasp-at-a-glance diagrams.

PART I of the new Dictionary is compiled from 37 IRE Standards on electronics terms, organized and published by The Institute of Radio Engineers over an 18 year period. Each definition cites the Standard from which it was taken; where a term is used in more than one field, and has more than one meaning, each is listed under a separate entry.

PART II of this 225 page volume is a Dictionary of Symbols, made up of five IRE Standards: Letter Symbols for Electron Tubes; Letter Symbols for Semiconductor Devices; Graphical and Letter Symbols for Feedback Control Systems; Graphical Systems for Semiconductor Devices; and Graphical Symbols for Electrical Diagrams. A four-page index to graphical symbols is included.

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Please send me ..... copies of your complete new Dictionary of Electronics Terms and Symbols.

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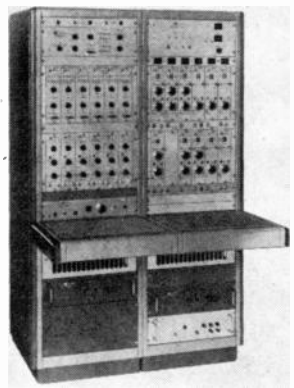
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NEWS New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your IRE affiliation.

(Continued from page 140A)

counters record complete test results including total rejected cores, total accepted cores, and the several causes of core rejection.



Characterized by very fast transistor drive circuitry, precise voltage discriminators, wideband core signal analysis circuits and a 10 nanosecond strobe pulse width, the Model 2045A has the ability to test the newer high speed memory cores and partial switching cores, in addition to cores switching in the 1 to 4 μsec range.

Current pulses in the Model 2045A are continuously variable from 60 ma to 1.25 amperes and have a voltage amplitude of 55 volts maximum. Both rise and fall time are linear and can be adjusted independently from 50 nanoseconds to 3 μsec. Duty cycle is 75% maximum into a 50-ohm load and source impedance is 3,000 ohms minimum. Each current pulse driver in the system is driven by its own independently variable width and delay generator. Pulse width varies from 100 nanoseconds to 10 μsec and pulse delay ranges from 0 to 10 μsec.

In operation, the Model 2045A develops a periodically repeated 8 step basic pulse program which has been selected for the particular type memory or switch core under test. Once the program has been selected, the core signal analysis circuits are set up to organize the system for acceptance or rejection of the cores on criteria relating to signal amplitude and switching time. The tester has three precision strobed voltage discriminators which test core output voltage amplitude at accurately predetermined instants of time, and three peak-type voltage discriminators which test core output voltage amplitude for conformance to preset maximum or minimum peak amplitudes. A "Fail-Safe" monitoring feature designed into the system assures rejection of all bad cores and disqualification of specific core tests in the event of system failure or test parameter drift.

Trinistor Controlled-Rectifier

A 16-ampere rms forward current "Rock Top" Trinistor controlled-rectifier line (JEDEC series 2N1842A through 2N1849A) with a maximum junction temperature of 125° C is available from the **Westinghouse Electric Corp.**, Semiconductor Div., Youngwood, Pa. Forward blocking voltage ratings are 25 through 400 volts and repetitive peak reverse voltage ratings are 30 to 480 volts.



The 125° C junction temperature is 25 degrees above previous units in this current range. This increase in maximum temperature offers the user the advantages of cooler operation, smaller heat sinks, smaller size, and less weight. The device will carry an average current of 10 amperes at 180° conduction angle; and a 90° C stud temperature as compared to the maximum stud temperature of 55° C under the same conditions for previous types. One-cycle peak surge current rating is 125 amperes and the maximum forward voltage drop is 1.6 volts at 10 amperes.

Typical applications of the four-layer, three-terminal devices are in proportional control systems, such as heating controls, dc motor control, pulse generators and residential light dimming control.

The devices feature all hard soldered internal construction and hermetically weld-sealed cases for reliable operation, even under severe application conditions.

The line is available from stock at authorized Westinghouse distributors. The price range is $7.35 to $24.65 per unit in quantities of 100 or more.

For more information, write to Westinghouse.

Motor-Pot Potentiometer Data Sheet

Two standard Motor-pot models are presented in a new data sheet issued by **Helipot Div., Beckman Instruments, Inc.,** 2500 Harbor Blvd., Fullerton, Calif.

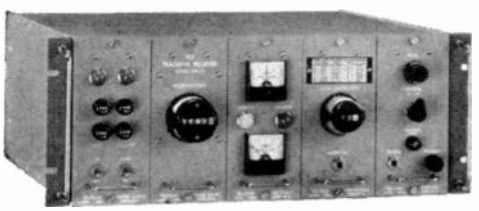
They are: Model 45, a $1\frac{1}{16}$ diameter unit built around a single-turn pot; and the 2" diameter Model 46 with a 10-turn potentiometer. Both models feature precision Helipot potentiometers and high performance Beckman servomotors and gearheads.

The Motor-pot concept combines a pot, servomotor and gearhead in a single, compact package—with the gearhead and part of the motor placed inside the pot. The result is a saving of approximately 30% over the size of typical in-line units.

Copies of Motor-pot Data Sheet 62834 are available free from Helipot Technical Information Service.

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Field evaluation by two leading government agencies confirmed the Textran receiver to be the best available. A phase-comparator and servo-driven phase-shifter yield automatic phase-locked operation. The receiver, with an external local frequency source, becomes a complete and integrated time/frequency measuring system—nothing else to buy or attach.

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Frequency Standard

A new single-unit secondary frequency standard for communications frequency measurement for every allocated frequency in the spectrum is announced and is available from the **Communications Dept., Allen B. DuMont Laboratories, Division of Fairchild Camera and Instrument Corp.,** 750 Bloomfield Ave., Clifton, N. J. Established list price is $650.



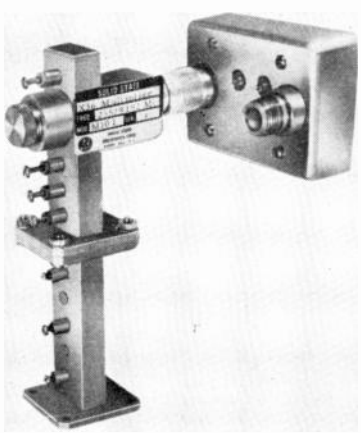
Termed the DuMont Fairco Mark III Frequency Meter, the new frequency standard can reduce to 5 kc throughout the ham, citizens, and mobile band the more widely separated frequency standards from radio station WWV.

The new meter will complement and add to the company's existing communications test equipment units including the 5819-A Universal Test Meter for servicing and tuning, as well as the 5890-B Transistorized Frequency Meter for adjacent or split channel systems.

The meter has an accuracy exceeding FCC requirements; the frequency of both receiver and transmitter can be set at the same time with one single cable connection.

Solid State Multiplier

Micro State Electronics Corp., 152 Floral Ave., Murray Hill, N. J., has announced production of the Model M101 Solid State X36 Multiplier. This harmonic multiplier provides highly reliable multiplication for precision systems such as with atomic standards.



The Model M101, using only one Micro State diode for maximum reliability, operates with a 255.3 mc input at 250 mw for a X36 output of 9192.6 mc at 100 mw. Rejection of spurious outputs of the nearest harmonics (35th and 37th) is at least 40 db below the desired frequency. The Model M101 is completely passive, requiring no battery or power supply.

Models of other multiplication factors and output frequencies and powers are also available.

For complete technical information on the X36 Multiplier, write to the firm.

Field Intensity Meter

The NM-22A Radio Interference-Field Intensity Measuring Equipment, developed by **Stoddart Aircraft Radio Co., Inc.,** 6644 Santa Monica Blvd., Hollywood 38, Calif., operating over the frequency range 150 kc to 32 mc, has built-in capability for data and spectrum signature recording. Both a superheterodyne receiver and a calibrated frequency selective microvoltmeter, the NM-22A features constant gain over the frequency range; two selectable 6 db bandwidths (3 and 10 kc) and analog voltage outputs for frequency versus amplitude recording.

Its extreme sensitivity (up to 0.1 microvolt) and broad measurement range of 140 db permits a wide range of applications for both laboratory and field use. Measurements using the NM-22A can be made on business machines, diathermy equipment, induction heating devices, condensers, filters, communications equipment of all types, radio-controlled devices, and so forth. Measurements are simplified by use of a Stoddart RFI Calculator.

(Continued on page 144A)

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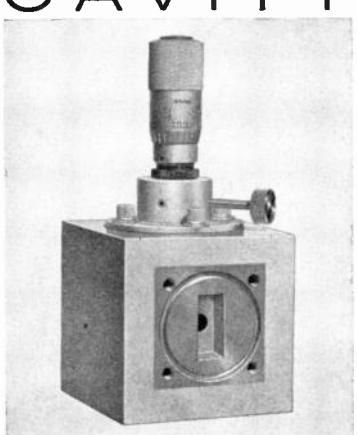
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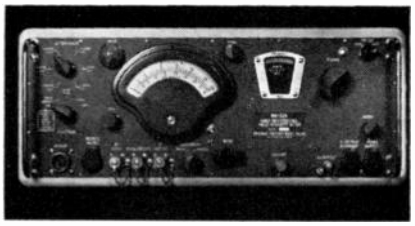
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NEWS New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your IRE affiliation.

(Continued from page 143A)



Additional front panel outputs allow connection to remote meter, recorder, oscilloscope and headphones. Its front panel IF output (1.6 mc) provides for connection to external devices for panoramic display, and so forth. All of these outputs may be in operation simultaneously without disabling the main instrument output meter.

The NM-22A incorporates an internal Impulse Calibrator, three detector functions; Field Intensity, Quasi-Peak and Peak, and operates from Stoddart's 91923-2 Universal Power Supply.

Its operating accessories include Remote Rod and Loop Antennas, and is available with either Rack Mount or Standard Case.

Price of the NM-22A RFI Meter is $2950; it is available now.

50 MS Delay Line

A new 50 millisecond delay line, believed to offer the world's longest sonic wire delay, has been designed by **General Electric Co., Heavy Military Electronics Dept.,** Syracuse, N.Y.



For use with advanced radar signal processing equipment being developed by this department, the unique device employs a special nickel alloy wire measuring over 800 feet long and 0.025 inch thick.

Incorporating high efficiency piezoelectric transducers, the high stability delay line combines low insertion losses with a bandwidth of 100 kc. The device provides a time-bandwidth product of 5000 and a data storage capacity of 10,000 bits of binary information.

Shown standing, left to right, are co-designers of the new device: Carl E. Wellman and William H. Jones.

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IS *YOUR* COMPANY ON THE *OFFENSE* FOR *DEFENSE?*

SIGNAL is your introduction to the men who control government's growing $6.5 billion radio-electronics spending

Never before have our armed forces so badly needed the thinking and products of the electronics industry. Advertising in SIGNAL, the official journal of the *Armed Forces Communications and Electronics Association*, puts you in touch with almost 11,000 of the most successful men in the field—every one a prospect for your defense products!

Share in the defense and the profits! Company membership in the AFCEA, with SIGNAL as your spokesman, puts you in touch with government decision-makers!

SIGNAL serves liaison duty between the armed forces and industry. It informs manufacturers about the latest government projects and military needs, while it lets armed forces buyers know what *you* have to offer to contribute to our armed might. SIGNAL coordinates needs with available products and makes developments possible.

But SIGNAL is more than just a magazine. It's *part of an over-all plan!*

A concerted *offensive* to let the government, which has great faith in industry and the private individual producer, know exactly what's available to launch its far-sighted plans. Part of this offensive is the giant AFCEA National Convention and Exhibit (to be held next year in Washington, D.C., June 4-6). Here, you can *show* what you have to contribute directly to the important buyers. Your sales team meets fellow manufacturers and military purchasers and keeps "on top" of current government needs and market news.

Besides *advertising* in SIGNAL which affords year-round exposure by focusing your firm and products directly on the proper market . . . besides *participation* in the huge AFCEA National Convention and Exhibit . . . the over-all plan of company membership in the AFCEA *gives your firm a highly influential organization's experience and prestige to draw upon.*

As a member, you join some 175 group members who feel the chances of winning million dollar contracts are worth the relatively low investment of time and money. On a local basis, you organize your team (9 of your top men with you as manager and team captain), attend monthly chapter meetings and dinners, meet defense buyers, procurement agents and sub-contractors. Like the other 60 local chapters of the AFCEA, your team gets to know the "right" people.

In effect, company membership in the AFCEA is a "three-barrelled" offensive aimed at putting your company in the "elite" group of government contractors —the group that, for example in 1957, for less than $8,000 (for the full AFCEA plan) made an amazing total of *459.7 million dollars!*

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(2) Group membership in the AFCEA, a select organization specializing in all aspects of production and sales in our growing communications and electronics industry; and

(3) Attending AFCEA chapter meetings, dinners and a big annual exposition for publicizing your firm and displaying your products.

If *you're* in the field of communications and electronics . . . and want prestige, contacts and exposure . . . let SIGNAL put your company on the *offense* for *defense!* Call or write for more details—now!

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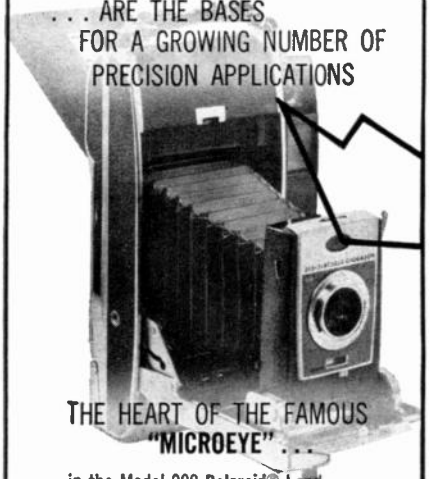
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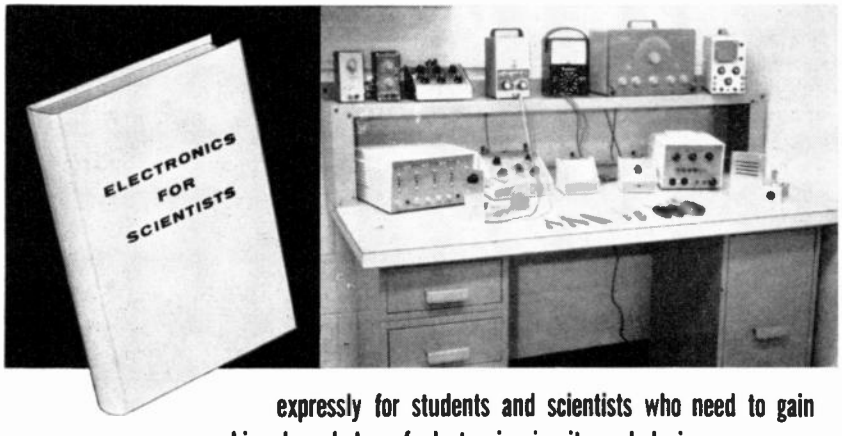
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Complete description in McGraw-Hill Radiation Laboratory Series, Volume 1, page 284 and page 209, and Volume 26, page 233.

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Output pulse power 25 KV at 40 amp. Max. duty ratio: 002. Uses 6C21 pulse tube. Pulse duration .25 to 2 microsec. Input 115 volts 60 cycle AC. Includes power supply in separate cabinet and driver. Fully guaranteed as new condition. Full Desc. MIT. Rad. Lab. Series "Pulse Gen."

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(A) 31 KV at 60 amps 5C22 Thyratron 2 microsec. 360-400 pps. .002 Duty Cycle Ideal for 5J26 at 500KW $950. (B) 30KV at 70 amps. 5C22 Thyratron 2 microsec. 285-315 pps. .001 Duty Cycle. Hyd. Thyr. $1250 w/pulse transformer.

500 KW PULSER

5C22 Hyd. Thyr. Modulator. 22 KV at 28 amps. W/HV & Fil. supplies. 3 pulse length rep. rates: 2.25 usec 300 pps. 1.75 usec 550 pps. .4 usec 2500 pps. 115V 60 cy AC input. Will deliver nominal 225 KW X Band using 4J50 magnetron.

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6 MEGAWATT PULSER

5949 Hydrogen thyratron incl. 1 and 2 microsec. PFN; 15KV 500MA power supply. Complete. 7ftx5ftx7ft.

L BAND RF PKG.

20 KW peak 990 to 1040 MC. Pulse width .7 to 1.2 micro sec. Rep rate 180 to 420 pps. Input 115 vac. Incl. Receiver $1200.

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300 to 2400MC CW, Tuneable, Transmitter 10 to 30 Watts. Output. As new $375.

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2800 MC PKG. 500W

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AN/SPS-6

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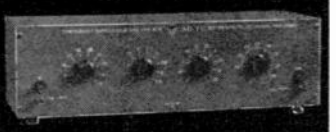
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CONTINUOUSLY VARIABLE
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ULTRA-FAST RISE TIME
Type 10T, 11T Series



Rise time can be less than 1% of total delay. Phase linear beyond two-thirds of cutoff frequency. Distortion less than 2%. Temperature coefficient better than 0.005% per degree C.

Type 10T: Total delay 0.27 to 18 us. Impedance 50 to 700 ohms. Input voltage 500 volts.

Type 11T: Total delay 400 to over 200,000 us. Impedance 475 to 15,800 ohms. Input voltage 100 volts.

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Type 602–3 Series

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Type 551–6 Series: Resolution time less than 8×10^{-11} sec. Total delay 0.029 to over 20 us. Bandwidth over 30 mc. Mechanical rotation, 22 complete turns. Impedance 50 ohms to 800 ohms.

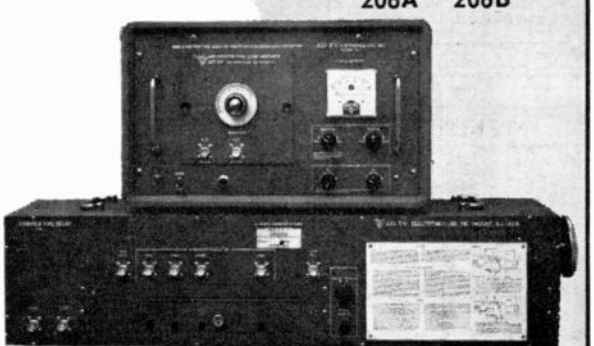
Type 521 Series: Total delay from 6 us to 30 us. Impedance 500 ohms to 1200 ohms. Two-knob control: coarse – 60 positions; vernier – continuously variable, resolution less than 8×10^{-11} sec.

MICROWAVE AND TIME DETECTORS

Type
205B1-B2-B3
206A 206B

FEATURES

Frequency range 15 mc to 18,000 mc. Accuracy ±1% or ±0.05°. Resolution time infinitesimal. Minimum signal input 1 mv.



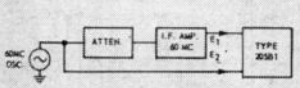
SPECIFICATIONS

Phase Range: 0 to 360°. CW, pulse-modulated or amplitude-modulated signals.

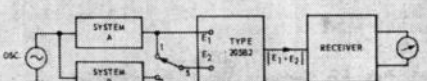
Frequency Range: 205B1 – 15 mc to 400 mc; 205B2-B3 – 100 mc to 1500 mc; 206A – 300 mc to 4000 mc; 206B – 3.9 kmc to 18 kmc.

Characteristic Impedance: 50 ohms nominal.

Time Delay: Continuously variable and in steps up to 48 mus.



Measure phase shift of an IF amplifier.



Comparison of phase angle. Accuracy can be less than 1 uus.

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POLARIS — U. S. Navy operational long-range missile, uses seven types of Texas Instruments transistors in its Mark 84 fire control system and navigational computer system. As many as 39 transistor parameters meet a 0.65 AQL to fulfill the stringent reliability demands of Polaris.

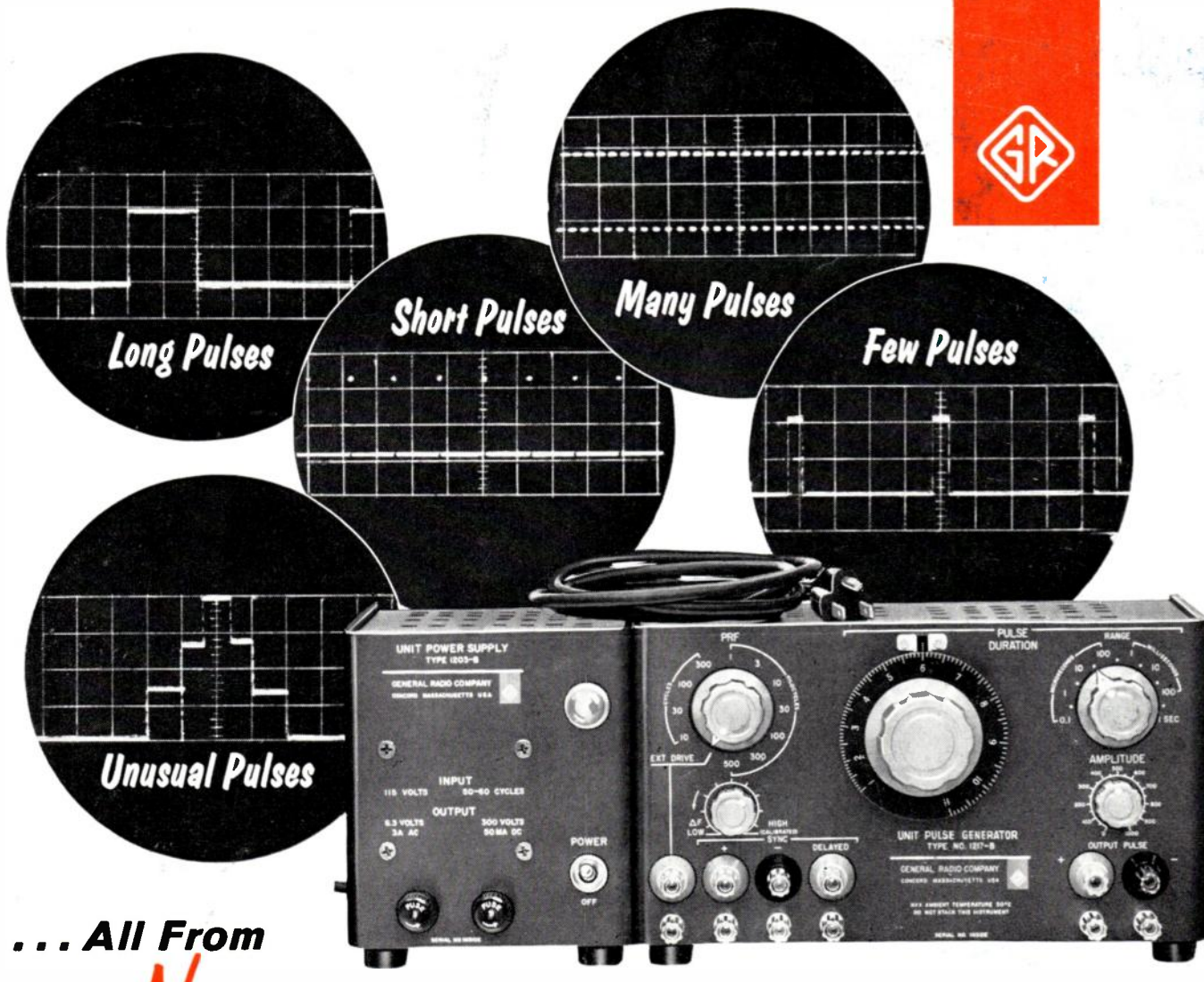
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As Many As 39 Parameters Meet 0.65 AQL . . .

Now you can design this same Polaris transistor performance into your circuits, products or systems. TI's Circuit Development-Applications department would welcome the opportunity to consult with you on your military or industrial design requirements. ■ The following TI device types are the counterparts of the Polaris units delivered to a 0.65 AQL. They are produced by the same technology, equipment and personnel that produce "Polaris quality" transistors . . . and made possible by TI's capabilities in precise process control. **2N1047 series, 2N1893/2N696 series, 2N1132 series, 2N743 series, 2N490 series, 2N1302 series, 2N428 series.** When you add TI reliability to your products, you add it at low cost because many years of high-volume production have allowed TI to develop extensive mechanization techniques. ■ More than 800 separate automatic production and testing facilities provide the manufacturing flexibility to meet your special product requirements . . . again at low cost. ■ You will find the transistors you need among the over 400 standard TI device types already in stock for your convenience. ■ Put TI transistors and TI's creative engineering assistance to work for you to provide better reliability for your products at lower cost. Call your TI sales representative, or write directly to TI's Circuit Development-Applications department about your specific requirements. You'll find out why so many already specify TI.

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. . . All From the *New* 1217-B Unit Pulse Generator

Price: $250 (less power supply)

With repetition rates from 2.5c to 500 kc (up to 1 Mc with external drive), durations from 100 nsec to 1 sec, and a typical rise time of 12 nsec, the 1217-B is one of the best pulse-generator buys to come along in recent years. Unusual composite pulse shapes can be readily generated by connecting a number of Pulse Generators in parallel. No external mixers or adding networks are necessary since the 1217-B has a linear, dc coupled output.

SPECIFICATIONS

PRF Range: 2.5c to 500 kc, internal drive; dc to 1Mc external drive

Duration Range: 100 nsec to 1 second in 7 decade ranges

Rise and Fall Time: Less than 20 nsec into 50- or 100-ohm load (typical rise and fall time about 12 nsec).

Duty-Ratio Restrictions: None

Output: Positive and negative 40-ma pulses available simultaneously (40 volts into 1 kilohm load)

Shape: Overshoot and noise less than 5% of amplitude; ramp-off is less than 1%; jitter is less than 0.01% with regulated power supply and 0.05% with unregulated supply.

Sync Pulse: Positive and negative 10-volt pulses having 150-nsec duration

Delayed Sync Pulse: A 5-volt, 100-nsec negative-going pulse coincident with late edge of main pulse followed by positive going, 5-volt 150-nsec reset pulse. Useful for triggering another 1217-B to produce composite pulses.

Type 1201-B Unit Regulated Power Supply . . . $90 • Type 1203-B Unit Power Supply . . . $50

Write for Complete Information

GENERAL RADIO COMPANY

WEST CONCORD, MASSACHUSETTS

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