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PROCEEDINGS OF THE I.R.E.

INSTITUTE NEWS AND RADIO NOTES SECTION

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J. W. McRae

Board of Directors, 1951–1952

J. W. McRae was born on October 25, 1910, in Vancouver, British Columbia. He received the B.S. degree in electrical engineering from the University of British Columbia in 1933, the M.S. degree in 1934 from the California Institute of Technology, and the Ph.D. degree from the same institution in 1937. Early in 1937 he joined the staff of Bell Telephone Laboratories, Inc., where he engaged in research on transoceanic radio transmitters. His work has included microwave research, especially a microwave oscillator for the National Defense Research Council.

In 1942 Dr. McRae was commissioned a major in the United States Army Signal Corps, assigned to the Office of the Chief Signal Officer in Washington, D. C. Here he worked on the co-ordination of development programs for airborne radar equipment and radar counter-measures devices, and received the Legion of Merit for his services.

Dr. McRae was transferred to the Headquarter of the Signal Corps Engineering Laboratories, at Bradley Beach, N. J., in 1944, as chief of the engineering staff, and became deputy director of the division, attaining the rank of colonel before returning to civilian life in 1945. Rejoining Bell Laboratories, he was appointed director of radio projects and television research in June, 1946, and was responsible, also, for work on the New York-Boston radio-relay project. With the addition of responsibility for electron dynamics research in February, 1947, he became director of electronic and television research. In January, 1949, he was appointed assistant director of apparatus development I. and on March 1, 1949, was made director of this work. He became director of transmission development on October 1, 1949, and on June 1, 1951, Dr. McRae was appointed vice-president in charge of the systems development organization.

Dr. McRae joined the Institute in 1937 as an Associate, and was made a Fellow in 1947. He has been a member of the Board of Editors since 1946, and served on the 1947 IRE Convention Committee. He was Chairman of the New York IRE Section in 1949, and is now a member of the Board of Directors and the Executive Committee. Dr. McRae also serves as Membership Relations Co-ordinator, and belongs to the IRE Admission, Appointments, and Awards Committees.

He has held membership in the American Institute of Electrical Engineers and Sigma Xi since 1936. Eta Kappa Nu awarded him an Honorary Mention in their recognition of Outstanding Young Electrical Engineers for 1943.

Mathematical Style and the Technical Paper

F. J. BINGLEY

Correspondence over many years with the membership of the Institute has shown a pronounced diversity of opinion concerning the importance and desirability of including mathematical papers in the PROCEEDINGS. Some have felt that such papers are basic, usefully instructive, and much to be encouraged. Others have regarded them as uninteresting and of but little value to the practicing engineer. A balanced viewpoint, which has guided the editorial readers, is that so-called "theoretical," as well as "practical" papers are of much value to the readers and should accordingly both be presented.

The following guest editorial offers a clear justification (if that were needed) of present Institute editorial and publication policies. Its author is a Fellow of the Institute, a member of the IRE Papers Review Committee, a noted contributor to the development of television broadcast engineering, and is associated with the Research Division of the Philco Corporation.

—*The Editor.*

The development and exposition of engineering ideas usually involve the extensive application of mathematical methods. It is certainly desirable that this should be the case because only in this way can experimental results be fitted into a framework from which a complete structure of knowledge can be erected. All this is well recognized, and we of the radio-engineering profession have made extensive progress by this method. We have learned that knowledge is built by a sort of dual process. The process usually starts with the discovery of some experimental effect which, after being the subject of many tests and measurements, appears to follow some law capable of mathematical expression. Upon the assumption that the law holds, further implications which would appear to follow from the law are derived by mathematical analysis. This leads us to a new set of physical facts which should then be true, whereupon these apparent facts are put to the test of experiment. If they pass this test, knowledge is further strengthened. More mathematical analysis follows, and more tests and measurements. And so it goes, with the structure of knowledge getting bigger and more solid with the passage of time.

Thus mathematics is an essential part of engineering. Just as we need clear, accurate, definitive experimental data, so must we have mathematical analysis of the same caliber. Just as we must use clear, lucid words to describe our experimental findings, so must we use clear mathematics for their further description and analysis. The quality of expository writing that fulfills these requirements is known as "style," which exists just as surely in mathematics as it does in literature. It avoids the florid and ostentatious use of complex methods where simpler processes exist; it does not leave the derivation of the solution as an exercise for the reader, but shows him clearly the shortest and best mathematical route to the desired end.

Achieving mathematical style in a paper requires specific effort on the part of the author. The first derivation of a new result is invariably made over a rough and circuitous path, and the only factor which makes such a difficult journey worthwhile is the satisfaction which the engineer himself derives from obtaining the result. But in describing his results later, the author should not require his reader to suffer the same hardships. He should, instead, smooth out and straighten the path so that the journey for the reader is easy. This means additional work and thought for the engineer, but it is not without its rewards. For he becomes the more sure of the accuracy of his findings as he beats other straighter paths towards them; he acquires new viewpoints, and perhaps discovers new facts as he covers the intervening territory more completely, and by increasing the readiness with which his findings may be assimilated by others, he accelerates the rate at which his ideas are adopted and accredited to him. The whole process may be likened to the optimizing of circuit constants to reduce unnecessary losses.

It is difficult for an engineer, newly triumphant over some complex problem and anxious to give his results to the technical world, to pause long enough to make sure that his presentation is in conformity with the thoughts expressed here. But reflection will prove to him that such time is well spent, and that the quality of mathematical style adds greatly to the value of his work.

The Organization of Research*

D. B. HARRIS†, SENIOR MEMBER, IRE

Discussion

IT IS RATHER generally agreed that the type of organization required for the effective conduct of a research activity differs from the usual business organization in two major respects; first, in that the organization must be built around the individual, contrary to the usual requirement that the individual adapt himself to the organization; and second, in that truly effective project planning cannot be done from the top, since it is only the individual research worker who is in a position to plan his next steps on the basis of information developed from previous steps. Hafstad[1] has expressed these concepts admirably in the following words:

" . . . one does not have an organization when he runs a big research show, but has instead a collection of individuals who have their own individual peculiarities." And:

"Research is . . . like a plant—like something that grows. The only way you can control it is to fertilize it a little. If it is growing well in a certain location, throw in a little fertilizer, which means more dollars. If it isn't growing well with reasonable care, there is no use wasting any more fertilizer or dollars on it. It won't grow anyway. You can't raise water lilies in a desert."

That these conclusions should now be rather generally adopted as almost axiomatic is not surprising. The character of research is so extremely individualistic that in the case of a research organization the importance of individualism is an almost self-evident fact. The surprising thing is that it should have taken the scientists and researchers to recognize a situation which, in fact, may not at all be confined to research, but may actually exist rather generally in the business field.

The success of any organization certainly depends to a great extent on the people in it, whether the business involved is the determination of basic facts of nature, the development of new types of electronic gear, or the manufacture of breakfast food. That organizations must be built, at least to some extent, around people is therefore a general principle which may be more strikingly demonstrated in the case of scientific enterprise, but which also applies to almost any kind of endeavor.

In considering the question of research organization, we accordingly need to bear two general principles in mind: first, that this is only a specialized example of a general situation; and second, that the root of the problem is people. For effective results, we must so constitute our organization that the people in it can function with satisfaction and efficiency. If this objective is not achieved, no amount of "master-minding," "cracking the whip," "regimentation," or detailed attention to specific problems will produce satisfactory results.

In determining what specific steps need to be taken in order to tailor an organization to these requirements, it would be most helpful if we might be able to make a list of the things which keep people "satisfied and efficient." Such lists have been made under the heading of "Wages and Working Conditions," and include such things as adequate salaries, well-lighted quarters, adequate working space, satisfactory drinking water, premium pay for overtime, vacations, pension plans, and wage progression schedules. These are all things desired by almost anyone. But many executives have been surprised to find that, in some organizations in which the greatest amount of attention has been given to these matters, the morale and accomplishment of the group are fairly often the most inferior.

These things are not enough. They are essential, it is true, but something over and beyond satisfactory wages and working conditions is required to produce good morale in an organization. This additional requirement, unlike wages and working conditions, which have to do with the physical welfare of the employees, is a matter of the spirit, and it is here that the personnel director steps out and the operating executive comes into the picture, along with his personality and his operating practices. These are the things which really determine the morale of an organization—the way employees are treated in the day-to-day performance of their tasks.

Our quest for information as to what we need to put into our organization to keep people satisfied and efficient, is, therefore, limited, for purposes of the present discussions, strictly to the operating organization of the enterprise, the assignment of responsibility and authority, operating routines, and the effect these things may have on the happiness and efficiency of personnel.

Here we immediately find that it is not possible to make a list of types of operating procedures which will be most palatable to the personnel and at the same time the most effective. For people differ in what they demand from an organization in the way of responsibility, security, and achievement. And their desires are not always compatible with the welfare of the enterprise.

In analyzing this question, it is a temptation to classify individuals into two general groups with respect to their aspirations. Such classifications of people are nearly always fallacious, because some people, in fact most people, cannot be fitted precisely into any one of a limited number of groups. Nevertheless, for purposes of

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[1] G. P. Bush and L. H. Hattery, "Scientific Research; its Administration and Organization," American University Press, Washington, D. C.; 1950.

illustration, and bearing in mind the fact that actual individuals will necessarily fall into intermediate classifications not considered, let us examine two extreme types of mental outlook.

First, there is the well-qualified routine worker. As far as his job is concerned, he has no particular desire for achievement, and his objective is his pay check, which he may use to buy for him those satisfactions which he does not obtain from his work. He may have ambition for advancement and the resulting higher salary, but his primary concern is security. He prefers to be told what to do, in the greatest detail, and feels uncomfortable when called upon to make decisions. He is a skilled and efficient worker, the backbone of the organization, so long as his work is in a field with which he is familiar. He learns almost entirely through experience and feels that it is a responsibility of the company to provide him with training by means of which he can acquire the experience necessary to permit him to perform his job with greater efficiency.

Second, there is the ambitious "career man." To him the satisfactions of accomplishment derived from his job are a predominant motivation in his life, and he would be willing to work for a small salary, if he were permitted to feel the pride of achievement. He is highly intelligent, and a rugged individualist. He despises routines and regimentation, and doubts if there is any true necessity for them, since he himself is able to make satisfactory decisions without recourse to them. He is apt to be a disturbing element in the organization machinery since his first inclination is to cut red tape and get the job done regardless of rules, regulations, and the necessity for co-ordination. He learns through processes of deduction, rather than through trial and error, or experience. His goal in life is achievement, and he is completely ineffective, dissatisfied, and useless to the organization if confined to a routine job, or unduly restricted in the latitude of decision permitted him.

These two extreme cases illustrate a discrepancy which must be taken into account in planning an organization. It is not possible to provide a type of organization, having uniform rules applicable to all employees, in which both types of individuals can function with a high degree of ease and efficiency. The routine man, of the first type, demands close supervision, and rigid and complete regulation of all operations which he performs. These are the exact conditions under which the abilities of the individualist, of the second type, are circumvented, with the result that, if he does not leave the job and go elsewhere, his valuable potentialities are generally completely wasted.

Does this mean that we must choose one or the other, that we must either have a completely regimented organization of routine workers, or an entirely uncontrolled, anarchical organization of individualists? No, because here, a third factor comes into play: the requirements of the organization itself, as a mechanism. These are twofold, insofar as the question of routine practices is concerned. In the first place, no organization can afford to be completely without rules and regulations, because an organization is a co-operative enterprise, and, like a society, it must contain provisions for channeling its activities with a reasonable degree of effectiveness so that the necessary compartmentalization and specialization can be maintained, and intergroup activities can be made to run smoothly. Thus, the rugged individualist must be prepared, when working in a group, just as in his social relationships, to sacrifice some of his personal independence in return for the assistance received from specialists in other branches of the organization. In society, he would not expect to buy and charge a pair of shoes at a department store without being prepared to handle the paper work that accompanies the transaction, in this case a charge slip and a bill; similarly, in an organization, if he wants the services of a machinist, he must be willing to conform to routine requirements relating to the preparation and routing of a shop order. Of course, in society, he might elect to make the shoes himself, or in an organization, to go out and do his own machining, just for the sake of being independent of routine, but ordinarily, no sensible person would deny himself the benefits of specialization in this fashion.

Second, any organization which expects not only to progress, but even to hold its own in competition, must number among its personnel at least a few individualists, aggressive, independent thinkers. They are the vital spark in the enterprise, the policy makers, the leaders who are able, when confronted with problems beyond the scope of their own experience or that of the group, to solve them by logical deduction, and guide the enterprise through uncharted waters. Without them, an organization is inevitably confronted ultimately with failure, brought about usually through obsolescence.

Thus, there is an essential place in any organization for both types of individuals, and we reach the conclusion that routine practices must somehow be adapted to accommodate them both. Actual experience shows that in most business enterprises, there is no difficulty in accommodating the routine worker. The mechanical requirements of an organization naturally result in the development of a mass of routine practices specifying, in the smallest detail, the exact procedure to be observed in any contingency. This is definitely the "machine" approach; planners are very apt to regard the organization as a machine, in which every part is expected to function in a certain prearranged manner in conjunction with every other part, within very narrowly specified tolerances. All that is needed to bring about this result is the application of industry to the problem of writing practices, and the lapse of time. It is axiomatic that the larger an organization gets, the more specialized, compartmentalized, and routinized it becomes.

The real problem then is to take care, while recognizing the value and necessity of rules and regulations, to avoid their glorification to the extent that the organization eventually becomes totally regimented, and loses all its individualists, who should be its leaders and planners. This is a very real danger, which has actually overtaken some of our larger concerns. It is ironical, that in this country, the last bulwark against the tides of socialistic totalitarianism and regimentation, many of the independent capitalistic enterprises, which we are defending, have themselves become so regimented, that, from the standpoint of the employee, he might as well be working for a totalitarian state. It is actually a fact that in not a few organizations, instructions have been issued to employees on how to wipe their hands. These instructions, it is true, were issued for the purpose of conserving paper towels, and may therefore be said to have resulted from one of the "mechanical" requirements of the organization viewed as a machine: but the damage done to employees' morale by such insulting practices far outweighs any savings achieved.

The problem is aggravated by an unstable condition of equilibrium which necessarily exists. If an excessively routine worker, of the first type considered in the foregoing, gets into a position of authority, it is natural for him to perpetuate his own ideas in the organization. It thus becomes even more strongly regimented than before, additional routine workers are elevated into positions of authority, and the ultimate result is an organization solidified into a condition of complete rigidity, with no hope of ever relaxing it, short of a complete and cataclysmic replacement of its entire supervisory staff. Such individuals with an excessively routine slant are very apt to regard conformity as the sine qua non for employment. They resent independent thinking not based on experience, feel that it is presumptuous for a subordinate to express original views, and if he does so, regard him as an egotist, and a revolutionary. To them, discipline is everything; rules and regulations are an end unto themselves to be observed religiously, without exception, purely for the sake of conformity; and the individualist is a dangerous character whose presence in the organization is an unnecessary hazard. Many of these things may be true, to a degree, but any organization which has reached this extreme, in which the individualist cannot function because by his very nature he cannot be a complete conformist, has nothing to look forward to but ultimate failure and dissolution because of the very lack of leadership.

Such a situation may, in fact, be as disastrous as its exact opposite, a complete lack of any organization or regulation. This lack of organization also actually exists in many of the smaller concerns, which have not as yet had an opportunity to gear their operations to their rapidly growing size. Definite responsibilities are not fixed, and not only is the supervisor frequently in doubt as to which of the workers is reporting to him, but the workers themselves do not know who the boss is. Individual workers receive most of their instructions directly from the head of the division, short-circuiting intermediate supervisors and leaving them completely in the dark as to happenings in their own department. Routine procedures are largely lacking, and the procurement of parts and components, properly handled by smoothly running routines, may turn out to be a major project for individuals in supposedly responsible positions, who are forced to hand-carry papers through the organization at the expense of time taken from their proper duties, and with the probable result that someone gets left out, and records are not corrected. Due to the lack of assignment of responsibilities, subordinate supervisors are not in a position to handle their jobs effectively, and top supervisors find themselves more and more loaded down with routine paper work as their organization grows. Politics are rampant due to the lack of well-delineated job requirements, and individuals frequently reach positions of dominance merely by dint of intrigue and aggressiveness, rather than through ability. There are no regulations for keeping costs under control, and efficiency drops to a minimum. Organizations in this condition can operate only at a financial loss, and must take steps to remedy the situation by setting up a reasonable organization procedure, if they expect to continue to operate at all.

Two extreme philosophies of management have been applied to various enterprises, with varying results. One, typified by some of our larger corporations, treats the entire organization strictly as a machine, and makes extreme provisions for standardization, which facilitate the application of specialization over large geographical areas, and throughout extensive fields of endeavor. This type of management philosophy makes it possible and desirable to employ personnel of average intelligence, who are not only willing to accept, but actually demand the rigid, detailed working practices required to bring about perfect co-ordination of the vast and intricate organizational mechanism. Such organizations may be highly efficient, may provide excellent service or products at a reasonable price and may show a fair profit made possible at least in part by their low salary costs. But they are extremely vulnerable to obsolescence, since their excessive regimentation discourages the development of future leaders having the vision and imagination needed to cope with the competition induced by technological and social changes.

The other philosophy of management referred to is that which was employed, more or less through force of circumstances, in operating some of the governmentally controlled and financed research enterprises during World War II. Scientists of exceptional reputation and brilliance were brought together from universities and industries all over the country into laboratories established on the spur of the moment, and on a supposedly temporary basis, for the purpose of working in specific fields of endeavor. Ample funds were provided, and, in general, though nominal controls were maintained by the government, the workers were given almost complete latitude to do as they wished, only the broadest possible delineation of a research program having been set up by the administrative personnel at the top of the organization. The results were astounding. New inventions and developments of old inventions, fundamental technological improvements, and discoveries in almost every field of scientific investigation started pouring out of the new laboratories at a rate at least an order of magnitude greater than that previously considered normal in industry. Developments which would ordinarily have been expected to take twenty years to perfect were completed in two years. Proposals for outlandish new devices which had previously been rejected as unworkable were accepted, projects were set up, and all of the devices were made to function.

This philosophy of organization proposes the employment of the most brilliant staff available, the provision of ample funds, and then an attitude of almost complete laissez faire on the part of management. The success achieved with these methods during the war indicates that the policy is not far from wrong. Certainly, the latitude given the workers resulted in constructive planning by a much larger number of qualified people then would have been possible, had all policy been dictated by management. Certainly, this type of management afforded a congenial atmosphere for the individualist, and ample opportunity for his development as a leader. On the debit side, however, it must not be overlooked that the lack of controls resulted in tremendous waste, which could have been tolerated only by the government during wartime. Moreover, the lack of organization in many cases actually impeded the scientists in their work, because sufficient management mechanisms to assist them in obtaining material, test equipment, and other facilities were not always provided. Finally, this sort of organization approach is feasible only when it is possible to obtain the services of large numbers of highly paid and extraordinarily qualified men capable of accepting the responsibilities thrust upon them. Under normal conditions, the availability of such personnel might be limited, and in any case, the salary cost might be prohibitive.

We reach the conclusion that a certain amount of organization and routine is essential, because, for mechanical reasons, no group effort can hope to succeed without it, and because the routine workers demand it. On the other hand, the organization must be so constituted that the individualists are exposed to a desirable minimum of regimentation.

How, then, may this result be brought about? The following general principles are proposed, as at least a partial solution of the problem:

1. Management Should Plan and Serve, Not Rule

Too often, rules and regulations are not the result of organizational necessity, but merely of empire building on the part of some member or members of the management. The writer is familiar with one organization in which it was actually necessary at one time to refer a recommendation through 5 organizational levels, with discussion at each level, in order to obtain a carefully engineered job order for moving an electric time clock 6 inches from its original position, a job which could easily have been done by a reasonably skilled carpenter under the direction of the lowest level of supervision. Such requirements merely tie up the supervisory organization in useless discussion, and serve no purpose other than aggrandizement of the position of the top man.

Aside from the question of practices, a similar attitude sometimes prevails in the supervision of day-to-day operations. Some supervisors inadvertently seem to feel that they are able to maintain their power and authority only by displaying a position of autocratic domination, typified by a constantly critical attitude toward subordinates and an unwillingness to accept independently made suggestions, which, in some way, are always considered by them to be a reflection on their omnipotence and omniscience.

On the contrary, it is the proper duty of a supervisor, both in his planning of the organization and in his daily supervision of its operations to maintain constantly an attitude of helpfulness. It is his job to help his subordinates to accomplish the objectives which he has previously set out for them in a general way. Such assistance can generally best be rendered by maintaining an open-minded attitude, accepting suggestions made, if they are valid, and taking whatever co-ordinating action may appear to be necessary to place them in affect. If recommendations do not appear to be valid, the supervisor needs to explain their shortcomings carefully to the subordinate, so that he understands the reasons why they are not considered acceptable, and will therefore be in a position to act correctly in future cases. With proper job training of this kind, an organization can eventually practically be made to run itself, to the extent that the supervisor has nothing left to do but take care of replacements and advancements, and shoot trouble. Under these conditions, he may be accused of being merely a rubber stamp; nevertheless, it is a fortunate supervisor who is able to have at his command a smoothly running organization of this kind and he can well afford to overlook such jibes, with the consciousness that actually he was the one who, through his liberal policies, brought about this salutary result.

2. Delegate Responsibility to the Lowest Level Which Can Handle It

The abilities of low levels of management frequently surprise those at the top, who have, perhaps, had the feeling that if they, personally, do not do a job, it will not be done right. There is no reason why the supervisor in the lowest level should not be given a description of the fields in which he is expected to function, and then given latitude to function independently within the boundaries of his assigned responsibility. Management must then be prepared in accordance with (1) above, to furnish him with all the information and help he needs to discharge these responsibilities satisfactorily. Only in this way can the organization make full use of the capabilities of its employees of supervisory caliber. Frequently, in a growing organization, a supervisor who has been accustomed to performing his own detail work, continues to do it after he has reached a high position in the management. The only result is to create a bottleneck at the high supervisor's level, and to deprive lower levels of the freedom needed to function efficiently.

3. Delegate Authority Along with Responsibility

The spectacle of the hapless supervisor who is loaded down with responsibility for results, but given no authority to achieve them, is all too prevalent. Such a situation may come about either through a failure of management to define the status of the supervisor in the organization, or through the retention by management of discretionary powers needed by the supervisor for the discharge of his duties. For example, a supervisor may be held accountable for production, but the procurement of the ordinary tools which he needs to achieve that production may require approval by management and be withheld. Such weasel-worded assignment of authority achieves nothing but resentment. When a supervisor is given a responsibility, his position in the organization should be made clear to his subordinates, his colleagues, and his superiors, and he should be given the authority to act, and the authority to procure means for accomplishing his objectives, within financial limitations which can as well be administered by himself as by the top levels of management.

4. Establish Working Practices Needed to Make the Organization Function Properly and to Maintain Reasonable Control over Expenses, and Only Such Rules and Regulations

Aside from the obvious requirement that rules are required to discharge contractual and financial responsibilities, the proper test of a working practice is "does it help the worker to function effectively?" If the answer is "no," the practice should not be adopted. Moreover, this rule must be interpreted with judgment. We might say, for example, that the establishment of a highly specialized service department might not only save money, but also help the worker by relieving him of detailed work which he had been performing himself. Ordinarily this would certainly be true, and working practices designed for co-ordinating operating activities with such a department are the very kind of working practices which are well justified in an organization. But if the service department becomes so extensive and so organized within itself that it can be reached only by going through levels clear up to the top of the organization and back down again, the paper work may become so overwhelming that it actually may be questionable whether the service department and its practices are a help or a hindrance to the operating department.

A second consideration has to do with what constitutes "proper" functioning of an organization. No organization can ever be made to function perfectly, and any attempt to make it do so will inevitably result in the establishment of an excessively elaborate structure of procedures not worth the cost of administration. Procedures should not be written solely for the purpose of avoiding minor errors which might be expected to occur with negligibly small frequency. For example, an accounting practice requiring detailed accountability for every screw and nut used in an assembly might eliminate the occasional loss of a few screws and nuts, but would certainly not justify the expense of its own maintenance. Almost any executive in an enterprise of long standing will, if he examines his practices, be amazed to find how many of his records and procedures, established over the years, were really set up with the objective of providing safeguards against unimportant contingencies having a negligible probability of occurrence. Obviously, such practices merely produce unnecessary rigidity and add to the cost of administration without achieving any worth-while result.

On the other hand, it is highly essential that enough working practices of general scope be set up to make the organizational machine run smoothly. Such practices should cover in detail those co-ordinating operations which are known to be performed frequently. Assuming the availability of able, intelligent workers, it would be possible, of course, to let each individual improvise his own routines as the occasion arose; such improvised routines, however, may be expected to be only partially satisfactory, since even the most fertile imagination cannot anticipate all of the consequences of a relatively insignificant act affecting several parts of an organization. The only true test of a procedure is the success of its application, and success rarely comes the first time. Intelligently balanced routines are truly a help and time-saver to all concerned, including the rugged individualist, who might, at first, be inclined to play by ear.

5. Adapt the Working Practices to Fit the Requirements of the Type of Workers to Whom They Are Intended to Apply

On the basis of suggestion 4, detailed working practices may legitimately be quite restrictive, because some of the routine workers really need full and complete instructions to function effectively. But practices intended for observance by supervisory levels, such as interdepartmental co-ordinating routines, should be examined carefully for undue restrictiveness before approval. As a corollary, no routines should be adopted solely for the sake of discipline; they should stand on their own feet as a desirable and necessary organizational tool or should not be established at all. There may be a tendency in an organization to attach too much importance to ritual, and to consider that it is good merely because it is ritual; this attitude should be avoided.

6. Assign Routine Workers to Routine Jobs: Assign Workers Having Supervisory Abilities to Supervisory Jobs

This principle, difficult as it may be to observe in practice, is the key to the problem of creating an organization in which routine workers and individualists can both function with ease and efficiency. The routine tasks of the organization, of which there are necessarily many, should be performed by employees having abilities for this sort of work, and having the sort of mental outlook which enables them to feel at ease in it. On the other hand, assuming satisfactory abilities, no employee of supervisory caliber should be permitted to remain long in a routine position. This will, of course, entail the exercise of judgment in hiring new employees; no more supervisory employees should be hired than can be accommodated easily in the proposed organization.

Applications

Having formulated these general principles, which, if they are valid at all, appear to be applicable to any sort of organization, it would now seem proper to determine just how they should be applied to the organization of research. The answer is simple: they should be applied just as they would be applied in any industrial organization. The only difference between the two lies in the caliber of their personnel. In a research organization, by the very nature of the activity, there must be more individualists than in a production enterprise. The scientists themselves must be treated as supervisors, which, in effect, they are, both from the standpoint of their abilities and qualifications, and in the sense that they do direct the activities of technicians in the groups. In a research organization, most of the operating practices will therefore be of the liberal type intended for application by supervisory personnel, and there will be relatively fewer practices, even of this type, than in an industrial organization. More responsibility will be delegated, and management will be particularly careful to avoid dictatorial attitudes, and to dispose the available qualified personnel in positions of responsibility, insofar as possible. In fact, in many cases, it will actually be desirable to modify specific job requirements in order to suit the wishes and qualifications of individual members of the staff and enable them to work easily in the positions created for them.

In short, while it should indubitably be the prime function of management in any line of endeavor, to create, by means of its organizational planning and its day-to-day operation of the business, an atmosphere which encourages enthusiasm and voluntary industriousness on the part of all its employees, the opportunity, to achieve satisfactory results by these means is particularly available to a research enterprise, by virtue of the caliber of its personnel. On the other hand, if advantage is not taken of this opportunity, a research enterprise is for the same reason particularly susceptible to failure. It may be hoped that through the exercise of these principles, a concern engaged in research and development may be made to produce not only important scientific results, an end in itself to many of its personnel, but also a sound and sustained profit record.

Experimental Radio-Telephone Service for Train Passengers*

NEWTON MONK†, SENIOR MEMBER, IRE

This paper is published with the approval of the IRE Professional Group on Vehicular Communications, and has been secured through the co-operation of that Group.—*The Editor.*

Summary—**Experimental public radio-telephone service for train passengers was inaugurated by the Bell Telephone System several years ago. Initial installations operated in conjunction with a series of urban mobile base stations. More recently, highway mobile systems have been used for this service, and this paper describes a typical train installation operating through a highway channel. All of these early systems utilized an attendant on the train. The cost of providing an attendant has, in some cases, been found excessive. Consequently, experiments have been initiated in which a coin box is used on the train. The arrangements for this purpose are also described.**

Introduction

COMMUNICATION with moving trains is not new. The use of radio for this purpose has been the subject of investigation for many years, and many railroads in the United States use radio communication for promoting safety and expediting operations.

Telephone service for passengers has not received such intensive development. However, in 1929 the Canadian National Railways experimented with a service of this type, using the inductive system, between Toronto and Montreal. About the same time, the Germans inaugurated a similar system between Berlin and Hamburg. While these systems functioned reasonably well, they were soon abandoned for economic reasons.

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During the past 25 years, the Bell System has co-operated with the railroads in an effort to find a solution to this problem. It was not, however, until the introduction of general mobile telephone service that it became practical to establish an experimental radio-telephone service for train passangers. In August, 1947 such a service was introduced for the first time on several trains of the Baltimore and Ohio and Pennsylvania Railroads, operating between New York and Washington. This was followed by installations on the New York Central and the New York, New Haven, and Hartford Railroads, and later on the Southern Pacific between Los Angeles and San Francisco.

All of these systems utilized a train attendant furnished by the railroad. Experience has indicated that the cost of providing such an attendant is excessive. Consequently, experiments have recently been initiated in which a coin-box telephone is used instead. One such experimental installation was made on the Congressional Limited of the Pennsylvania Railroad in the early part of 1950, and recently the telephones on two cars of the New Haven Railroad were converted from attended to coin-box operation.

The original installation between New York and Washington operated through four urban mobile systems which had already been established in four of the principal cities along the route of the trains. These urban systems are individual systems not co-ordinated with each other. All, however, operate in the same frequency range, thus making a more or less continuous system possible. Such an arrangement, obviously, can be employed only where there are a series of urban mobile systems along the railroad line.

Besides the urban mobile systems, the Bell System operates a considerable number of highway mobile systems designed to provide service to trucks and to other vehicles moving along a highway. These systems may also be used to provide a train service where the highway system roughly parallels the railroad right of way. Most of the later train passenger systems were established to operate in conjunction with highway systems.

The New York-Washington service was described in a paper presented by Monk and Wright at the March, 1948 National Convention of the IRE and published in the September, 1948 issue of the PROCEEDINGS OF THE I.R.E. This paper is, therefore, confined principally to a description of the other services and, more particularly, to that installed on the New York Central Railroad between New York and Buffalo; the latter may be taken as representative of a system operating through a highway channel. Following this, the arrangements employed to eliminate the attendant and provide coin-box operation are described briefly.

Common-Carrier Mobile System

A common-carrier mobile system[1] consists of one or more fixed base transmitters, each radio transmitter having associated with it one or more radio receivers. Each base transmitter and its associated receivers are connected by wire lines to a control terminal, at which point the radio transmission in each direction is combined and connected to the wire plant. The control terminals are connected to mobile service operators' positions at nearby toll switchboards, and a connection from a land telephone is set up through the local central office in the customary manner. Where only one base transmitter is employed and the system covers a limited area, such as an urban community, it is known as an urban system. Where a number of base transmitters and their associated receivers are arranged to provide substantially continuous coverage to vehicles moving along a relatively long highway, it is called a highway system. A block diagram of a base transmitter, together with its receivers and control terminal, is shown in Fig. 1. This figure also illustrates how a connection is made to a land telephone.

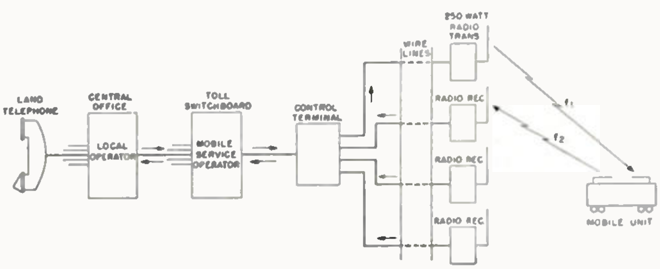

Fig. 1—Block diagram of general mobile station.

The arrangement shown may be considered to be the base station of an urban system or one station of a highway system. The principal difference resides in the frequencies employed. All urban systems operate in the 152 to 162-mc band, whereas the highway systems utilize frequencies between 30 and 44 mc.

There is another essential difference: In a highway system it is desirable that the service area for the several stations overlap. Since all stations use the same frequency, unless proper precautions are taken, interference may result when a mobile unit is operated in these overlapping areas and two adjacent base transmitters are in operation. Accordingly, the base stations of a highway system are co-ordinated by wire circuits, and operating practices are set up which prevent a vehicle in an area of overlapping transmission from obtaining transmission from more than one station at a time.

Over-All System Aspects of the Train Service

A map showing the route of the New York Central Railroad between New York and Buffalo and the location of the base radio stations associated with the New York-Buffalo highway system is indicated in Fig. 2. It will be noted that the railroad route is covered by nine base transmitters and approximately twelve associated

[1] H. I. Romnes and R. R. O'Connor, "General mobile telephone system," *AIEE Trans.*, vol. 66, pp. 1658–1666; 1947.

base receivers. These are used for the train service. The transmitters employed frequency modulation and are rated at 250-watts output. In the particular installation described, each base-station transmitter employs a frequency of 35.66 mc and the base receivers accept a frequency of 43.66 mc from the mobile units.

Basically, the equipment on the train is similar to that on an automobile and the same type of radio transmitters and receivers are employed. There are, however, important differences. Special antennas to meet railroad clearance requirements are necessary. Arrangements are employed whereby the push-to-talk feature generally used in automobiles is eliminated. Also, arrangements are included to permit operation of the equipment by an attendant who must know the train's exact location and who is familiar with the general layout of the system.

Connection to Telephone Network

As previously explained, the transmissions to and from the mobile unit are combined and connected to the land lines in the control terminal. A schematic diagram of a control terminal as used in the railroad service is shown in Fig. 3. The terminal includes a conventional hybrid coil with a pad, amplifier, and vogad[2] in the transmitting branch of the circuit. The vogad operates to produce, within limits, full modulation of the radio transmitter at all times, regardless of the speech volumes applied to its input. In the receiving branch of the circuit there is an amplifier, noise reducer,[3] and a pad for adjusting the incoming speech levels. Signaling oscil-

[2] S. B. Wright, S. Doba, and A. C. Dickieson, "A vogad for radio-telephone circuits," Proc. IRE, vol. 27, pp. 254–257; April, 1939.

[3] C. C. Taylor, "Radio-telephone noise reduction by voice control at receiver," *Elec. Eng.*, vol. 56, pp. 971–974; August, 1937.

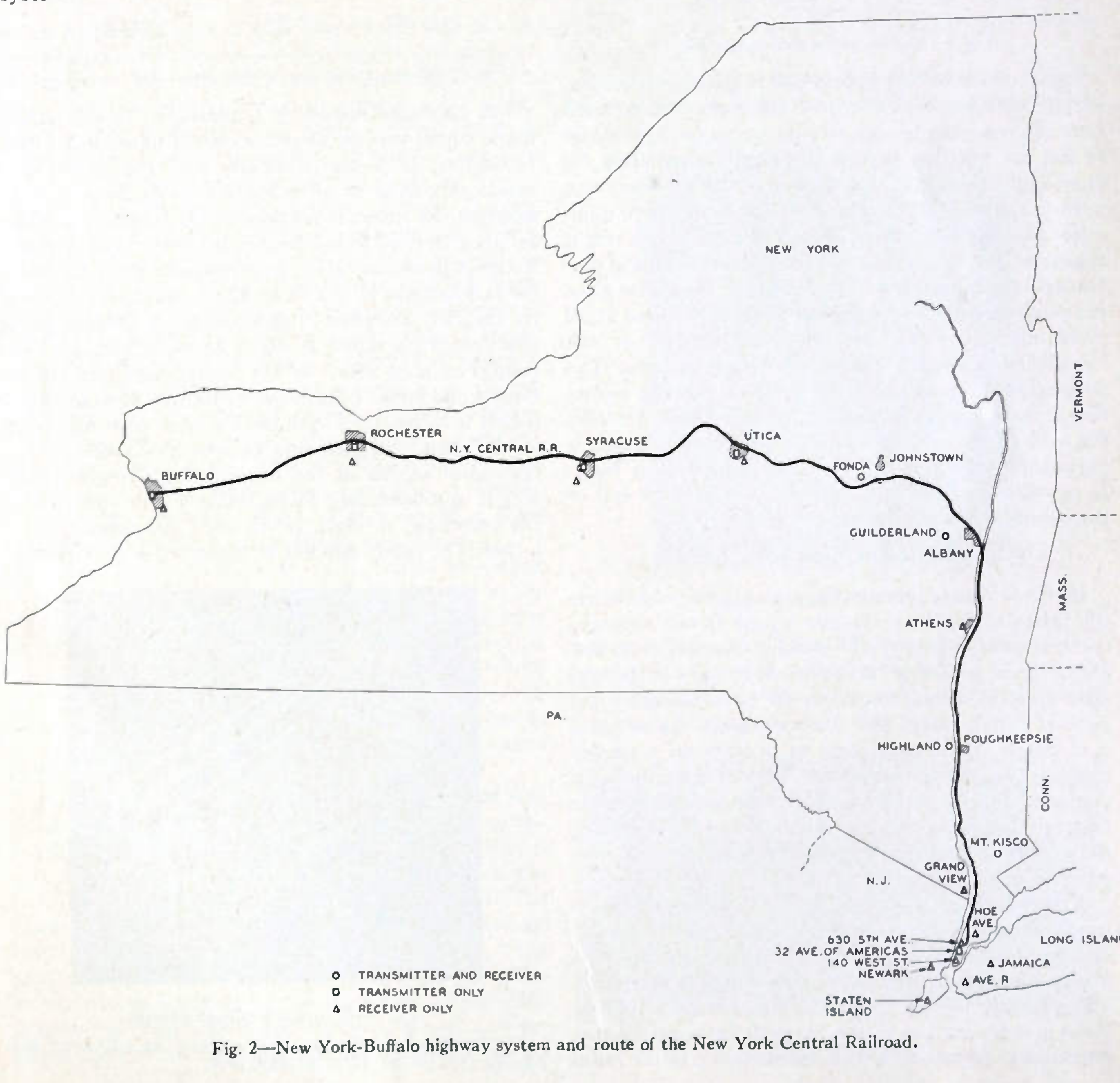

Fig. 2—New York-Buffalo highway system and route of the New York Central Railroad.

lators for generating the tones used for selective calling are included as a part of the control terminal and are connected to the transmitting wire line at the output of the vogad, as indicated. Certain other miscellaneous equipment units for checking the frequency of the transmitter and monitoring transmission are also included, but for simplicity are not shown in the figure.

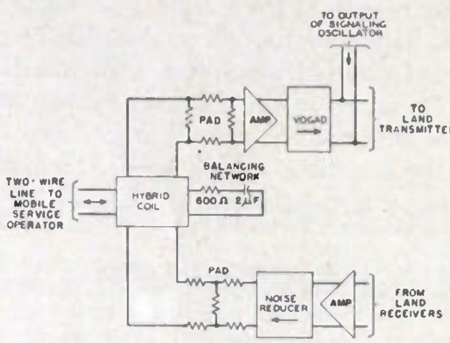

Fig. 3—Mobile-system control terminal.

Associated with the hybrid coil is a compromise balancing network which allows sufficient transmission from the receiving branch to pass across the hybrid coil to the transmitting branch to permit re-radiation for busy indications and for making calls between two push-to-talk mobile units. At the same time it minimizes the transmission trainward of noise which might appear at the output of the base receivers. This is particularly important in the case of duplex operation since the carrier from the mobile unit is being radiated at all times during a call and the base receiver is not "squelched" when the mobile customer is listening. The purpose of the noise reducer is to effect a further reduction in incoming noise. This device operates to diminish the noise in the intervals between speech sounds, and is particularly effective in improving transmission trainward when a low-volume talker is on the land end of the connection and the vogad gain is high.

Description of Train Installation

In the ordinary mobile units employed with either the urban or highway systems a push-to-talk arrangement is used, whereby the mobile transmitter is disabled during incoming transmission. It was not thought that such an arrangement would be satisfactory for general public usage, and the train installations have, accordingly, been engineered to operate on a duplex basis so that the telephone may be used in the ordinary manner. This is accomplished by providing separate transmitting and receiving antennas and by operating the mobile transmitter and receiver simultaneously instead of sequentially.

A train installation, including two antennas, two radio transmitters, two radio receivers, a relay panel, a power-distribution panel, an attendant's control unit, two telephone sets, and suitable interunit cabling, is illustrated schematically in Fig. 4. At the frequencies employed there is sufficient loss in the air path between the two antennas to permit simultaneous operation of the radio transmitter and receiver without densensitizing the radio receiver. Consequently, no filter, such as that found necessary in the 152 to 162-mc installations, is required.

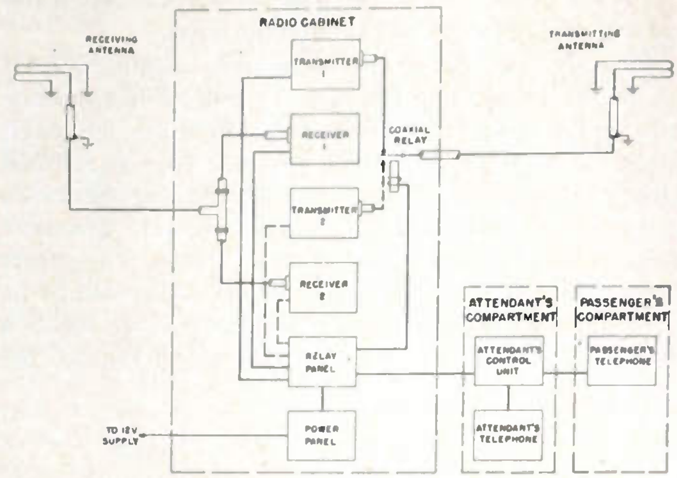

Fig. 4—Schematic diagram to 30 to 44-mc train installation.

For clearance reasons it is necessary to use an antenna whose vertical height is limited to not much over 14 inches. It is also desirable that the antenna be sufficiently rigid to provide a handhold for a man. In addition, all external portions of the antenna must be solidly grounded to the car roof for protection purposes. To meet these and other requirements a special antenna for railroad use in the 30 to 44-mc band was designed by the Bell Telephone Laboratories.[4] A photograph of this antenna is shown in Fig. 5. It consists of three elements, each of which is $11\frac{1}{2}$ inches high and are extended and folded over in the horizontal plane to form a folded inverted L. A trombone section is incorporated in the horizontal elements to tune the antenna. The horizontal section of the transmitting antenna (35.66 mc) is approximately 55 inches in length and that of the receiving antenna (43.66 mc) approximately 70 inches. The inside element of the antenna is connected

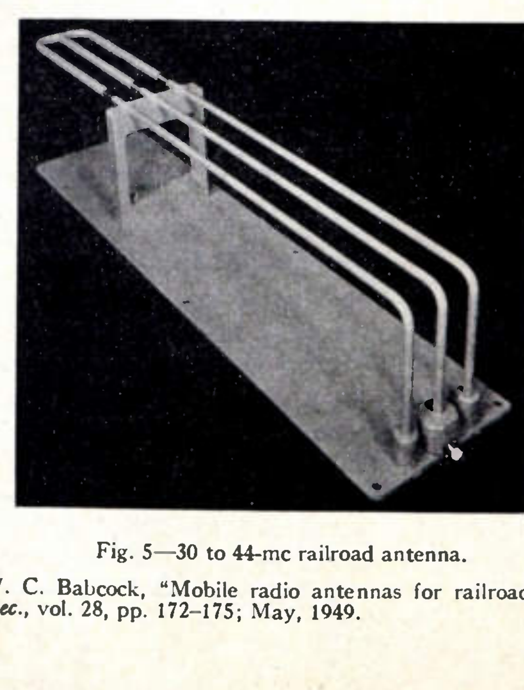

Fig. 5—30 to 44-mc railroad antenna.

[4] W. C. Babcock, "Mobile radio antennas for railroads," *Bell Lab. Rec.*, vol. 28, pp. 172–175; May, 1949.

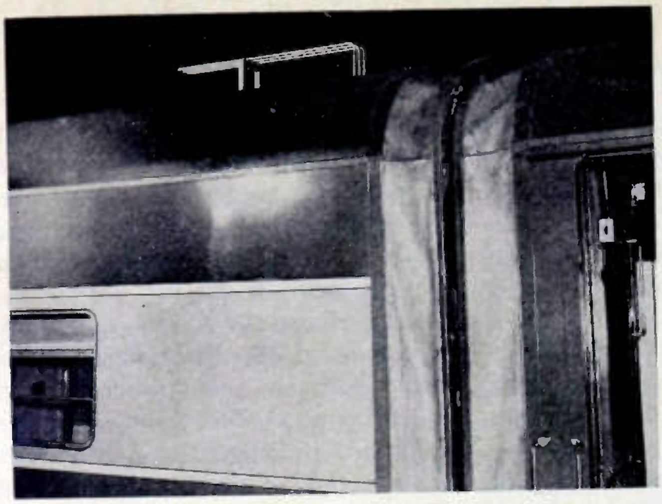

Fig. 6—Antenna installed on train.

Fig. 7—Attendant's position on the 20th Century Limited.

to the inner conductor of the coaxial cable lead-in and the two other elements are solidly connected to ground. The antenna has an impedance of approximately 70 ohms, and tests have indicated that the over-all radiation efficiency is about 1.5 db down with respect to a quarter-wave whip. A photograph of one of these antennas installed on a train is shown in Fig. 6.

The radio equipment proper consists of two Western Electric 39B Radio Transmitters and two 39A Radio Receivers. This equipment is the same type used in automobiles associated with the highway systems. Each transmitter and receiver operates from a 12-volt source. The transmitter has a radio frequency output of about 30 watts and the receiver has a sensitivity of about 1 microvolt. One transmitter and one receiver are paired to operate together as are the other transmitter and receiver. Two sets of equipment are employed in the experimental installations. This provides a flexible arrangement as the second set may be arranged to operate on a different frequency if found desirable at a later date. In the meantime it is available as a spare.

The equipment on the train is controlled by means of an attendant's control unit which is located at an attendant's position in the club car. Associated with the control unit is a relay panel which is mounted with the radio equipment and contains the various relays required for remotely operating the equipment from the attendant's control unit. A photograph of the attendant's position is shown in Fig. 7. It shows the control unit in position and attendant's telephone set.

Both the attendant's and passenger's telephone sets are standard Bell System wall type sets, suitably modified for 4-wire operation. Two small bells, one associated with each radio receiver output, are mounted in the attendant's telephone to indicate incoming calls. The passenger's telephone is mounted in a small sound-treated booth or compartment to reduce room noise and insure privacy. This telephone contains a buzzer which may be operated by the attendant. Connections between the relay panel and the control unit and between the control unit and the two telephones are effected by means of standard multiconductor, interior wiring cables.

Fig. 8 shows the attendant's control unit in more detail. It includes a lock-type power off-on switch and the

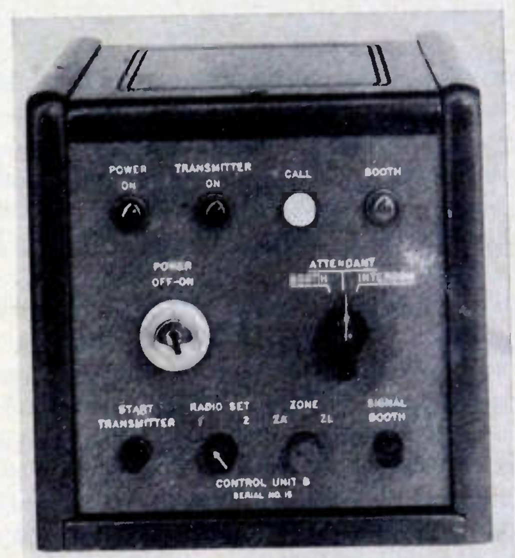

Fig. 8—Attendant's control unit.

necessary lamps and controls for operating the equipment. One important control is the button at the lower left corner of the unit. When momentarily depressed, this button energizes the transmitter, which is then

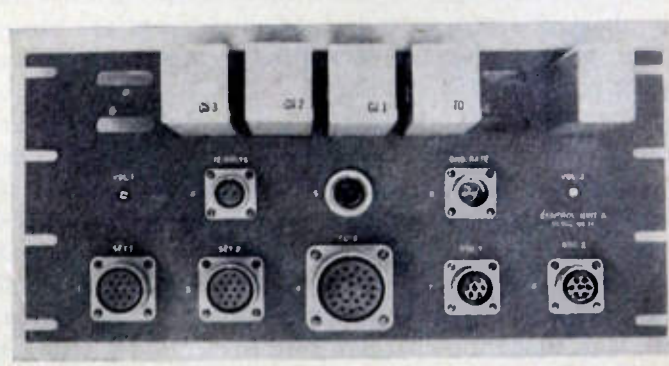

Fig. 9—Relay panel.

locked on under control of the attendant's and passenger's telephone switchhooks so that both handsets must be hung up to turn the transmitter off.

A photograph of the relay panel is shown in Fig. 9. It will be noted that all connections to the panel are made by means of plug-in connectors so that the complete unit may be readily removed and replaced should trouble develop in it.

The 12-volt supply is connected to the radio equipment through a power switch and fusetrons, both of which are mounted on the power-distribution panel. A photograph of this panel is shown in Fig. 10. Here again, external connections are made by means of plug-in connectors for ready replacement of the panel. An interesting feature of this panel is the utility connector in the lower right-hand corner. This provides a means for picking up the 12-volt supply for a trouble lamp or soldering iron.

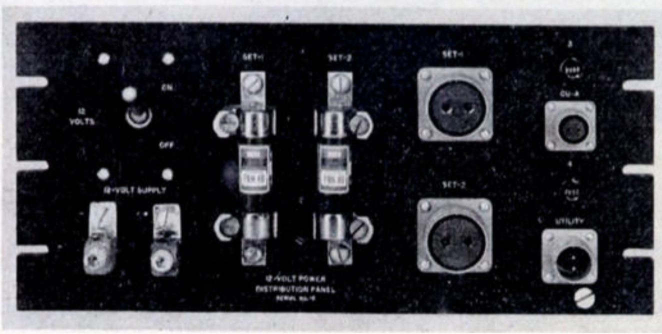

Fig. 10—Power panel.

Primary power for the radio sets and miscellaneous equipment is obtained from the battery on the car which normally supplies power for the lights and other electrical equipment. This battery is usually 32 or 64 volts. Since the radio equipment operates from a source of 12 volts, arrangements must be provided to obtain this voltage. These arrangements are supplied by the railroad, and three schemes were experimented with.

In the original installations on the Pennsylvania Railroad two 12-volt batteries were added. These were arranged so that at any given time one was connected to the radio equipment and the other to the 32-volt car battery for charging through suitable voltage dropping and regulating equipment. Thus, one battery was used while the other was on charge. At the end of each run the batteries were interchanged. Because of the heavy drain of the radio equipment this arrangement did not work out too satisfactorily.

For the New York Central a somewhat different arrangement was used. In this case a single 12-volt battery was installed for the radio equipment. This battery is charged from the car battery, which in this case is a 64-volt battery, through a vibrator, transformer, and rectifier. The 12-volt battery is trickle-charged at all times except when the radio transmitter is in operation. At such times the charging rate is increased, and this high rate is maintained until the battery voltage is restored to normal. This arrangement has proved reasonably satisfactory.

Experiments have also been made with motor generator sets operating from 32 volts with 12-volts dc output. Since the 32-volt battery fluctuates from some 24 to 40 volts, a high degree of regulation was required for these machines. To date their operation has been quite satisfactory even though they do require a certain amount of maintenance.

All of the equipment, except the telephone sets and attendant's control unit, is mounted in a small cabinet provided with forced ventilation and arranged so that the equipment is readily accessible for maintenance purposes. No shock mounting is provided for the radio equipment. The arrangement of the apparatus in one of the New Haven cars, is indicated in Fig. 11. This photograph shows only the lower part of the apparatus cabinet and one radio transmitter and receiver. The second transmitter and receiver are mounted directly above those shown on the two additional shelves.

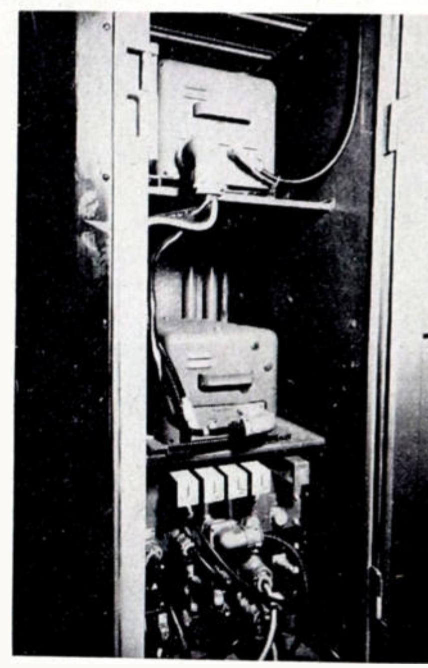

Fig. 11—Radio equipment installed on train.

Maintenance

In the design of the arrangements employed on the trains every effort was made to utilize, insofar as possible, the same equipment as is used on automobiles associated with the highway system and to make all units readily replaceable. At night the trains lay over at the terminals and minor repairs are effected by skilled telephone-company repairmen. Defective units are replaced and sent to repair shops provided by telephone companies for servicing automobile equipments.

An interesting result of the train service has been its effect on system maintenance. The repetitive daily movements of the railroad cars permit frequent comparisons of transmission in the same geographical locations. In this way system abnormalities and land-station troubles are more readily apparent to the operating forces than when calls are made from random spots at random times. As a result the train service has proved of considerable value in over-all system maintenance.

Radio-Transmission Considerations

When this project was initiated, tests were made in co-operation with the railroad to determine whether adequate coverage could be obtained from the existing highway system. During these tests occasional noise, fade outs, and flutter were noted, but, in general, transmission considered to be at least commercial was obtained over practically the entire route. Two additional base receivers were, however, found to be necessary.

With the new base-station equipment in operation and the final installations on the railroad cars completed, further tests were made to check the operation of the equipment and to determine definitely the boundaries of transmission from the various base stations for the use of the train attendants. Results of these final coverage tests are shown in Fig. 12. This figure shows the coverage, based on satisfactory transmission both trainward and landward, along the railroad from each highway station. In general, the coverage from adjacent base stations overlaps, or is contiguous, and depends primarily upon the location of the base stations with respect to the railroad right of way, the terrain involved, and the local noise conditions.

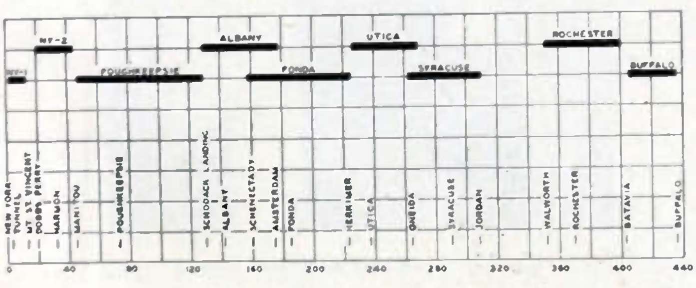

Fig. 12—Two-way coverage between New York and Buffalo.

An interesting feature, noted in the preliminary tests, was a type of relatively high received noise which was not experienced to any extent in previous tests in the 152 to 162-mc band. This noise occurs only under the following three conditions:

1. The train transmitter must be on the air.
2. The train must be in motion (or another train passing close by).
3. The train must be in an area of low received signal where the receiver is working with substantially full gain.

An investigation of this noise indicated that it was similar to a type of noise termed "stay noise" since it was first observed on ship installations and traced to the movement of the stays. The stay noise is created locally and is probably caused by the movement of metallic contacts and joints in the presence of the strong radio frequency field near the mobile transmitter. These moving parts became small generators which create noise over a wide band of frequencies, including those accepted by the receiver of the mobile unit. Measurements of stay noise indicated that it is approximately 25 db above set noise. It is not present on mobile units employing push-to-talk operation in which the transmitter is not energized during periods of reception.

Coin-Box Operation

Experience with attended train telephone installations on a number of railroads has indicated that the service fills a need and that it is generally satisfactory to the traveling public. The expenses of the attendant are, however, a major item in the cost to the railroad providing the service. Accordingly, experiments have been initiated utilizing coin-box telephones on the trains. One of these is being conducted on the two Congressional Limited trains of the Pennsylvania Railroad, one train running in each direction daily between New York and Washington. Service on these trains was originally established on an attended basis, and the installations are thus in the nature of a conversion of existing equipment rather than new installations.

The service between New York and Washington is operated through four urban mobile base stations and is not continuous, there being gaps in coverage between the several radio stations. From the viewpoint of service this is not desirable; however, advantage was taken of it in this particular installation since it permits control tones, which can be picked up and utilized on the train without co-channel interference, to be radiated from the base-station transmitters.

The system employed on the Congressional Limited functions as follows: A control tone of 4,100 cycles is arranged to be modulated on the carrier and transmitted continuously from each base station during the period the train is in range of that station. However, if the radio channel is busy, the 4,100-cycle modulation is removed. The 4,100-cycle control tone is received on the train in the regular radio receiver and compared with the received noise. If the tone-to-noise ratio exceeds a predetermined value, a lamp called the "service available lamp" is lighted on the train. If the tone-to-noise ratio is below this value or no tone is being received, the service-available lamp will not light. If the service available lamp is lighted, the circuit is so arranged that, when a customer removes the telephone handset from its switchhook, the mobile transmitter is energized and remains so until the telephone handset is replaced. If the service available lamp is not lighted, the customer, upon removing the handset, hears a busy tone which is generated locally on the train.

The telephone set employed is a standard Bell System coin collector, modified for 4-wire operation and arranged for postpayment operation with no provision for refunding coins. While arrangements for doing this have been devised, it was felt that, for experimentation, the additional complications would not be warranted.

When a customer desires to make a call, he first observes whether the service available lamp is lighted. If it is, he removes the handset from the hook and contacts the operator. When the operator has his party on the line, she requests the customer to deposit the charges, whereupon she completes the connection. If he talks overtime, it is necessary for the operator to ring him back and request the additional money.

The arrangement of the equipment on the train is shown schematically in Fig. 13. The figure shows the two antennas with their coaxial cable lead-ins and, in this case, a single radio transmitter and receiver. A coaxial stub filter is included in the lead-in to the receiver since, in this case, the installation operates in the 152 to 162-mc range. The control equipment, which contains the tone-to-noise ratio circuit and the several relays required, together with the busy tone generator, is indicated below the radio receiver. Underneath this unit is the power panel on which are mounted a power switch and fusetrons for protecting the radio equipment. To the right is the coin-box telephone and its associated bell box. Actually, there are two service available lights which operate in parallel. One of these is located directly above the coin box and the other outside of the telephone booth where it is visible to anyone in the car.

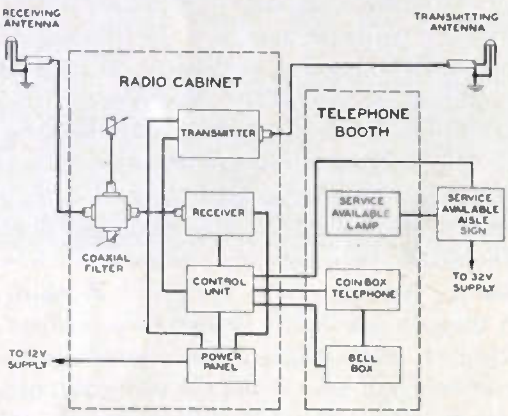

Fig. 13—Schematic diagram of coin-box installation on Congressional Limited.

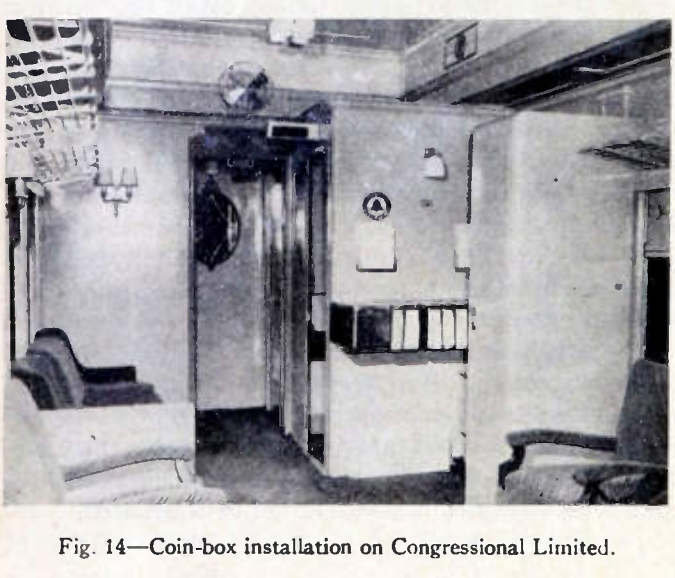

Fig. 14—Coin-box installation on Congressional Limited.

Fig. 14 shows a photograph of the installation as seen from the interior of the car. In the right foreground is the equipment cabinet. Immediately behind this are the telephone directories, located in the compartment formerly used by the attendant. Just beyond is the customer's booth with the service available lamp over the door and extending outward into the aisle.

The telephone booth is shown in Fig. 15. This photograph shows the coin-box telephone with the small service available lamp located directly above it and instructions for operating the telephone conveniently mounted for the benefit of the customer. Below the telephone is a writing shelf. A chair, not visible in the photograph, is also provided.

Fig. 15—Interior of telephone booth.

Besides the installations on the Congressional Limited two other coin-box installations have recently been made on the New Haven Railroad. In this case, a somewhat different arrangement is employed since the New Haven trains operate through a highway mobile system in which the transmission ranges of the several stations in the highway system overlap. Thus, the arrangement using control tones as employed in the Pennsylvania Railroad system could not be used. Instead, a very simple arrangement was devised in which the entire burden of operation is placed on the customer and the mobile service operator. No control tones or service available lamps are used. The customer determines by listening, whether or not the circuit is idle.

The operation of this system is as follows: To initiate a call from the train the customer first removes the handset from the switchhook and listens on the radio channel. If it is busy, he replaces the handset and tries again after a suitable interval has elapsed. If, on the other hand, he finds the channel idle, he momentarily depresses a "start-transmitter" button. This energizes the radio transmitter, sends out a carrier from the mobile unit, and brings in the mobile service operator. After the called party is on the line and the proper charges are deposited in the coin box, the operator completes the connection. At the conclusion of the call, when the customer hangs up, the radio transmitter is turned off and the equipment restored to normal.

Conclusion

Experiments so far have indicated that radio transmission between moving trains and Bell System mobile facilities along rail routes is generally satisfactory for public telephone service. A number of installations requiring an attendant on the train for their operation have been made. The public reaction to these installations has been favorable, but the railroads have found the cost of an attendant burdensome. Experiments using coin-box operation to eliminate the attendant are under way, and this method appears promising.

It should be noted, however, that a train telephone service of the type described, using either urban or highway channels, requires base-station equipment suitably located along the route of the railroad and a radio channel capable of handling the additional traffic. At the present time neither of these requirements is met for most of the railroad mileage in the United States. It appears, therefore, that provision of train passenger service on any extended scale will be dependent upon the availability of frequencies and the addition of the necessary base-station equipment.

Seven-League Oscillator*

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Summary—**A bridge-type *RC* oscillator is described which is continuously adjustable over a frequency range of 20 cps to 3 mc in one sweep of a two-gang linear potentiometer control. Tracking requirements of the two-gang control are not severe. The output is available in four phases, and the frequency is an approximately logarithmic function of the linear potentiometer setting. Practical limits of the frequency range are tentatively 0.01 cps and 10 mc. Accuracy of setting of the order of one per cent is attainable with ordinary components. Frequency stability is of the order of 2 per cent per db of tube gain variation.**

HERETOFORE a continuous sweep of a wide frequency range with a single oscillator control has been difficult to realize. Frequency bands of 3 to 1 ratio are obtainable with fixed-inductance variable-capacitance tuned oscillators for a 9 to 1 variation of capacitance. Still wider ranges are available with variable inductor tuning, and with variable resistance-capacitance tuning in configurations such as Wien bridges,[1] bridged-T circuits[2] and double-T circuits.[3] Continuous ranges of perhaps 1,000 to 1 may be obtained with variation of both resistance and capacity. Still other methods of achieving wide frequency ranges have been presented in different types of phase-shift oscillators.[4,5]

The heterodyne type of oscillator alone has been capable of frequency variation from the order of 1 cps or less to 1 mc or more in one continuous sweep. Because the output frequency at the low end of the range depends on the small difference of two large quantities, the stability is poor at low frequencies. Low frequency outputs also require elaborate precautions to prevent locking into step of the beating oscillators.

A considerable simplification of control and extreme widening of the frequency-band ratio is possible with a two-bridge circuit, in which three arms of each bridge may be resistive and the fourth arms are *RC* combinations capable of covering about four octaves for each two pairs of elements included, one pair in each bridge. Such bridges are readily designed to provide a nearly constant transmission at the desired frequency settings over wide ranges of adjustment. The frequency range is limited mainly by the tube gain available.

A simple version of an oscillator based on such bridges is shown in Fig. 1. The *RC* arms are shown as made up of two sections in series, but may be extended

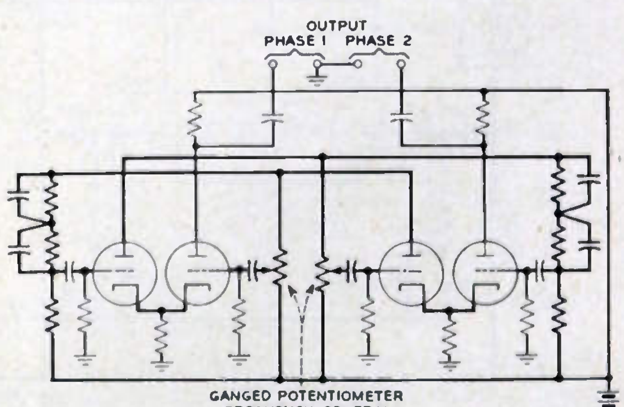

Fig. 1—Basic wide-range variable oscillator.

to eight or ten sections each to provide frequency ranges of a billion or more extreme ratio.

* Decimal classification: R355.914.3×R355.911.5. Original manuscript received by the Institute, March 16, 1950; revised manuscript received, October 26, 1950.

† Bell Telephone Laboratories, Inc., New York, N. Y.

[1] W. R. Hewlett, United States Patent No. 2,268,872, January 6, 1942.

[2] P. G. Sulzer, "Wide-range *RC* oscillator," *Electronics*, vol. 23, p. 88; September, 1950.

[3] H. H. Scott, "A new type of selective circuit and some applications," Proc. I.R.E., vol. 26, pp. 226–235; February, 1938.

[4] G. Willoner and F. Tihelka, "A phase-shift oscillator with wide-range tuning," Proc. I.R.E., vol. 36, pp. 1096; September, 1948.

[5] M. E. Ames, "Wide-range deviable oscillator," *Electronics*, vol. 22, p. 96; May, 1949.

Tuning is accomplished by adjustment of the two-gang potentiometer. Two outputs are shown, which are substantially in quadrature.

Fig. 2 shows a schematic of an oscillator which covers a tuning range of 20 cps to 3 mc. This is a simplified preliminary model. The *RC* networks shown in the bridges have six sections. The bridges shown are alike; it is not essential nor even desirable that both bridges be alike, as will be shown later.

The thermistor shunted across the second bridge in the plate circuit of tube *V*1 serves to limit the oscillation amplitude. The thermistor and *RC* network in the output circuit of tube *V*4 provide additional regulation of the output to a power amplifier.

A resistance potentiometer across the plate voltage supply provides a positive bias to the tube grids in order to permit large resistances in the cathode returns to improve frequency stability.

Fig. 3 shows the dial calibration of the oscillator of Fig. 2. The measured points oscillate slightly about a semilogarithmic asymptote over the low frequency range. The deviation from the straight line asymptote at high frequency is caused by parasitic capacitance, but it serves to spread out this part of the range somewhat.

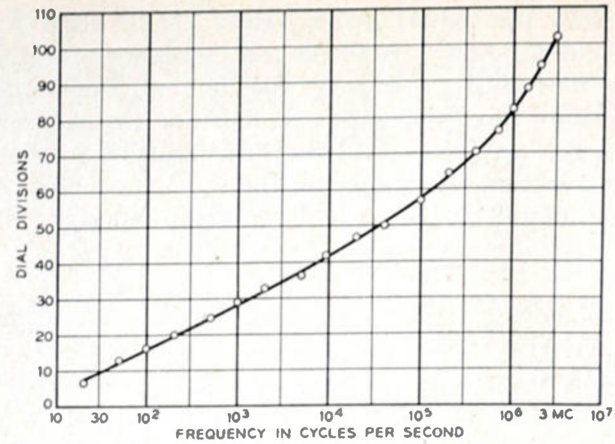

Fig. 3—Dial calibration.

Principle of Oscillator

The oscillator is based on the dissymmetrical bridge of Fig. 4. Three arms of the bridge are resistive. The fourth arm is composed of an *RC* network which provides over the whole frequency range a nearly constant reactive component of transmission to the output of the bridge. The resistive component of transmission through the *RC* network varies with frequency, and is balanced at the desired frequency of oscillation with equal resistive component from adjustable arms on the other side of the bridge. This leaves only the reactive component at the frequency of oscillation. Ninety-degree phase shift between bridge input and output results.

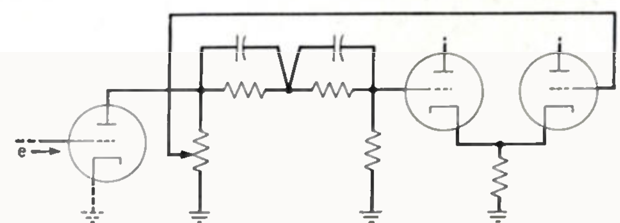

Fig. 4—A type of bridge useful in an oscillator.

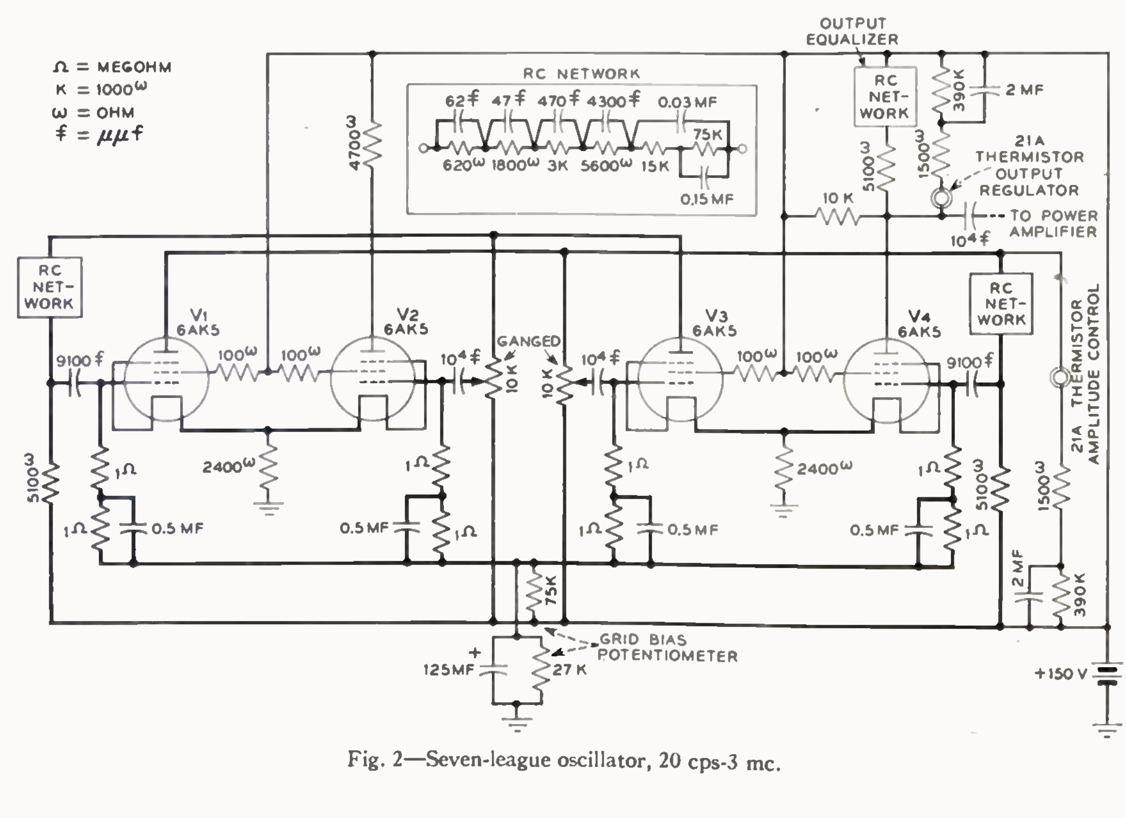

Fig. 2—Seven-league oscillator, 20 cps-3 mc.

The output of the bridge is split nearly equally between the grid and cathode of one tube, and the cathode and grid of the other tube, because of the large cathode coupling impedance employed. The outputs of the two tubes are nearly equal, and of opposite phase. One output of the required phase is connected to a second bridge, which is adjusted along with the first by a gang control, to provide a phase reversal and an additional ninety-degree phase shift. One of the outputs from this second bridge is fed back into the first bridge to complete the oscillator loop. The two quadrature phase shifts and the phase reversal provide the 360° or 0° phase shift required at the frequency of oscillation. The theory outlined is idealized, and can be approximated well enough if attention is given the networks and parasitics comprising a practical oscillator loop.

Bridge Design

The design of the adjustable bridge is based on circle diagrams of transmission through RC networks.

Fig. 5 shows a resistance R, tapped at aR and abR. The portion $(1-a)R$ is bridged with a capacitance C. The transmission ratio is

$$\frac{v_1}{e} = b\,\frac{a + j\dfrac{f}{f_c}}{1 + j\dfrac{f}{f_c}} \quad \text{where} \quad f_c = \frac{1}{2\pi CR(1-a)a}. \tag{1}$$

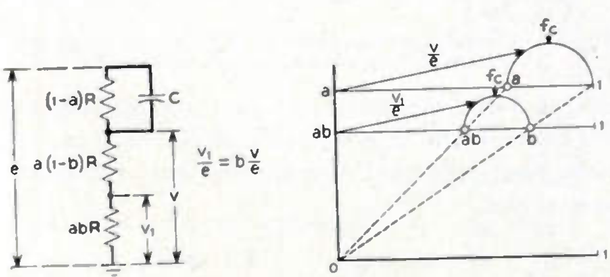

Fig. 5—Transmission to a resistance through an RC network.

Equation (1) is of bilinear form, and gives rise to a semicircular plot in the complex plane as in Fig. 5. The semicircles have been shifted along the imaginary axis to emphasize the scale factor b.

If the resistance aR is tapped at abR and the portion $a(1-b)R$ is bridged with a capacitance C_1, as in Fig. 6, we have for the transmission ratio

$$\frac{v_1}{v} = \frac{b + j\dfrac{f}{f_{c_1}}}{1 + j\dfrac{f}{f_{c_1}}} \quad \text{where} \quad f_{c_1} = \frac{1}{2\pi C_1 aR(1-b)b}. \tag{2}$$

If the critical frequencies f_c and f_{c_1} are chosen so that $f_{c_1} \gg f_c$, the transmission ratio for the combination of Fig. 7 tends to approach the semicircle between ab and b in the region near and below f_c, and the semicircle between b and 1 in the region near and above f_{c_1}. Between f_c and f_{c_1} the transmission ratio swings gradually from one semicircle to the other.

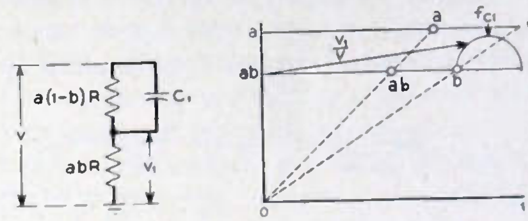

Fig. 6—Transmission to a resistance through an RC network.

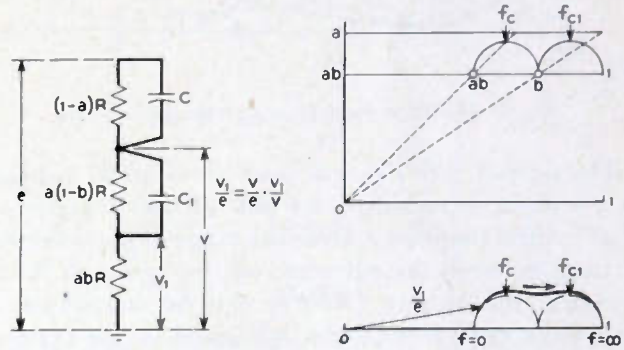

Fig. 7—Transmission to a resistance through an RC network.

If f_{c_1} is chosen such that $f_{c_1} = 16 f_c$, more or less, the transmission envelope will be a smoothly flowing envelope of the two semicircles of Fig. 7.

If the factors a and b are chosen so as to provide semicircles of equal diameters, the reactance component of the over-all transmission will be nearly constant between f_c and f_{c_1}.

Several RC combinations can be connected in series to extend the frequency range as in Fig. 8. Design relations for the RC network will be discussed later.

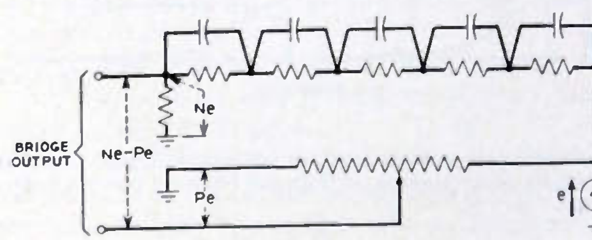

Fig. 8—Output voltage from an adjustable bridge.

If a variable frequency voltage e is connected to the RC network in Fig. 8, a voltage to be defined as Ne will be developed across the shunt arm of network. The reactive component of Ne will be nearly constant over a wide band of frequencies.

If a variable control, such as a resistance potentiometer, is connected to the same generator of voltage e, a voltage to be defined as Pe will be developed between the slider and the generator return. The control can be adjusted to obtain a voltage Pe equal to the in-phase component of the voltage Ne at any desired frequency.

This adjustment will be different for every frequency. At the frequency of in-phase component balance the output voltage (Ne-Pe) will be only the reactive component of Ne as in Fig. 9. This voltage is in quadrature with e, and nearly constant over a wide range of potentiometer settings at the corresponding frequencies of balance.

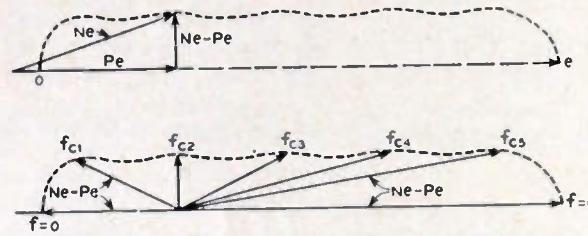

Fig. 9—Output voltage from an adjustable bridge.

The output (Ne-Pe) will lead the applied voltage e by an angle decreasing from 180° at zero frequency to 0° at infinite frequency. Over the range of potentiometer settings between the extreme reactive peaks of the RC network, the output (Ne-Pe) will be maximum for zero frequency, reduce to a minimum in the region of the 90° balance point, and increase toward another maximum at infinite frequency. At settings well beyond the extreme reactive peaks of the RC network, as for f_1 or f_2, the bridge output will exhibit a single peak with no dip to a minimum as in Fig. 10.

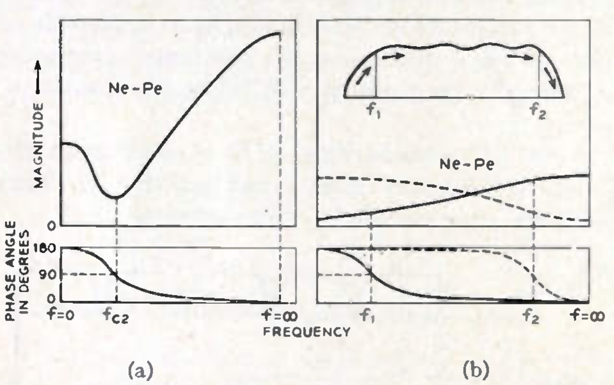

Fig. 10—Output voltage from an adjustable bridge. (a) Typical medium potentiometer setting. (b) Extreme potentiometer settings.

The effect of parasitic capacitance will first be considered for the potentiometer. The transmission through the potentiometer side of the bridge is determined by the tap k at low frequencies. At high frequencies the parasitic capacitance C in Fig. 11 bridged on the slider and its connected equipment changes the transmission as indicated in (3).

$$\frac{v}{e} = k\frac{1 - j\dfrac{f}{f_c}k(1-k)}{1 + \left(\dfrac{f}{f_c}\right)^2 k^2(1-k)^2}. \tag{3}$$

Equation (3) also gives rise to semicircle transmission plots as in Figs. 11 and 12. The curves for several frequency ratios f/f_c are plotted in Fig. 12 by constructing circles for $k(1-k)=0.25$, 0.2105, 0.177, 0.1485, and so forth (1.5-db intervals) and plotting intercepts with lines through the origin having slopes of these values. These lines represent v/e transmission ratios having phase angles

$$\theta = \text{arc tan}\left[-\frac{f}{f_c}k(1-k)\right].$$

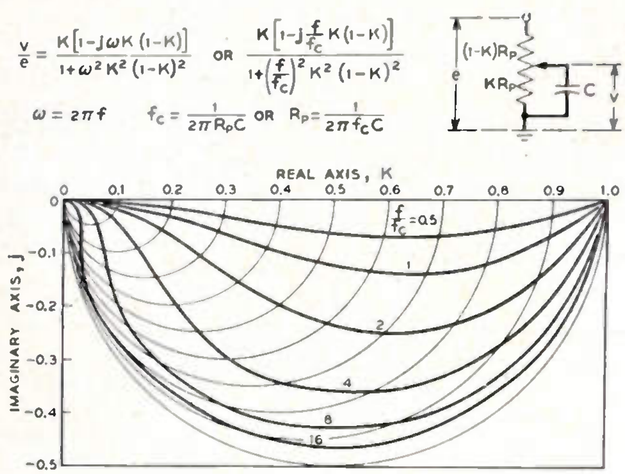

Fig. 12—V/e voltage ratio diagram for resistance potentiometer with capacitance between slider and ground.

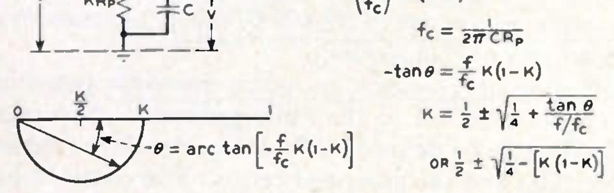

Fig. 11—Transmission through a capacitance bridged resistance potentiometer.

Table I in the appendix shows corresponding values of k and $k(1-k)$, over a range of four octaves, which are useful in plotting the curves of Fig. 12.

The effect of parasitic capacitance on the resistance arm connected to the RC network is considered next. In the frequency range where this capacitance is effective, the reactances of all the RC network capacitances

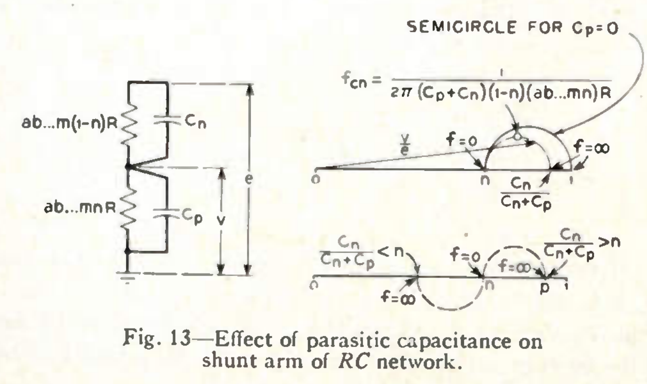

Fig. 13—Effect of parasitic capacitance on shunt arm of RC network.

but one, designated C_n in Fig. 13, are negligible. The associated resistances may be considered short-circuited for practical purposes. This leaves the resistances $ab \cdots m(1-n)R$ and $(ab \cdots mn)R$, bridged with capacitance C_n and the parasitic capacitance C_p.

The capacitance C_n across the resistance $(ab \cdots m)(1-n)R$ will form with C_p a potentiometer of some loss at infinite frequency, so that transmission will never become unity. The transmission for these two RC sections will be represented by (4), which is of the bilinear form, indicating a semicircular locus.

$$\frac{v}{e} = n \frac{1 + j\omega C_n R(ab \cdots mn)}{1 + j\omega(C_n + C_p)R(1-n)(ab \cdots mn)}. \quad (4)$$

If the infinite frequency transmission is less than the dc transmission, the semicircle swings below the real axis.

We can now consider the RC arm and the capacitance-shunted resistance arm connected to it. In the extension of the process used for Fig. 7 we aim to divide a resistance R into the desired number of sections $(n+1)$ as in Fig. 14. The first section lies between the top end of

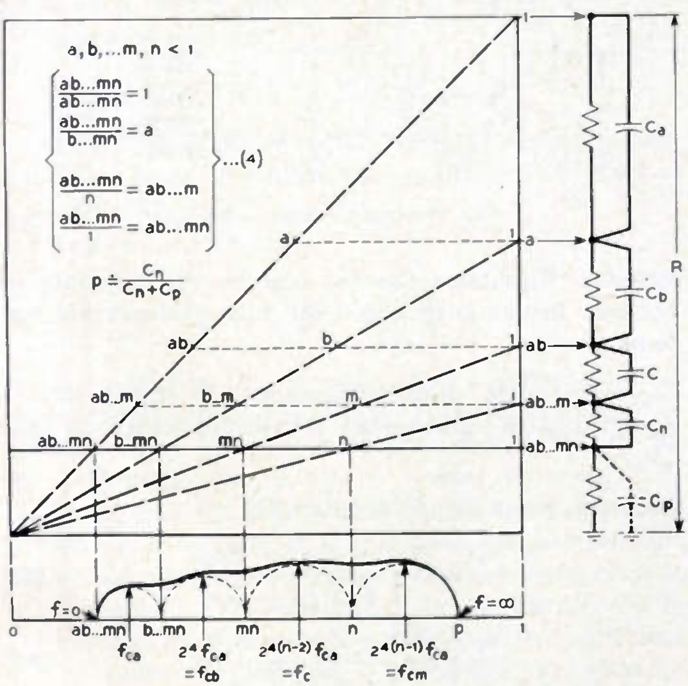

Fig. 14—Relations between transmission and elements of an RC network.

the resistance, at value R, and the tap, at value aR. The second section lies between the taps at values aR and abR. The last tap lies between the tap at value $(ab \cdots mn)R$ and the bottom end of the resistance, at value zero.

The first step is to lay down the semicircles required for the desired RC network transmission. These semicircles may be equal, but, as shown later on, should be tapered a bit.

The relations between the real axis intercepts of the component semicircles, n, mn, and so forth, and the taps aR, abR, and so forth, are shown in Fig. 14. The intercept p is set by trial, and may be readjusted if necessary after the completion of a trial design. The ratio of the two intercepts of the lowest frequency RC semicircle is $ab \cdots mn/b \cdots mn$ or a, of the next higher frequency section is b, and so on. Equation (5) gives a convenient method of evaluating the factors a, ab, and the like, which is to divide all the semicircle intercepts into the product $(ab \cdots mn)$, which may be set as desired

$$\left[\begin{array}{l} \dfrac{ab \cdots mn}{ab \cdots mn} = 1 \\[2ex] \dfrac{ab \cdots mn}{b \cdots mn} = a \\[2ex] \cdots\cdots\cdots\cdots \\[2ex] \dfrac{ab \cdots mn}{n} = ab \cdots m \\[2ex] \dfrac{ab \cdots mn}{1} = ab \cdots mn \end{array}\right]. \quad (5)$$

$$a, b, \cdots, m, n < 1$$

The critical frequencies are chosen, by proportioning of capacitances and resistances, to increase in geometric ratio, as f_c, $2^4 f_c$, $2^8 f_c$, $2^{12} f_c$, $\cdots 2^{4m} f_c$. The capacitance values are determined by equating reactance at the desired f_c to the resistance facing the capacitance, assuming all lower frequency RC sections to be shorted to a zero impedance generator, all higher frequency RC sections to be pure resistances.

The value of the lowest tap on the resistance R, at value $(ab \cdots mn)R$, should be in general less than the reactance of the shunting capacitance C_p at the highest frequency of oscillation. It should be kept large enough to sustain the gain required to overcome the bridge loss at the highest frequency. The total resistance R is not critical, and is determined by the values $(ab \cdots mn)R$ and $(ab \cdots mn)$ selected. It should be, in general, large, and may be set to fit with favorable values of capacitances as well as can be done.

Because the RC system shunts the paralleled potentiometer, as in Fig. 4, the impedance Z_{PL} presented to the tube driving the bridge falls as frequency increases. The transmission between the grid circuit of the driving tube and the output of a bridge with a constant reactive transmission will likewise fall. This falling off can be compensated by suitable shaping of the RC arm transmission as follows.

The plate load impedance Z_{PL} is approximately

$$Z_{PL} \doteq \frac{R_P(ab \cdots mn)\dfrac{R}{r}}{R_P + (ab \cdots mn)\dfrac{R}{r}} \quad (6)$$

where r is the transmission ratio of the RC system.

The voltage gain of the tube driving the parallel combination will be $g_m Z_{PL}$, where g_m is the transconductance of the tube. The transmission from the grid of the tube to the output of the RC network will be $N g_m Z_{PL}$, and should be unity or slightly larger. This requires that

$$N \geq \frac{1}{g_m Z_{PL}}.$$

This gives rise to a straight-line minimum boundary of the form

$$N = \frac{r}{(ab \cdots mn) R g_m} + \frac{1}{R_P g_m} = \frac{1}{g_m Z_{PL}}. \tag{7}$$

This is shown in Fig. 15.

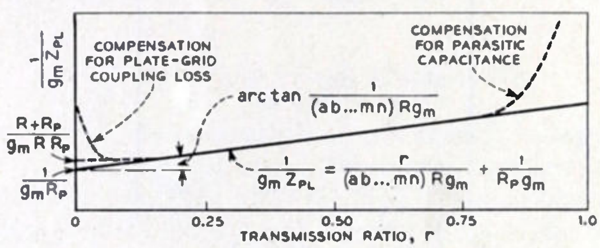

Fig. 15—Method of approximating required transmission characteristic of RC network.

This straight line is the minimum requirement for N. The semicircles representing the RC network are fitted under this line as a guide as in Fig. 16. Some margin should be allowed for network deviations and g_m variation. The effective g_m is reduced by about one half in the circuit, as will be shown later.

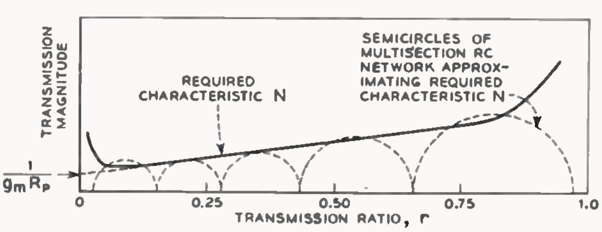

Fig. 16—Method of approximating required transmission characteristic of RC network.

The low-frequency end of the guide line is bent upward as required to allow for plate-grid coupling network loss. The high-frequency end is bent upward to allow for parasitic capacitance losses.

Potentiometer-RC Network Bridge Transmission

We are now ready to combine the transmission of the RC network and the potentiometer to derive the transmission through the bridge as in Fig. 17. The voltage output from the bridge (Ne-Pe) is determined, for a particular setting of the potentiometer, by the vector difference between the RC network output Ne and the potentiometer output Pe, over the $0-\infty$ frequency range. All settings of the potentiometer from $k=0$ to $k=1$ must be considered.

The transmission characteristics of Figs. 12 and 14 are combined in Fig. 17 to evaluate the transmission through the bridge. The curves of Fig. 17 show the magnitude and phase of this transmission.

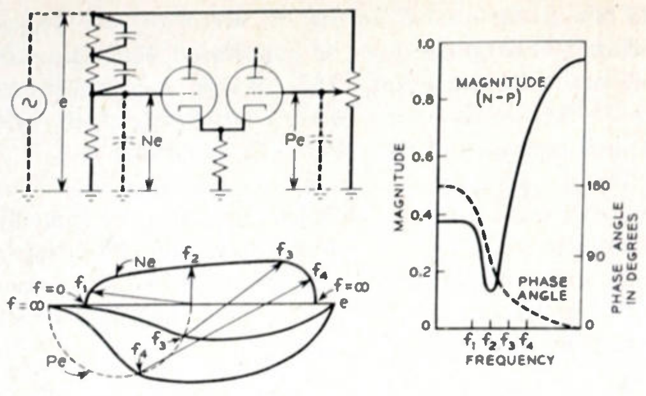

Fig. 17—Bridge output voltage (Ne-Pe) variation with frequency.

The transmission from the bridge through the tubes is next considered. Fig. 18 shows the voltage relations at the bridge output and the connected tube grid-cathode systems. Equations (8a–8d) express the transmission between bridge output and the tube grid-cathode systems.

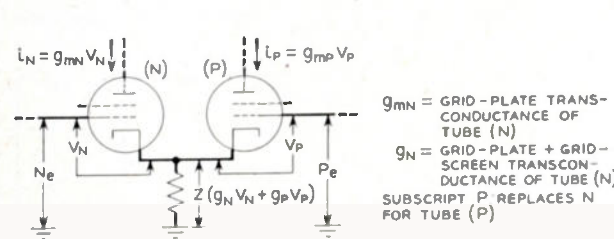

Fig. 18—Input voltage to tubes.

G_{mN} = grid-plate transconductance of tube (N).
g_N = grid-plate + grid-screen transconductance of tube (N).

Subscript $_P$ replaces N for tube (P).

$$\frac{v_N}{e} = \frac{N(1 + Z_{gP}) - P Z_{gP}}{1 + Z(g_N + g_P)}. \tag{8a}$$

$$\frac{v_P}{e} = \frac{P(1 + Z_{gN}) - N Z_{gN}}{1 + Z(g_N + g_P)}. \tag{8b}$$

For Z_{gN} and $Z_{gP} \gg 1$, and $g_N = g_P$, (8a) and (8b) reduce to

$$\frac{v_N}{e} \doteq \frac{N - P}{2} \tag{8c}$$

$$\frac{v_P}{e} \doteq \frac{P - N}{2} \quad \text{or} \quad -\frac{v_N}{e}. \tag{8d}$$

Cathode Coupling Impedance Factor

As shown in (8a) and (8b), the multiplying factors operating on the N and P vectors depend on the cathode impedance Z. These factors are readily analyzed by means of the approximation formulas given in (9a) and (9b). For $Z_{gP} > 1$, and $g_N = g_P = g_m$,

$$\frac{1 + Z_{gP}}{1 + Z(g_N + g_P)} \doteq \frac{1}{2}\left(1 + \frac{1}{2Zg_m}\right). \qquad (9a)$$

$$\frac{Z_{gP}}{1 + Z(g_N + g_P)} \doteq \frac{1}{2}\left(1 - \frac{1}{2Z_{g_m}}\right). \qquad (9b)$$

Normally g_N and g_P will be almost equal and (9a) and (9b) will show the effects of the Z_{g_m} factor for values greater than unity. Fig. 19 shows how this factor affects the multipliers of the N and P vectors of (8a) and (8b).

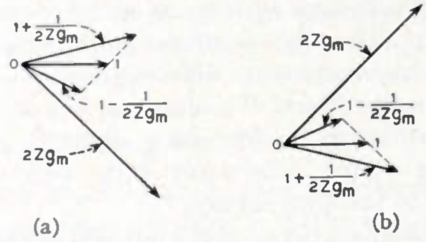

Fig. 19—Cathode impedance effects. (a) Z capacitive. (b) Z inductive.

For Z having a negative phase angle, an important high-frequency case, the factor Z_{g_m} likewise has a negative phase angle. The quantity

$$\frac{1}{2}\left(1 + \frac{1}{2Z_{g_m}}\right),$$

which multiplies the N vector in (8a), then has a positive phase angle, and rotates the multiplied N vector in a positive direction.

Likewise the quantity

$$\frac{1}{2}\left(1 - \frac{1}{2Z_{g_m}}\right),$$

which multiplies the P vector, has a negative phase angle, and rotates the multiplied P vector in a negative direction. The N vector is also lengthened slightly by the multiplying factor, the P vector shortened. The difference between the multiplied vectors, as expressed by (8a), is a measure of the transmission between the bridge output and a tube grid-cathode system, such as of $V1$ of Fig. 2. This difference $N-P$, as operated on by the multipliers, is increased by the capacitive impedance Z. These relations hold since N multiplied will ordinarily lead P multiplied by an angle between 0° and 180°, and not between 180° and 360°. However, in (8b), the roles of the P and N vectors are interchanged. The effect of a capacitance Z_{g_m} factor is to rotate the multiplied P vector in a positive direction, the multiplied N vector in a negative direction. The difference between the multiplied vectors, as expressed by (8b), is a measure of the transmission between the bridge output and a tube grid-cathode system, such as of $V3$ of Fig. 2. This difference $N-P$, as operated on by the multipliers, is decreased by the capacitive impedance Z. To compensate for this decrease the RC network driving tube $V4$ may be designed with a larger reactive transmission component at the highest frequencies.

The remainder of the oscillator loop design offers problems similar to those met in negative feedback amplifiers.[6] In the case of the oscillator, one frequency of oscillation is required. The network design problem over the 0–∞ frequency range, except for this one frequency of oscillation, consists of avoiding any tendency to oscillate at another frequency. This means that if the oscillator loop transmission passes through a 0° phase shift at any frequency except the desired frequency of oscillation, it must do so with a magnitude of less than unity, or 0 db. Exceptions may be made for conditional stability as shown by Nyquist,[7] although special measures may be necessary to establish oscillation in the desired mode.

The shaping of the transmission loop frequency characteristic is readily accomplished with proper choice of components. No further discussion is necessary.

Summary of Loop Transmission

Figs. 20 through 22 show transmission magnitude and phase for the feedback loop of the oscillator of Fig. 2 at extreme and middle frequency settings. Prominently demonstrated are the extremely broad bands, stretching well beyond the actual frequency adjustment range, in which negative feedback exists. The feedback magnitudes range up to some 50 db.

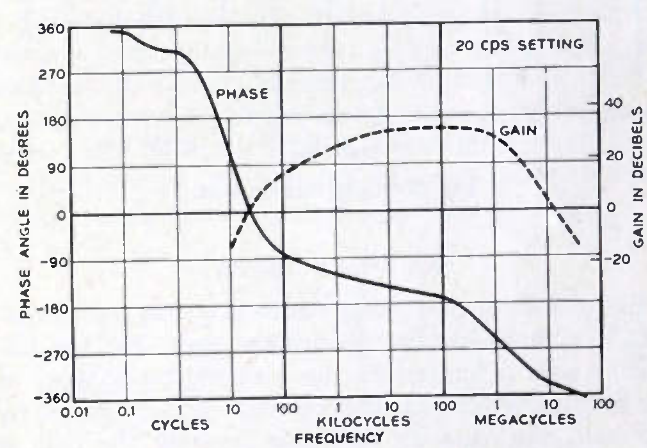

Fig. 20—Loop transmission.

Smoothing of Transmission Ripples

The ripples in the RC network transmission can be compensated somewhat by staggering the critical frequencies f_C of the two RC networks. The configuration of the RC arm also may be modified to advantage, and the frequency ratios of 16 to 1 for successive sections may be changed.

Transit-Time Effects

At frequencies in the megacycles, tube transit times add to the loop phase shift, and cannot be neglected.

[6] H. W. Bode, "Network Analysis and Feedback Amplifier Design," D. Van Nostrand Co., New York, N. Y., 1945.

[7] H. Nyquist, "Regeneration theory," *Bell Sys. Tech. Jour.*, vol. 11, p. 126; January, 1932.

For a 6AK5 tube this factor may add a phase shift of the order of 0.2° per mc per stage. The physical size of the loop also contributes, but is a smaller factor in a normal design.

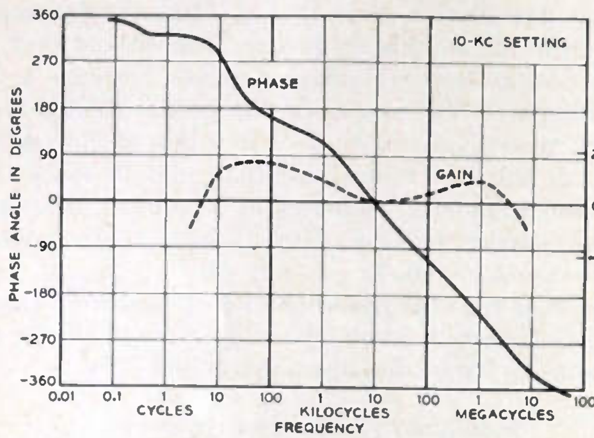

Fig. 21—Loop transmission.

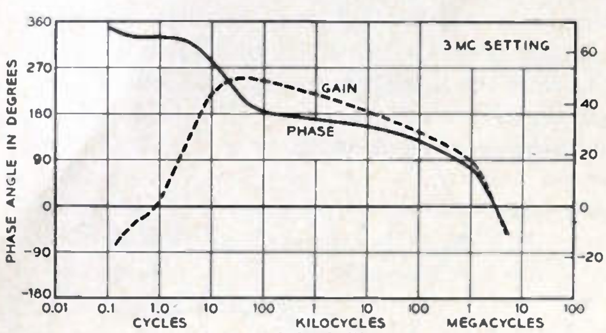

Fig. 22—Loop transmission.

Frequency Range

Extreme frequency range depends on the gain available to compensate for the bridge losses. The top frequency gain is limited by the parasitic capacitance of the amplifier system. If the top frequency is reduced by one half, the capacity reactance limiting the gain is doubled, and the realizable gain is doubled too. The bridge loss then can be doubled by halving the semicircle diameters. Twice as many semicircles can be accommodated and the extreme highest-to-lowest frequency ratio is roughly squared. The lowest frequency can be reduced decades below 1 cps to regions in which insulation resistance will be a limiting factor.

Frequency Calibration of Dial

The frequency calibration of the dial is nearly a logarithmic function of linear scale divisions with linear control potentiometers. This relation arises from the nearly linear variation of the resistive component of the *RC* network impedance with the logarithm of frequency.

The linear asymptote through which the real component of the *RC* arm transmission oscillates would give rise to a logarithmic frequency adjustment dial scale, but for several factors. Most prominent are that the *RC* arm envelope asymptote has a slope and that the plate grid coupling networks and parasitic capacitance introduce phase and magnitude changes as the extreme frequencies are approached.

Phase Angle Between Outputs

The phase angle between stage outputs is nearly 90°. This arises from the 180° reversal in the two bridges, leaving 180° of phase shift to be split between the two bridge units. The split is almost even, with some unbalance caused by the two different multiplying factors operating on the *N* and *P* vectors (see equations (8) and (9)). Deviations of the frequency-adjusting potentiometers from perfect alignment with each other also contribute to the dissymmetry.

The phase angle between the outputs of the *N* and *P* tubes of the same stage is roughly 180°, which may be modified by the different multipliers of the *N* and *P* vectors, and different plate load impedances.

Thus the outputs of $V1$ and $V2$ in Fig. 2 are roughly opposite in phase, as are those of $V3$ and $V4$. The output of $V3$ is displaced roughly 90° from that of $V1$. It follows that the outputs of $V2$ and $V4$ are similarly related. Thus a four-phase output, with phases distributed at roughly 90° intervals, may be derived from the four plate circuits of the tubes.

Direct connection may be made to plates of tubes $V2$ and $V4$ without seriously disturbing the oscillator. Connection to either of plates of $V1$ and $V3$ preferably is made through a buffer amplifier to avoid reaction on the oscillator from different connected loads.

Alignment of Frequency-Adjusting Potentiometers

The effect of misalignment of the frequency-adjusting potentiometers results in a frequency roughly halfway between those corresponding to both potentiometers set in turn to align with each other.

Amplitude Stabilization

If the oscillator output is limited by extreme tube overloading, the wave form is degraded. It is preferable to use some other sort of gain control, such as grid-bias automatic voltage control, or a thermistor as shown in the plate circuit of tube $V1$ of Fig. 2. The resistance bypassing the condenser in series with the thermistor reduces the voltage peak of the thermistor characteristic by providing a dc bias current. The resistance in series with the thermistor serves to adjust the slope of the regulation characteristic.

The oscillator output is taken from the plate circuit of $V4$. A second shunting thermistor circuit helps to reduce the output voltage variations. Further reduction of these variations over the frequency range is provided by an RC network, similar to those in the bridges, shunted across the plate load of the tube $V4$. An article on fundamentals of thermistors has appeared recently in this journal.[8]

Harmonic Output

Harmonic output is proportional to the excursions of the tube currents. The tube system will act as an amplifier being driven by the fundamental frequency of the oscillator, and the harmonics incidental to the fundamental output will be present. In general, the smaller this output, the smaller the distortion.

Frequency Stability

Frequency stability is dependent on many factors, the most prominent of which is the tube gain stability. The nonlinearity of the transmission loop also contributes. Factors such as stability of the frequency-control elements will not be discussed here.

Fig. 23 shows typical frequency shifts for interchanges of $V1$ and $V2$ or $V3$ and $V4$ in cases where the transconductance of $V1$ is 12 per cent higher than that of $V2$. The frequency shift is 5 per cent per db of tube gain change for the simple cathode impedance of a resistance and a small coil, bridged with parasitic capacitance. Above 100 kc the parasitic capacitances in the circuit begin to exact toll, and frequency stability becomes poorer.

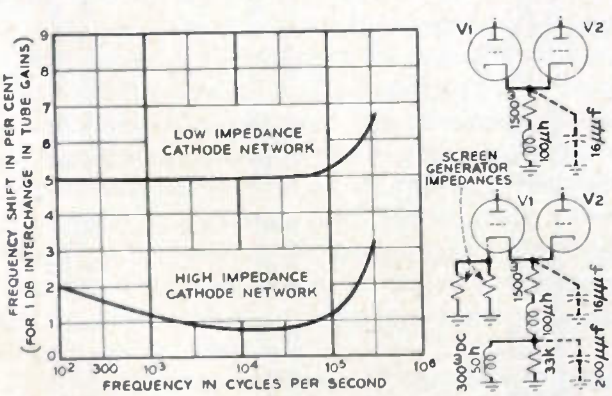

Fig. 23—Frequency stability with gain variation.

Increase of the cathode impedance by a more complex network improves the frequency stability. The tube screen grid generator resistances tend to limit the maximum impedance and the improvement possible.

The tube gain variation is the largest source of frequency drift, and demonstrates the need for a large cathode feedback. The components of the the transmission ratio v_N/e, shown in Fig. 24, indicate how the phase shift resulting in frequency change is derived from changes in the multiplied N and P vectors, N_1 and P_1.

[8] J. H. Bollman and J. G. Kreer, Jr., "Application of thermistors to control networks," Proc. I.R.E., vol. 38, pp. 20–26; January, 1950.

Equations (10a) and (10b) give the changes in the components of the effective grid-cathode voltages v_N and v_P for changes in the tube gain factors g_N and g_P.

$$d\frac{(v_N)}{(e)} = dN_1 - dP_1. \tag{10}$$

For $g_P = g_N = g_m$,

$$dN_1 = \frac{N_1}{2}\left[\left(1-\frac{3}{2Zg_m}\right)\frac{dg_P}{g_P} - \left(1-\frac{1}{2Zg_m}\right)\frac{dg_N}{g_N}\right] \tag{10a}$$

$$dP_1 = \frac{P_1}{2}\left[\left(1+\frac{1}{2Zg_m}\right)\frac{dg_P}{g_P} - \left(1-\frac{1}{2Zg_m}\right)\frac{dg_N}{g_N}\right]. \tag{10b}$$

The differential factors operating on the components of v_N/e as shown in the last two equations are almost equal, and become more so as $Zg_m \doteq \infty$. The variation coming from g_P contributes to the inequality, and so to the frequency shift. The contribution from g_N tends to cancel. This would suggest special attention in the way of regulation of supply voltages and perhaps a negative feedback for the P tube. Similar measures are indicated for the N tube of the pair using the output of the P tube to drive a bridge.

The high dc resistance of the common cathode return tends to keep the total current of the N and P tubes constant. If the current in one tube starts to fail, the current in the other tube increases to make up the difference. Since transconductance is somewhat proportional to space current, the common cathode connection tends to accentuate frequency drift caused by tube aging.

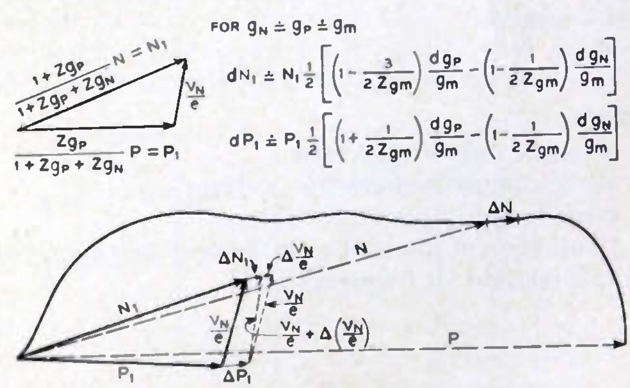

Fig. 24—Frequency stability with gain variation.

Fig. 25 shows separate cathode resistances for each tube. The cathodes are coupled together through condensers which reduce the performance to that of the paralleled cathode circuit impedances at sufficiently large frequencies. The large dc feedbacks acting on each tube separately tend to hold the space currents constant. The transconductances tend to stay more nearly constant over much of the tube life, and the oscillator frequency is thus more stable.

The cathode coupling condensers enter into the low-frequency transmission around the oscillator loop, and must be designed with the cathode return impedances and grid-plate coupling systems to make a stable feedback gain and phase characteristic. The cathode returns may be complex networks instead of plain resistances to provide large impedances in the oscillator working band.

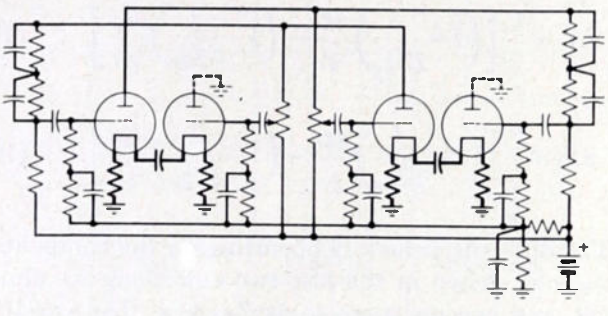

Fig. 25—Direct-current stabilization of gain and frequency.

Oscillator Frequency Shift as a Function of Output Amplitude

Frequency shift of an amplitude-regulated oscillator as a function of output well below overload may be evaluated by means of a power series analysis. The frequency shift is caused by phase shift of the fundamental frequency resulting from quadrature components of the second-order difference product of fundamental modulating with second harmonic, and the third-order difference product of fundamental modulating with third harmonic, and so forth. For small amplitudes of fundamental, say 30 db below full output, only the second-order modulation is noticeable, and a two-term power series suffices

$$I = I_1 + a_2 I_1^2 \tag{11}$$

where

I = total current in a tube

I_1 = fundamental frequency current

a_2 = a factor of proportionality.

Evaluation of the actions in the feedback loop results in this relation for frequency shift

$$\Delta f = -\left.\frac{df}{d\theta}\right|_{f=f_1}\left[2\,\frac{I_2^2}{I_1^2}\,\frac{j\sin\phi_2}{|1-A_2|}\right] \tag{12}$$

where

Δf = frequency shift in cycles

$d\theta/df$ = slope of phase-frequency characteristic of feedback loop in radians per cycle

I_2 = second harmonic current corresponding to I_1, measured with no feedback

A_2 = feedback to second harmonic.

$$\frac{1}{1-A_2} = \frac{\cos\phi_2}{|1-A_2|} + \frac{\sin\phi_2}{|1-A_2|}\,.$$

$d\theta/df$ is negative at f_1; for θ_2 between 0° and −180°, $\sin\phi_2$ is negative and Δf is negative. Thus increasing amplitudes of I_1 result in a downward shift of frequency, proportional to the square of the amplitude, as long as I_2 is proportional to the square of I_1. This holds for small amplitudes. For larger amplitudes the power series of (11) must include more terms.

Appendix

TABLE I

$$k = \frac{1}{2} \pm \sqrt{\frac{1}{4} + \frac{\tan\theta}{\dfrac{f}{f_c}}} = \frac{1}{2} \pm \sqrt{\frac{1}{4} - k(1-k)}$$

$k(1-k)$	k	
0.25	0.50	0.50
0.21	0.30	0.70
0.177	0.23	0.77
0.149	0.18	0.82
0.125	0.15	0.85
0.105	0.12	0.88
0.088	0.10	0.90
0.074	0.08	0.92
0.063	0.07	0.93
0.053	0.06	0.94
0.044	0.05	0.95
0.037	0.04	0.96
0.031	0.03	0.97
0.026	0.03	0.97
0.022	0.02	0.98
0.019	0.02	0.98

Recording Demagnetization in Magnetic Tape Recording*

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Summary—An analysis of the magnetic tape recording process employing supersonic excitation is presented by considering the effect of the spatial distribution of the magnetic field around the recording head air gap on the magnetic history of an unmagnetized element of tape as it tracks across the recording head. This leads to an effect which is termed "recording demagnetization" and serves to explain certain performance characteristics. An experimental technique developed for the measurement of this recording demagnetization is described, as is the method of measuring the air-gap field distribution. Finally, the correlation of the measurements of the recording demagnetization with normal recording performance characteristics is reported.

Introduction

RAPID PROGRESS in the development of practical magnetic sound recording has been achieved in recent years. The use of a supersonic excitation or bias in the recording process has contributed substantially to the high fidelity attained in present systems, but has likewise increased the complexity of the recording process for which several theories have been proposed in the literature. Noteworthy among these is the work of such men as Toomin and Wildfeuer,[1] Holmes and Clark,[2] Wetzel,[3] and Camras.[4] However, most of this work was restricted to a steel wire medium and none included the effect of the spatial distribution of the magnetic field around the air gap of the recording head. The magnetic field measurements of Clark and Merrill[5] also were made on wire recording heads and were used quantitatively in an analysis of the reproduction process. The analysis proposed herewith pertains primarily to the effect of the magnetic field distribution of the recording head air gap on the recording process when the medium is a magnetic tape undergoing longitudinal magnetization.

In order that the analysis and corroborating measurements may have practical significance, the performance characteristics of Figs. 1 and 2 are presented. Curves A, B, and C of Fig. 1 are the frequency-response curves for a constant recording current, and for recording and playback speeds of 4.4, 8.3, and 16.4 inches per

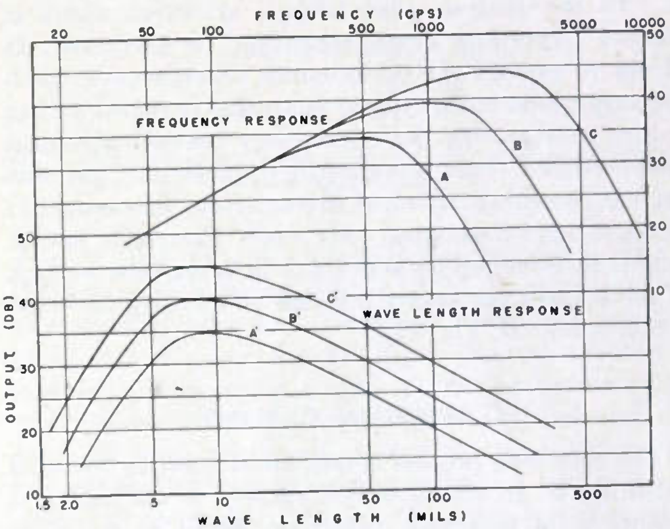

Fig. 1—Frequency and wavelength response characteristics for constant current recordings. Curves A and A'; B and B'; C and C', are for tape speeds of 4.4; 8.3; and 16.4 inches per second, respectively.

second, respectively. Curves A', B', and C' show the same data considering the wavelength of the recorded signal on the tape as the independent variable. It follows directly from curves A', B', and C' that, for constant current recording, the output varies directly with

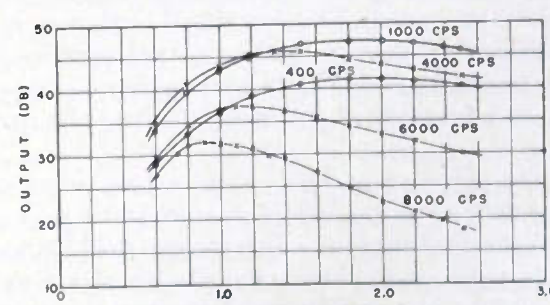

Fig. 2—Experimental curves showing variation of signal output voltage as a function of the magnitude of the supersonic excitation with signal recording current and tape speed both held constant.

the tape speed, and the factors influencing the shape of the characteristics are independent of tape speed. It was considered advantageous therefore, in the analysis of the recording and reproduction processes, to use the physical wavelength on the tape as the independent variable, for then the tape speed factor was eliminated.

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[1] H. Toomin and D. Wildefeuer, "The mechanism of supersonic frequencies as applied to magnetic recording," PROC. I.R.E., vol. 35, pp. 664–668; November, 1944.

[2] L. C. Holmes and D. L. Clark, "Supersonic bias for magnetic recording," *Electronics*, vol. 18, pp. 126–136; July, 1945.

[3] W. W. Wetzel, "Review of the present status of magnetic recording theory, Part II," *Audio Eng.*, vol. 31, pp. 12–16; December, 1947.

[4] Marvin Camras, "Graphical analysis of linear magnetic recording using high frequency excitation," PROC. I.R.E., vol. 37, pp. 569–573; May, 1949.

[5] D. L. Clark, and L. L. Merrill, "Field measurements on magnetic recording heads," PROC. I.R.E., vol. 35, pp. 1575–1579; December, 1947.

The curves of Fig. 2 illustrate how the signal output voltage of different frequencies varies with the magnitude of the supersonic excitation, all other factors remaining constant. Two features of these characteristics are to be noted. The maximum output occurs for different excitation for different signal frequencies, moving toward lower excitation currents as the frequency is increased, and the output of the higher frequency signals is substantially reduced for higher excitation currents. This latter effect has been reported by Holmes[6] and Wetzel,[7] but no satisfactory explanation has been presented before this time.

In the analysis that follows, an effect which is termed "recording demagnetization" is predicted. It serves to explain the performance characteristic when the magnitude of supersonic excitation is varied and to reduce considerably the discrepancy between the ideal and observed short wavelength response for constant current recording. By ideal characteristic is meant that characteristic depending only upon Faraday's law of magnetic induction for which a drop in response voltage of six decibels per octave increase of wavelength would result (curve *D*, Fig. 13).

Theoretical Analysis

The recording process is best understood by focusing attention on an unmagnetized element of the tape and following its magnetic history in detail as it tracks across the recording head. On the assumption of sinusoidal signals and excitations, this history is a function of the speed of the tape, the amplitude, frequency and phase of the signal, the amplitude, frequency and phase of the supersonic excitation, the nonlinear magnetic characteristic of the tape, and the magnetic field distribution around the recording head air gap. An immediate simplification is effected by combining the speed and frequency into a wavelength function as was previously suggested. The functional variation of the magnetizing field that an element of tape experiences in moving across the recording head then may be expressed merely as a function of the distance along the path of the tape.

For the purposes of this paper, a long wavelength signal is defined as one whose wavelength is long compared to the distance over which the gap field has an appreciable value and a short wavelength signal as one whose wavelength is less than the effective air-gap field. The composite variation of the magnetizing force with distance across the recording head combining a long wavelength signal of fixed amplitude and phase, a supersonic excitation of a given amplitude and phase, and the gap field distribution is illustrated by curve *oabc* $\cdots$ *n*, in Fig. 3(a). The effect of the long wavelength signal is such as to produce an asymmetry in a given (positive

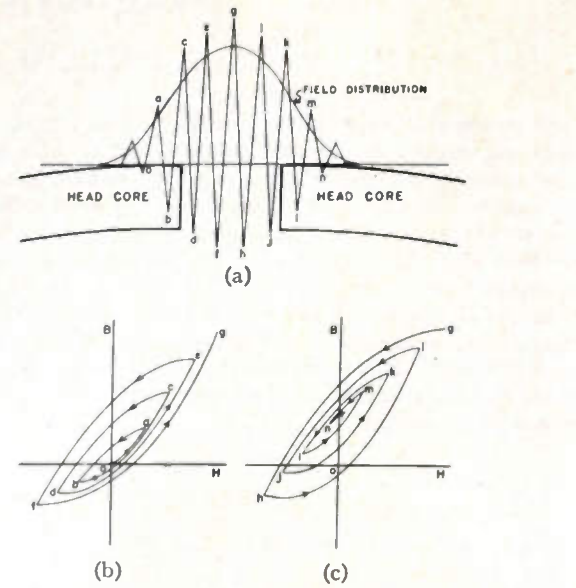

Fig. 3—(a) Schematic representation of the magnetic field variation across the recording head air gap during the recording process of a long wave-length signal. (b) The magnetizing sequence (not to scale) experienced by an unmagnetized tape element as it moves to the center of the air gap. (c) The magnetic sequence as the tape element leaves the center of the air gap.

for the phase shown) direction. For a long wavelength the signal amplitude varies little across the gap length and the effect is essentially that of a constant magnetizing force, its magnitude varying only as determined by the gap field distribution. A corresponding relationship is shown in Fig. 4(a) for a short wavelength signal of the same amplitude and phase and with the same supersonic

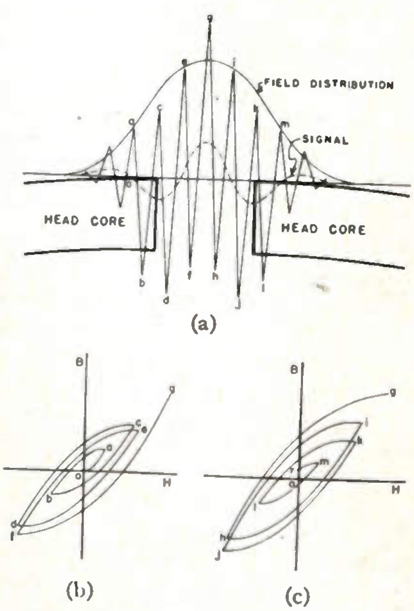

Fig. 4—(a), (b), and (c). The corresponding representations of Fig. 3 for a short wavelength signal.

[6] Lynn C. Holmes, "Some factors influencing the choice of a medium for magnetic recording," *Jour. Acous. Soc. Amer.*, vol. 19, pp. 395–403; May, 1947.

[7] W. W. Wetzel, "Review of the present status of magnetic recording theory, Part III," *Audio Eng.*, vol. 32, pp. 26–30; January, 1948.

excitation. Several cycles of the short wavelength signal extend over the effective air-gap field region producing corresponding reversals in the direction of the asymmetry, the peak values of which vary in accordance with the gap field distribution. In both cases a phasing has been assumed such that both the signal and supersonic excitation have a positive maximum at the center of the air gap.

The effects of applying the resulting magnetic field variations (curves $oab \cdots m$) of Figs. 3(a) and 4(a)) to the magnetization curve of the tape are illustrated in Figs. 3(b) and (c), and 4(b) and (c), respectively. An element of unmagnetized tape then in moving to the center of the air gap experiences a magnetic history shown by the lines *oabcdefg* and in moving from the center out of the air gap experiences a history shown by lines *ghijklmr*. Separate diagrams are used to illustrate the incoming and outgoing of the tape element so as to reduce confusion in following the curves. The remanence of the element of tape at this point in the recording process is represented by *Or* in Fig. 3(c) for a long wavelength signal, and by the smaller value *Or*, Fig. 4(c), for a short wavelength signal, the smaller value resulting from the asymmetry reversals. Both of these values are below the residual induction of a symmetric cyclical hysteresis loop having the same peak value of induction. This reduction in remanence due to the complex magnetic history of a tape element as it traverses the air gap is designated herewith as "*recording demagnetization.*"

It follows from the above discussion that the recording demagnetization is a function of the amplitude, wavelength and phase of the signal current, the amplitude, wavelength and phase of the excitation current, the gap-field distribution, and the magnetic characteristics of the tape. The complexity of the function is further increased by the nonlinearity of the tape characteristics and its value therefore was determined on a purely experimental basis.

The general influence of recording demagnetization on the performance characteristics follows from the above analysis plus an extended visualization of the magnetic cycle as certain factors vary. As the signal wavelength decreases, the remanent flux is reduced, and the tape output on reproduction is thereby reduced. This is in accordance with the wavelength response characteristic (Fig. 1). As the amplitude of the supersonic excitation increases, the effective air-gap fringing field extends over a greater distance. For a signal of a given wavelength then, the ratio of the signal wavelength to the effective gap length will decrease and the recording demagnetization becomes greater. This correlates with the output versus excitation amplitude curves of Fig. 2.

The final remanence of a tape element at the conclusion of the recording process is not necessarily a value such as *Or* in Fig. 3(c) or 4(c), but rather this value *Or* less the effect of self-demagnetization, an effect which has been discussed by Camras,[8] Wetzel,[9] and others. A correlation of the self-demagnetization effect with the recording demagnetization effect is illustrated in Fig. 5, in which the remanence *Or* of Fig. 3(c) or (4c) is represented by OB_3. The reduction in magnetic induction due to the self-demagnetization effect for a signal having

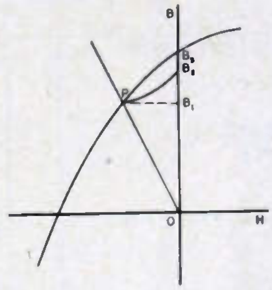

Fig. 5—A section of the magnetization curve illustrating the effect of self-demagnetization.

a wavelength to which the shearing line *OP* applies is the difference between OB_3 and OB_1. In the reproduction process, however, a recovery of induction is made along PB_2 to give an over-all loss due to self-demagnetization represented by the difference between OB_3 and OB_2. For long wavelength signals, e.g., those greater than 20 mils, the shearing line *OP* approaches the vertical so that the self-demagnetization loss is negligible and the value OB_3 (or *Or*) represents the final remanence.

Experimental Procedure

The experimental validation of the proposed theory leading to the concept of recording demagnetization required, in very brief terms, a determination of the fringing air-gap field of the recording head, a method of conducting a small section of the tape through the magnetic sequence corresponding to a given set of recording conditions, and finally the measurement of the resulting remanence of the tape section and its correlation with the actual recording system performance characteristics.

The gap-field measurements in this work were made on a ring-type head having an average gap length of 0.83 mil (0.00083 inch) as measured under a 100× microscope. This small gap length imposed severe limitations on the dimensions and positioning of the exploring coil to be used for the measurements. A final unit using 0.3-mil tungsten wire constructed as shown in Fig. 6 proved satisfactory. Fine adjustment of alignment of the 0.3-mil wire with respect to the air gap was provided by the mounting assembly. Horizontal motion of the coil across the gap and vertical motion above it at any horizontal position were made by cali-

[8] Marvin Camras, "Theoretical response from a magnetic-wire record," Proc. I.R.E., vol. 34, pp. 597–602; August, 1946.
[9] W. W. Wetzel, "Review of the present status of magnetic recording theory, Part I," *Audio Eng.*, vol. 31, pp. 14–17; November, 1947.

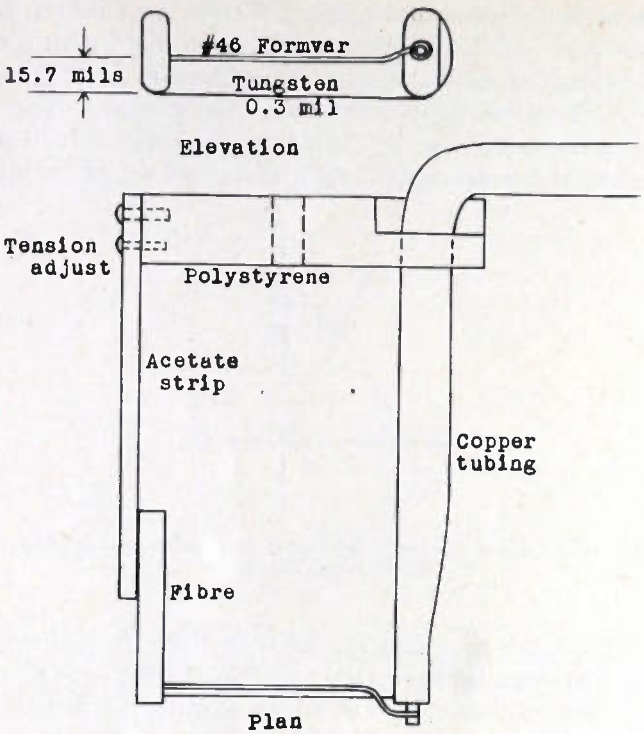

Fig. 6—Sketch of exploring coil used to measure the gap-field distribution.

brated screw adjustments. The sensitivity of these adjustments was such that horizontal displacements down to 0.05 mil and vertical displacements to 0.1 mil could be made and read on calibrated dials.

The exploring coil was connected to an amplifier the output of which supplied a wave analyzer unit in parallel with a vacuum-tube voltmeter and a cathode-ray oscilloscope, as shown schematically in Fig. 7. The

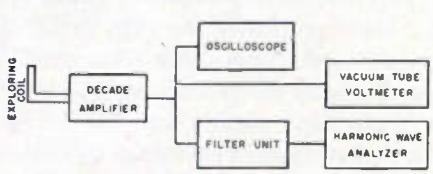

Fig. 7—Block diagram of the equipment used in measuring the air-gap field.

wave analyzer unit consisted of a cathode follower, a low-pass m-derived filter, and an amplifier stage feeding a Hewlett Packard Model 300A Wave Analyzer. The cathode follower was used to match impedance and the filter to attenuate the supersonic voltage when measurements were made with both the supersonic excitation and signal applied to the recording head.

The measurements made with the exploring coil were voltage measurements representing the voltage induced in the coil by the time rate of change of the magnetic flux linkages. Since the coil dimensions were large compared to the dimensions of the region for which the field distribution was to be measured, the change in the voltmeter reading with change in coil position rather than the actual voltage was of significance. If the coil were moved vertically the change in reading would be due to the change in the horizontal component of flux linkages, while if the coil were moved horizontally the change in voltage would be due to the change in the

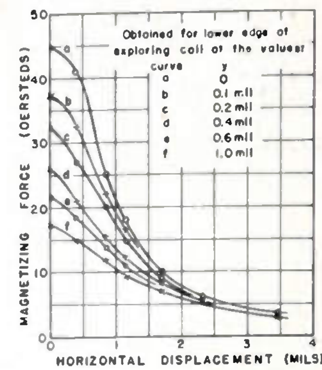

Fig. 8—The horizontal component of magnetizing force as a function of distance from the center of the recording head air gap, as measured with the lower edge of the exploring coil at different distances above the gap.

vertical component of flux linkages. Since the No. 46 Formvar return wire was 15.7 mils above the 0.3-mil wire (Fig. 6), the change in flux linkages due to the No. 46 wire was considered negligible for small movements of the coil, and the entire flux linkage change was assumed to be due to the movement of the 0.3-mil wire. The results of these measurements are shown in final form in Fig. 8.

With a knowledge of the variation of the horizontal component of the magnetizing force across the gap it was possible to combine the signal, supersonic excitation, and field distribution effects into a single function which represented the magnetic cycle of an element of tape as it passed across the gap. Examples of this are shown in Figs. 9 and 10 for signal wavelengths of 8 mils

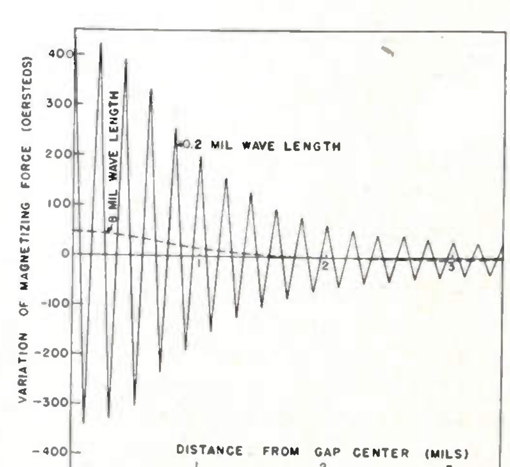

Fig. 9—The variations of the magnetizing force over one-half the air gap during recording with supersonic excitation and a long wave length signal.

and 1 mil, respectively, and for a signal current of 0.18 milliamperes rms, and an excitation current of 1.57 milliamperes rms. Throughout the work an excitation having a wavelength of 0.2 mil (corresponding to 40 kilocycles per second at a tape speed of 8 inches per second) was used. Since symmetry about the center line of the gap is assumed, the complete magnetic history of a tape element is illustrated by considering it as entering the gap region from right to left in Figs. 9 and 10, and leaving from left to right. Initially it was

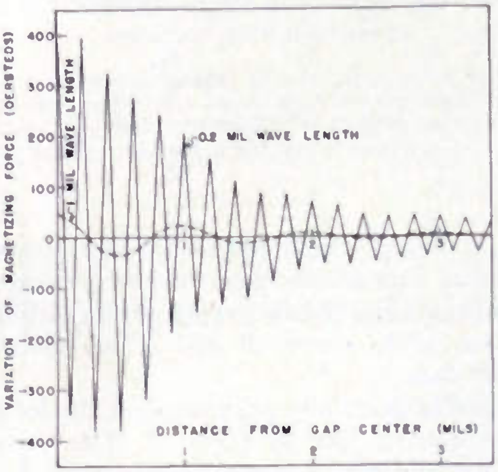

Fig. 10—The variation of the magnetizing force across one-half the air gap for the same magnitudes and phasing as in Fig. 9, but for a short wavelength signal.

thought that a maximum signal would be recorded when the signal and excitation reached a maximum simultaneously at the gap center, for under these conditions a maximum magnetizing force would be established. Accordingly, the curves of Figs. 9 and 10 were drawn for such a phase relationship.

With the magnetic cycle known, the problem then resolved itself into devising a method of magnetizing an element of tape with such a cycle and measuring its remanence. In order to do this, the tape sample was held in a fixed position with respect to the gap instead of moving as in actual recording. The current through the head was then varied smoothly between maxima and minima values similar to those on Figs. 9 and 10, which gave a field variation at the center of the gap corresponding to a given recording condition. In other words, the spatial variation of the field, such as is shown in Figs. 9 and 10, was transformed into a time variation of the field for the element of the nonmoving tape at the center of the gap. For each recording condition it was necessary, therefore, first to tabulate from the gap field distribution data the values of current that would produce the proper successive positive and negative peak values of magnetizing force.

Experimentally the sequence of current values corresponding to the successive positive and negative peak values of magnetizing force were obtained manually by means of the potentiometer control of the bridge circuit shown in Fig. 11. The remanence of the tape element directly over the gap center after it had been subjected to a magnetic cycle corresponding to a given recording condition would be a value such as *Or*, as shown in Figs. 3(c) and 4(c), and therefore would in-

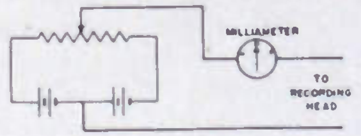

Fig. 11.—Manual control circuit by means of which a tape element was carried through a predetermined magnetic cycle corresponding to a given recording condition.

clude the recording demagnetization but not the self-demagnetization. Measurement of this remanence was made by a remanent pulse playback system which consisted of moving the tape across a play-back head, the output of which was amplified, integrated, and applied to a cathode-ray oscilloscope as shown schematically in Fig. 12. Because of the integration, the peak oscilloscope deflection was proportional to the remanence. Although most of the correlation was concerned with a qualitative verification of characteristic curve shapes, some quantitative measurements were made and therefore a calibration of the remanence measuring system was made.

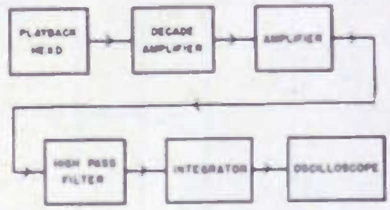

Fig. 12—Block diagram showing the method by which the remanent magnetic flux of a tape element was measured.

Correlation of Experimental Measurements with Theoretical Analysis

A threefold correlation of experimental measurements based on the proposed theory of the recording process and the measured results of normal recording was made. These included a quantitative correlation of the remanent flux for a long wavelength signal, a correlation of the wavelength response characteristic including the gap-length effect, and a correlation of the signal output versus supersonic excitation characteristic.

The quantitative correlation was made for a signal wavelength of 20 mils (400 cps at 8 inches per second) and an excitation wavelength of 0.2 mil. The signal recording current was 0.18 milliamperes and the excitation current 1.57 milliamperes. For these conditions the sequence of peak recording head current values were determined, and a tape element was carried through this sequence with the aid of the circuit of Fig. 11. The tape was then played back through the calibrated remanent pulse playback system shown in Fig. 12 to give a measured remanent flux of 0.041 maxwells. The measured output of a normal recording for equivalent conditions gave a peak remanent flux of 0.047 maxwells. The discrepancy between these is less than 12 per cent, the direct remanent measurement flux being lower than that of the actual recording. This represents approximately 0.9 decibel in the output, a difference which is well within the experimental precision.

In order to establish the correlation with the wavelength response characteristic, elements of the tape were magnetized along a magnetic sequence established for a recording current of 0.18 milliamperes. Initial measurements were made with a phasing such that both the signal and excitation had a positive maximum value at the center line of the air gap. Figs. 9 and 10 show such a cyclic variation. The results were somewhat surprising in that the 2-mil wavelength signal gave an output of virtually zero and shorter wavelengths down to one mil produced output pulses of reversed polarity. This indicated that for the shorter wavelengths the

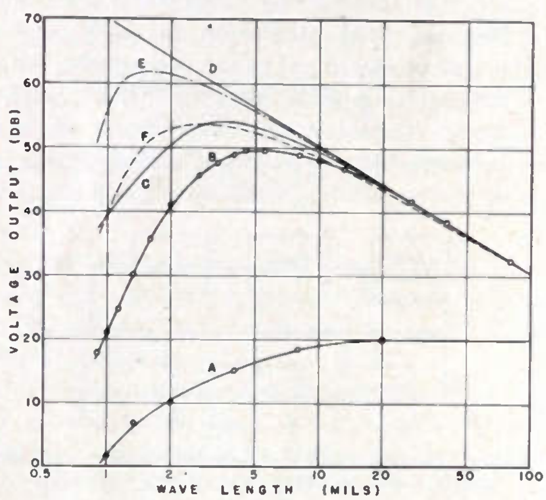

Fig. 13—Wavelength response characteristics. Curve A shows the measured effect of the recording demagnetization as a function of the signal wavelength.

negative part of the cycle was sufficient to counteract the positive magnetization and leave a reversed remanence. It was necessary, therefore, to rephase for such wavelengths and determine the magnetic cycle for the new phase. It was interesting that for a certain phasing the pulse became saddle-shaped, indicating it was only at the center of the gap that the reversed field was sufficient to produce appreciable demagnetization.

The results of the output measurements for a phase that would give maximum output are shown as curve A, Fig. 13, in which the 20-mil wavelength is taken as a reference and plotted arbitrarily as 20 decibels. For wavelengths greater than 20 mils, the signal magnetizing force is attenuated along the gap-field distribution line and therefore would suffer the same recording demagnetization as would the 20-mil wavelength signal. The droop in the curve with reduced wavelength represents the loss in decibels due to increased recording demagnetization above the 20-mil wavelength value.

Curve B, Fig. 13, is the wavelength response characteristic for normal recording under the same conditions. To this has been added the measured recording demagnetization losses (curve A) to give curve C which would represent the wavelength response characteristic including the effect of recording demagnetization. Curve D represents the idealized response in which there are no losses and which is closely approximated for the longer wavelength signals. The playback losses due to the gap length of the playback head neglecting the effect of the thickness of the magnetic coating on the tape were calculated from the expression[2,10]

$$\text{loss} = 20 \log \frac{\sin \dfrac{\pi l_g}{\lambda}}{\dfrac{\pi l_g}{\lambda}} \text{ db,} \tag{1}$$

where l_g is the gap-length and λ is the signal wavelength. These losses were subtracted from curve D to give curve E. In other words, the two upper curves of Fig. 13 are calculated, while curves B and C are plotted from measured data.

The author gratefully acknowledges the information communicated to him by D. G. C. Hare[11] concerning the effect of the thickness of the magnetic coating of the tape on the frequency response of the playback system. The factor to account for this effect was given as $\epsilon^{-2\pi d/\lambda}$ where d is the distance from the surface of the playback head to the center of the magnetic coating, and λ is the recorded signal wavelength. The playback loss function including the effect of the coating thickness then becomes

$$\text{playback loss} = 20 \log (\epsilon^{-2\pi d/\lambda}) \left| \frac{\sin \dfrac{\pi l_g}{\lambda}}{\dfrac{\pi l_g}{\lambda}} \right|_1 \text{ db.} \tag{2}$$

When the playback losses for the tape used are calculated from (2) and subtracted from curve D of Fig. 13, there results curve F which correlates well with the measured output of the actual recording, to which has been added the effect of recording demagnetization (curve C). The remaining discrepancy is attributed to the redistribution of the tape field in the presence of the play-back head, nonparallelism of the gap pole faces, and self-demagnetization effects.

The reduction of output with increased excitation (Fig. 2) was predicated on the basis of increased recording demagnetization with increased excitation for the shorter wavelengths. The experimental validation of

[10] Otto Konei, "Frequency response of magnetic recording," *Electronics*, vol. 20, pp. 124–128; August, 1947.
[11] The DGC Hare Company, New Canaan, Conn.

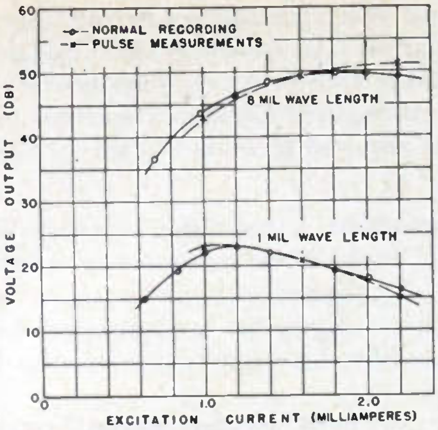

Fig. 14.—Recorded voltage output as a function of the excitation current as measured from a normal recording and from tape elements which had been magnetized by a magnetic cycle corresponding to the conditions of the normal recording.

this is shown in Fig. 14 for signal wavelengths of one and eight mils. The continuous lines pertain to results of normal recording, and the broken lines to the output pulse measurements of tape elements which had been magnetized by the appropriate magnetic cycle and for a phase giving maximum output. It was discovered that the phasing for maximum output was a function of the excitation current and had to be determined for each excitation. Since the self-demagnetization, gap-length effects, and the like, were not measured, the significance is in the shape of the curves and not in the relative outputs. Consequently, for comparison the curves were superposed for an excitation of 1.6 milliamperes. The correlation of the curves is within 2.5 db over the measured range, and the fact that the eight-mil curves rise and the one-mil curves fall effectively substantiates the proposed theory of recording demagnetizat on.

Standards on Transducers: Definitions of Terms, 1951*

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PART I

In this set of definitions, the term "wave" is a generic term intended to cover such concepts or ideas as signal, power, energy, response, stimulus, current, voltage, motion, pressure, and the like, whether these are constant or varying. The waves in the input or output may be of the same or different types (e.g., electrical, mechanical, acoustical, or other).

Active Transducer. A transducer whose output waves are dependent upon sources of power, apart from that supplied by any of the actuating waves, which power is controlled by one or more of these waves.

Bilateral Transducer. A transducer capable of transmission simultaneously in both directions between at least two terminations.

Conversion Transducer (Frequency Conversion Transducer). A transducer in which the input and useful output frequencies are different.

Converter. *See* **Heterodyne Conversion Transducer.**

Dissymmetrical Transducer (with respect to specified terminations). A transducer in which the interchange of at least one pair of specified terminations will change the transmission.

Electric Transducer. A transducer in which all of the waves concerned are electric.

Electroacoustic Transducer. A transducer for receiving waves from an electric system and delivering waves to an acoustic system, or vice versa.

Electromechanical Transducer. A transducer for receiving waves from an electric system and delivering waves to a mechanical system, or vice versa.

Frequency Conversion Transducer. *See* **Conversion Transducer.**

Frequency Divider. *See* **Harmonic Conversion Transducer.**

Frequency Multiplier. *See* **Harmonic Conversion Transducer.**

Harmonic Conversion Transducer. A conversion transducer in which the useful output frequency is a multiple or a submultiple of the input frequency.

Note—Either a Frequency Multiplier or a Frequency Divider is a special case of Harmonic Conversion Transducer.

Heterodyne Conversion Transducer (Converter). A conversion transducer in which the useful output frequency is the sum or difference of the input frequency and an integral multiple of the frequency of another wave.

Ideal Transducer (for connecting a specified source to a specified load). A hypothetical passive transducer which transfers the maximum possible power from the source to the load.

Note—In linear transducers having only one input and one output, and for which the impedance concept applies, this is equivalent to a transducer which (a) dissipates no energy and (b) when connected to the specified source and load presents to each its conjugate impedance.

Linear Transducer. A transducer for which the pertinent measures of all the waves concerned are linearly related.

Note 1—By "linearly related" is meant any relation of linear character whether by linear algebraic equation or by linear differential equation or by other linear connection.

Note 2—The term "waves concerned" connotes actuating waves and related output waves, the relation of which is of primary interest in the problem at hand.

Mode Transformer. *See* **Mode Transducer.**

Mode Transducer (Mode Transformer). A device for transforming an electromagnetic wave from one mode of propagation to another.

Passive Transducer. A transducer whose output waves are independent of any sources of power which is controlled by the actuating waves.

Reciprocal Transducer. A transducer in which the principle of reciprocity is satisfied.

Symmetrical Transducer (with respect to specified terminations). A transducer in which all possible pairs of specified terminations may be interchanged without affecting transmission.

Transducer. A device capable of being actuated by waves from one or more transmission systems or media and of supplying related waves to one or more other transmission systems or media.

Unilateral Transducer. A transducer which cannot be actuated at its outputs by waves in such a manner as to supply related waves to its inputs.

PART II

Related Transmission Terms

The term "loss" used with different modifiers has different meanings, even when applied to one physical quantity such as power. In view of definitions containing the word "loss" (as well as others containing the word "gain") which appear below, the following brief explanation is presented.

(1) Power loss from a circuit, in the sense that it is converted to another form of power not useful for the purpose at hand (e.g., Ri^2 loss) is a physical quantity measured in watts in the mks system and having the dimensions of power. For a given R, it will vary with the current in R.

(2) Loss may be defined as the ratio of two powers: for example, if P_o is the output power and P_i the input power of a network under specified conditions, P_o/P_i is a dimensionless quantity which

would be unity if $P_o = P_i$. Thus, no power loss in the sense of (1) means a "loss," defined as the ratio P_o/P_i, of unity.

(3) Loss may also be defined as the logarithm, or as directly proportional to the logarithm of a power ratio, such as P_o/P_i. Thus if loss $= 10 \log_{10}(P_o/P_i)$ the loss is zero when $P_o = P_i$. This is the standard for measuring loss in decibels.

It should be noted that in cases (2) and (3) the "loss" (for a given linear system) is the same whatever may be the power levels. Thus (2) and (3) give characteristics of the system, and do not depend, (as (1) does) on the value of the current or other dependent quantity.

Available Conversion Power Gain (of a conversion transducer). The ratio of the available output-frequency power from the output terminals of the transducer to the available input-frequency power from the driving generator with terminating conditions specified for all frequencies which may affect the result.

Note 1—This applies to outputs of such magnitude that the conversion transducer is operating in a substantially linear condition.

Note 2—The maximum available conversion power gain of a conversion transducer is obtained when the input termination admittance, at input frequency, is the conjugate of the input-frequency driving-point admittance of the conversion transducer.

Available Power. Of a linear source of electric energy, the quotient of the mean square of the open-circuit terminal voltage of the source divided by four times the resistive component of the impedance of the source.

Available Power Gain (of a linear transducer). The ratio of the available power from the output terminals of the transducer, under specified input termination conditions, to the available power from the driving generator.

Note—The *maximum available power gain* of an electric transducer is obtained when the input termination admittance is the conjugate of the driving-point admittance at the input terminals of the transducer. It is sometimes called "completely matched power gain."

Conversion Transconductance (of a heterodyne conversion transducer). The quotient of the magnitude of the desired output-frequency component of current by the magnitude of the input-frequency component of voltage when the impedance of the output external termination is negligible for all of the frequencies which may affect the result.

Note—Unless otherwise stated, the term refers to the cases in which the input-frequency voltage is of infinitesimal magnitude. All direct electrode voltages and the magnitude of the local-oscillator voltage must be specified, fixed values.

Insertion Gain. Resulting from the insertion of a transducer in a transmission system, the ratio of the power delivered to that part of the system following the transducer to the power delivered to that same part before insertion.

Note 1—If the input and/or output power consist of more than one component, such as multifrequency signal or noise, then the particular components used and their weighting must be specified.

Note 2—This gain is usually expressed in decibels.

Insertion Loss. Resulting from the insertion of a transducer in a transmission system, the ratio of the power delivered to that part of the system following the transducer, before insertion of the transducer, to the power delivered to that same part of the system after insertion of the transducer.

Note 1—If the input and/or output power consist of more than one component, such as multifrequency signal or noise, then the particular components used and their weighting must be specified.

Note 2—This loss is usually expressed in decibels.

Transducer Gain. The ratio of the power that the transducer delivers to the specified load under specified operating conditions to the available power of the specified source.

Note 1—If the input and/or output power consist of more than one component, such as multifrequency signal or noise, then the particular components used and their weighting must be specified.

Note 2—This gain is usually expressed in decibels.

Transducer Loss. The ratio of the available power of the specified source to the power that the transducer delivers to the specified load under specified operating conditions.

Note 1—If the input and/or output power consist of more than one component, such as multifrequency signal or noise, then the particular components used and their weighting must be specified.

Note 2—This loss is usually expressed in decibels.

Voltage Amplification. The ratio of the magnitude of the voltage across a specified load impedance connected to a transducer to the magnitude of the voltage across the input of the transducer.

Note 1—If the input and/or output power consist of more than one component, such as multifrequency signal or noise, then the particular components used and their weighting must be specified.

Note 2—By an extension of the term "decibel," this amplification is often expressed in decibels by multipling its common logarithm by 20.

Voltage Attenuation. The ratio of the magnitude of the voltage across the input of the transducer to the magnitude of the voltage delivered to a specified load impedance connected to the transducer.

Note 1—If the input and/or output power consist of more than one component, such as multifrequency signal or noise, then the particular components used and their weighting must be specified.

Note 2—By an extension of the term "decibel," this attenuation is often expressed in decibels by multiplying its common logarithm by 20.

Charge Storage in Cathode-Ray Tubes*

C. V. PARKER†, SENIOR MEMBER, IRE

Summary—**The charging process in cathode-ray tubes used for static storage of information is analyzed for both a stationary spot and a linear scan, with and without redistribution of secondary electrons. Approximate equations are derived for the surface potentials and charging current as functions of time and other parameters, such as primary beam current, writing speed, and initial potentials. The results are presented graphically in special cases for comparison with photographs of experimental wave forms.**

Introduction

THE PHENOMENON of static storage of electric charge in cathode ray tubes has recently found wide-spread application and is therefore worthy of intensive theoretical and experimental study. The purpose of the present paper[1] is to attempt an approximate analysis of the charging action and to deduce the effects of various parameters of the process on the output current. The analysis is undertaken in four well-defined parts which arise naturally from the increasing complexity of the phenomena. Parts I and II apply approximately to cathode-ray tubes containing grids closely spaced to the insulating surface[2–5] while parts III and IV apply to ordinary cathode-ray tubes used for storage purposes.[6–8] An analysis[9] of a barrier-grid tube with "backing-plate" modulation was published recently with results similar to some described in Part II of this paper, but differing in detail because of a different approach.

The generalized storage tube consists of an evacuated envelope containing at least the following parts: an electron gun, deflecting plates, a collector electrode, and a target electrode made up of an insulating material backed up by a conducting plate. The collector electrode is normally at the highest potential of any electrode in the tube (usually ground potential). The output terminal is the conducting plate of the target electrode which is connected to the collector through a suitable output resistor.

Part I—Spot Charging Without Redistribution

It is by this time well known that current flows in the output resistor when the primary beam is first turned on due to the charging current to the capacitance formed by the spot under the electron beam with the back plate of the target electrode. A secondary current, consisting of δ secondary electrons (on the average) for each primary electron, is emitted from the spot, which will temporarily be assumed to be initially at collector potential. These secondaries will be carried by their initial kinetic energy to the collector. If δ is greater than unity, the spot will charge positively due to the loss in electrons. Those secondaries with the lowest kinetic energy of emission are pulled back to the spot by its positive charge; the apparent secondary emission ratio is reduced. Higher and higher energy secondaries are pulled back as the spot charges until eventually a condition of equilibrium is reached in which the apparent secondary emission ratio is unity.

Referring to Fig. 1, the number N of secondary electrons of initial energy between E and $E+dE$ electron-volts emitted per second is assumed to vary with E according to a function $f(E)$. The total area under the

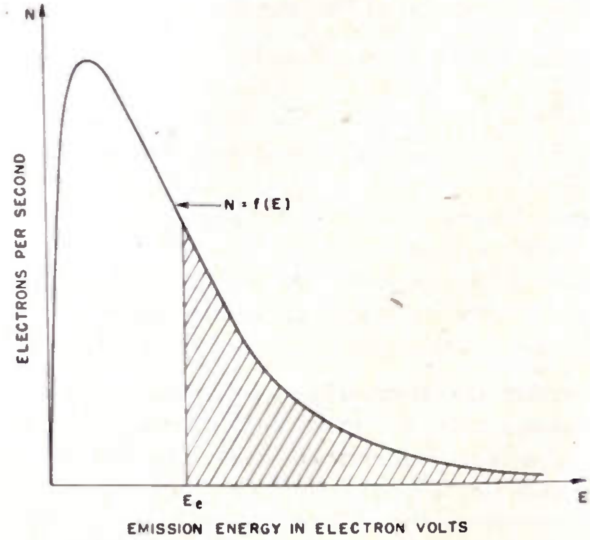

Fig. 1—Energy distribution curve for secondary electrons.

curve represents the number of electrons emitted per second δI_0, where I_0 is the beam current. The area not cross-hatched represents the number of secondary electrons which return to the spot per second after the equilibrium potential has been reached; the cross-hatched area represents the current I_0 which flows to the collector.

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[1] Condensed from Naval Research Laboratory Report 3648, dated March 17, 1950.

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[3] A. S. Jensen, J. P. Smith, M. H. Mesner, and L. E. Flory, "Barrier grid storage tube and its operation," *RCA Rev.*, vol. 9, pp. 112–135; March, 1948.

[4] R. C. Hergenrother, and B. C. Gardner, "The recording storage tube," Proc. I.R.E., vol. 38, p. 740; July, 1950.

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[7] F. C. Williams, and T. Kilburn, "A storage system for use with binary-digital computing machines," *Jour. IEE*, Part IIIA, vol. 96, p. 81; March, 1949.

[8] J. P. Eckert, Jr., H. Lukoff, and G. Smoliar, "A dynamically regenerated electrostatic memory system," Proc. I.R.E., vol. 38, p. 498; May, 1950.

[9] J. V. Harrington, "Storage of small signals on a dielectric surface," *Jour. Appl. Phys.*, vol. 21, p. 1048; October, 1950.

In order for a secondary electron to escape from the spot and reach the collector, it must possess at least as much kinetic energy initially as its potential energy in the field of the spot just before reaching the collector. If the penetration of the fields of auxiliary electrodes into the region of interest is negligible, this potential energy in electron volts is simply the potential of the spot. Thus, in this case of no redistribution, the abscissa E of Fig. 1 may be identified as the potential of the spot V. Hence, the fractional part of the secondary current δI_0 which returns to the spot may be defined as $S\delta I_0$ from which S is defined as

$$S = \int_0^{E=V} f(E)dE \Big/ \int_0^{\infty} f(E)dE. \qquad (1)$$

In the case of thermionic emission, the energy distribution of escaping electrons is Maxwellian. A plot of this distribution[10] curve has the same general shape[11] which has been found experimentally for secondary electrons from metals as in Fig. 1. Hence, we obtain

$$S = \int_0^{E=V} 2(E/E_T)^{1/2} e^{-E/E_T} d(E/E_T)^{1/2} = 1 - e^{-V/E_T} \qquad (2)$$

where E_T is the average energy due to Z-directed motion (the Z axis is taken perpendicular to the surface from which the electrons are emitted).

The output current consists of three components: The primary beam current $(-I_0)$, the secondary emission current (δI_0), and the secondaries which return to the surface after emission $(-S\delta I_0)$, as given by the following formula:

$$I = \delta I_0 - I_0 - S\delta I_0 = C\frac{dV}{dt} \qquad (3)$$

where C is the capacitance between the spot and the conducting back plate, and dV/dt is the time rate of change of the effective potential of the spot. Using the value of S from equation (2), there is obtained:

$$\frac{dV}{dt} - \frac{\delta I_0}{C} e^{-V/E_T} + \frac{I_0}{C} = 0. \qquad (4)$$

If the initial potential of the spot is V_i, and $\gamma = I_0/CE_T$, the solution is

$$V = E_T \ln\left[\delta + (e^{V_i/E_T} - \delta)e^{-\gamma t}\right]. \qquad (5)$$

Note that the potential approaches $V_e = E_T \ln\delta$. The output current is

$$\frac{I}{I_0} = \frac{C}{I_0}\frac{dV}{dt} = \delta e^{-V/E_T} - 1 = \frac{(\delta - e^{V_i/E_T})e^{-\gamma t}}{\delta - (\delta - e^{V_i/E_T})e^{-\gamma t}}, \qquad (6)$$

which may be approximated to a fair degree of accuracy by a linear function of potential and a simple exponential function of time. These results would be obtained if the exponential in equation (2) were expanded in a power series and only the first two terms retained; i.e., $e^{-x} \doteq 1 - x$. In order to obtain better agreement between the more exact expressions and their simpler approximations over a wide range of the independent variables, a constant factor m may be introduced so that $S = mV/E_T$. This approximation is useful in more complicated cases.

The physical situation makes it clear that the maximum output current obtainable is $(\delta - 1)I_0$ and the minimum current is $-I_0$ so that $0 \leqq S \leqq 1$. If the potential V is made negative, either by positive bias on the collector or by the deposition of negative charge on the dielectric surface, the derived equations do not apply since the output current can be no greater than $(\delta - 1)I_0$. However, if V is made more positive than the equilibrium potential, equation (6) correctly predicts the limiting current $-I_0$. In either case the charging process automatically brings the surface to zero potential with respect to the collector after which the derived equations apply.

The assumed condition of no redistribution may be approximated in the practical case by use of a wire mesh closely spaced to the insulating surface compared to the spot size. Only a small fraction of those secondary electrons whose Z-directed energy is insufficient to permit them to reach or pass through the wire mesh but whose radially-directed energy is large will fall outside the boundaries of the spot. The number of such electrons approaches zero with decreasing spacing between wire mesh and surface. Hence, it is only necessary to consider the Z-directed energy. The closer the physical situation approaches one of no redistribution, the closer we may expect our derived equations to apply.

Part II—Scanning Without Redistribution

As we have seen, the output current is a function of the potential of the spot. As long as the spot is stationary, the potential may be assumed the same over the entire spot. When the beam begins to move, the leading edge of the beam strikes a region whose potential is that left by preceding scans or the initial potential of the surface. The trailing edge of the beam strikes a region which has already been under bombardment for at least the time required for the beam to move from its initial position to its present position, and the parts of the beam between the leading and trailing edges strike regions at intermediate potentials. It is necessary, then, to integrate over the area of the spot.

The beam (and, hence, the spot) will be assumed rectangular in cross section of width a in the direction of scan and height b perpendicular to the direction of scan, and the writing speed $W = dx/dt$ will be assumed constant. The assumption will also be made that the previously derived equation for the potential may be applied to each incremental area bdx of the spot with the functional dependence of S upon the potential of the incremental area the same as in the case of the station-

[10] W. G. Dow, "Fundamentals of Engineering Electronics," John Wiley and Sons, Inc., New York, N. Y., p. 237; 1937.
[11] H. Salow, "Die Sekundarelektronenemission," *Fern. Tech. Zeit.*, Heft 6, p. 161; June, 1949.

ary spot. Defining the capacitance per unit area as C/ab, it follows that

$$dI = \frac{C}{ab}\cdot bdx\cdot\frac{dV}{dt} \tag{7}$$

is the contribution of current from area bdx. The total current is

$$I = \int_0^a \frac{C}{a}\frac{dV}{dt}dx. \tag{8}$$

When the beam is first turned on and the scan begins, a transient current flows, but after the time required for the spot to move its own width, a steady-state condition is reached in which the total current is constant. Since $W = dx/dt$, the integral for the steady-state current may be evaluated by changing the variable of integration from x to t. Thus:

$$I = \frac{C}{a}\int_0^{a/W}\frac{dV}{dt}W\,dt = \frac{WC}{a}\left[V\left(t=\frac{a}{W}\right) - V_i\right]. \tag{9}$$

Using the value of V from (5)

$$\frac{I}{I_0} = \frac{W}{\gamma a}\ln\left[\delta + (e^{V_i/E_T} - \delta)e^{-\gamma a/W}\right] - \frac{WCV_i}{aI_0}. \tag{10}$$

Fig. 2 shows I/I_0 and V/E_T plotted as functions of $\gamma a/W = I_0 a/WCE_T$ for the case of $V_i = 0$, $\delta = 4$. The V/E_T curve also indicates the way in which the output current in the steady state varies with primary beam current, since from (9) I is linear in V.

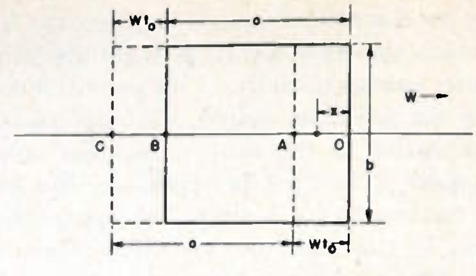

Fig. 3—Initial and present positions of spot for scanning beam transient analysis, no redistribution.

potential is the same for all points between A and B but varies between $x = 0$ and $x = Wt_0$ (point A). Thus, the integration must be carried out in two parts

$$I = \frac{C}{a}\int_{x=0}^{x=Wt_0}\frac{dV}{dt}(x)dx + \frac{C}{a}\frac{dV}{dt}(t_0)\int_{x=Wt_0}^{x=a}dx \tag{11}$$

$$\frac{I}{I_0} = \frac{W}{\gamma a}\ln\left[\delta + (e^{V_i/E_T} - \delta)e^{-\gamma t_0}\right] - \frac{WCV_i}{aI_0} + \frac{(\delta - e^{V_i/E_T})e^{-\gamma t_0}\left(1 - \frac{Wt_0}{a}\right)}{\delta - (\delta - e^{V_i/E_T})e^{-\gamma t_0}}. \tag{12}$$

It will be noted that when $t_0 = a/W$, this expression reduces to that for the steady state, but that for $t_0 < a/W$, the output current contains a component having the form of the stationary spot current multiplied by $(1 - Wt_0/a)$. The form of the resultant current is shown

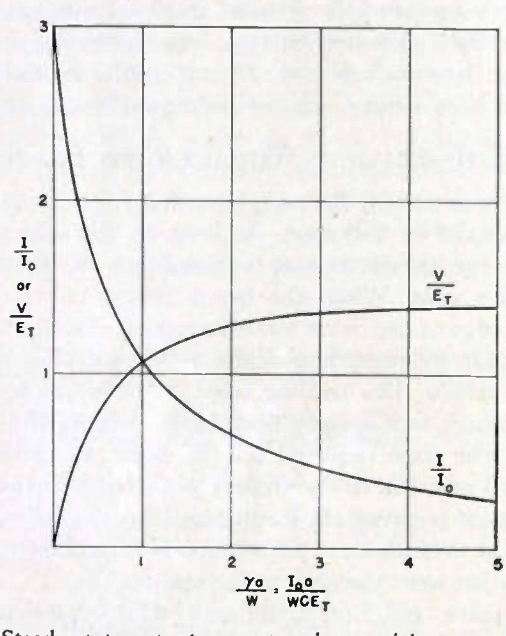

Fig. 2—Steady-state output current and potential versus $\gamma a/W$ for $\delta = 4$, $V_i = o$. Scanning beam with no redistribution.

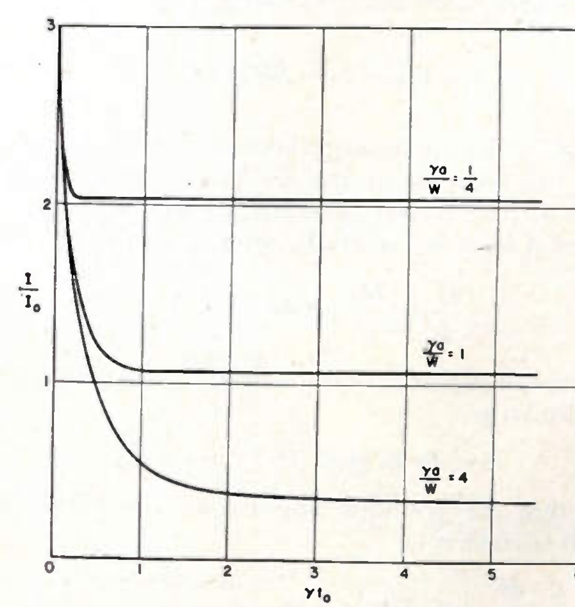

Fig. 4—Transient output current versus γt_0 for $\delta = 4$, $V_i = 0$. Scanning beam with no redistribution.

In the transient state, the integration over x must be carried out with t held fixed, and dV/dt expressed as a function of x. Referring to Fig. 3, the initial and present positions of the spot are indicated by the dotted and solid lines, respectively. The time rate of change of

in Fig. 4 for three different values of $\gamma a/W$ in the special case of $\delta = 4$, $V_i = 0$. The higher the writing speed the greater is the final output current. The initial current is independent of writing speed and is directly proportional to the beam current.

Figs. 5 and 6 are photographs of the experimental output current obtained on an early model of the Haeff memory tube.[2] All parameters except writing speed were held constant for these photographs. The beam current was turned on for 20-microsecond intervals in a sequence of three times on, followed by an off period during which the hold gun was pulsed on to provide erasure. The photographs show the large output on the first scan (after erasure) which charges the surface to equilibrium except at the beginning and end of the on periods. Hence, transient outputs appear at these points at subsequent on times. The off times show in these photographs as mere closures of the base line since the output during erasure is not shown. The similarity of the output current on the first scan after erasure to the curves of Fig. 4 is apparent including the dependence on writing speed.

Part III—Spot Charging With Redistribution

In this case it is necessary to consider the effect of those secondary electrons which return to the region around the spot. Since the secondary emission ratio is close to zero for low energies, most of these electrons "stick" where they land and depress the effective potential of this region. In the absence of these redistributed electrons, the dielectric surface in the vicinity of the positively-charged spot would be at a positive potential with respect to ground which varied inversely with distance from the spot. Thus there is a tendency for the returning secondaries to pile up in the immediate vicinity of the spot and for their number to fall off with distance in the same general way as the potential does. It is quite certain experimentally that most of the returning secondaries are deposited within two or three spot diameters of the center of the spot, although evidence of their presence is found out to many spot diameters. In order for additional secondaries to arrive at any point, the resultant potential of that point must still be positive with respect to the collector.

In order to handle this complex charge distribution even approximately, the idealizing concept of an equivalent constant-potential region of negative charge surrounding the spot is introduced. This region of negative charge will be designated as Region Two (2), with Region One (1) designating the positively-charged spot. The actual potential of any point of the screen may be written as some fraction of the potential of the spot V_1, plus some fraction of the (negative) potential of Region Two V_2, the fractions depending on the distance of the point from the center of the spot and the size of the spot.

Now the Z-directed initial energy required of a secondary electron to permit it to land at a given point depends not only on the location of the point, but also on the initial direction of emission. Those electrons emitted almost normal to the surface may possess considerable initial energy and still return to the spot while those electrons making a large angle with the normal may land in Region Two with much smaller amounts of Z-directed initial energy.

The argument of Part I must now be modified since there are two regions of interest at different potentials charging at different rates. Hence, there are two differential equations:

$$I_1 = \delta I_0 - I_0 - S_1\delta I_0 = C_1 \frac{dV_1}{dt} \tag{13}$$

$$I_2 = - S_2\delta I_0 \qquad = C_2 \frac{dV_2}{dt}. \tag{14}$$

These equations are not independent, however, since the fractional number of secondaries which return to Region One (S_1) depend on the potential of both Regions One and Two, etc. Hence, it is necessary to average over these regions by defining some new parameters dependent on the geometry, chiefly spot size, and the angular distribution and energy distribution of the secondaries characteristic of the emitting material. Then, by an argument similar to that of Part I, and making the approximation that was indicated in Part I

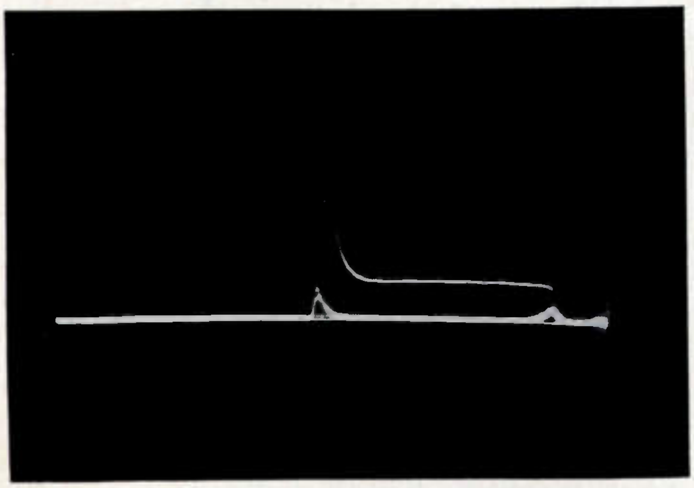

Fig. 5—Output current wave form obtained with early model Haeff Memory Tube. Writing speed, 0.0152 inch per microsecond.

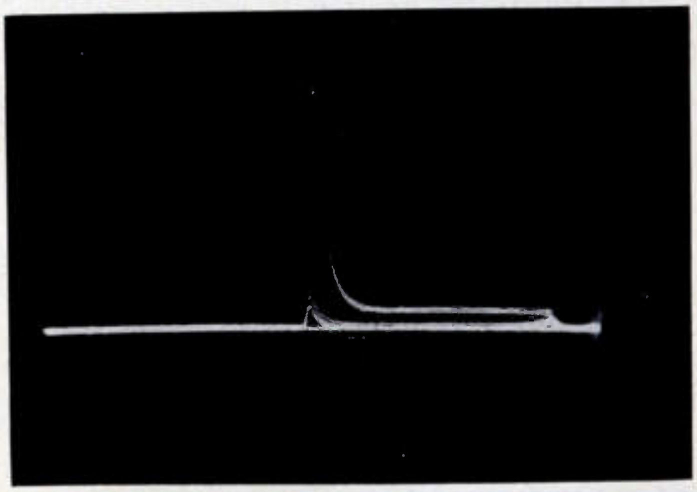

Fig. 6—Same as Fig. 5, except writing speed is 0.00715 inch per microsecond.

as being reasonable, $e^{-V/E_T} \doteq 1-(mV/E_T)$, we have

$$S_1 = m_1 V_1 + m_3 V_2 + S_{01} \tag{15}$$

$$S_2 = m_2 V_1 + m_4 V_2 + S_{02}. \tag{16}$$

Both S_1 and S_2 have been increased by the constant terms S_{01} and S_{02} in order to take account of residual fields which may penetrate in some cases to the region around the spot to cause the return of some secondary electrons to the screen, even when the screen bears no positive charge. The restrictions on S mentioned in Part I apply here also in the form $0 \leqq (S_1+S_2) \leqq 1$, limiting the total output current to $(\delta-1)I_0$ and $-I_0$, respectively. The differential equations now become

$$C_1 \frac{dV_1}{dt} = (\delta - 1)I_0 - \delta I_0(S_{01} + m_1 V_1 + m_3 V_2) \tag{17}$$

$$C_2 \frac{dV_2}{dt} = -\delta I_0(S_{02} + m_2 V_1 + m_4 V_2) \tag{18}$$

with the solutions

$$V_1 = X_1 e^{-\delta I_0 N_1 t} + Y_1 e^{-\delta I_0 N_2 t} + V_{e1} \tag{19}$$

$$V_2 = X_2 e^{-\delta I_0 N_1 t} + Y_2 e^{-\delta I_0 N_2 t} + V_{e2} \tag{20}$$

where N_1 and N_2 are given by the two solutions

$$N = \frac{1}{2}\left[\frac{m_1}{C_1} + \frac{m_4}{C_2} \pm \sqrt{\left(\frac{m_1}{C_1} + \frac{m_4}{C_2}\right)^2 - 4\left(\frac{m_1 m_4 - m_2 m_3}{C_1 C_2}\right)}\right]. \tag{21}$$

We will designate by N_2 the value of N corresponding to the use of the positive sign and by N_1 the negative sign. Hence, N_2 will represent the sum of two positive terms while N_1 represents their difference. The output is the sum of I_1 and I_2 or

$$\frac{I}{\delta I_0} = [H_1(V_{e1} - V_{i1}) + H_2(V_{e2} - V_{i2})]e^{-\delta I_0 N_1 t} + [H_3(V_{e1} - V_{i1}) + H_4(V_{e2} - V_{i2})]e^{-\delta I_0 N_2 t}, \tag{22}$$

where H_1, H_2, H_3, and H_4 are constants involving C_1, C_2, m_1, m_1, m_3, and m_4. From their definitions, N_2 would be expected to be larger than N_1. This is borne out experimentally with the factor N_2/N_1 being large. The output current turns out to be the sum of a small amplitude, long time-constant component; and a large amplitude, short time-constant component. The latter component is of positive polarity experimentally if both initial potentials are near zero and may be interpreted as chiefly representing the net positive current to the spot which, having small capacitance to ground because of its small size, charges rapidly to its equilibrium potential. The former component may be interpreted as the redistributed-electron current which continues to charge the relatively-large capacitance of the region about the spot for a long time. Experimentally, this longer time-constant does not seem to remain constant with time, but approaches a limiting value which is independent of beam current after hundreds of microseconds. Apparently the initial assumption that Region Two remained fixed in area indefinitely is not valid after a sufficient time, as might have been expected, but that, on the contrary, the higher-energy secondaries continue to land on the border so that the boundary slowly spreads out.

The general behavior of the output is experimentally as predicted, however, as shown by Figs. 7 through 9.

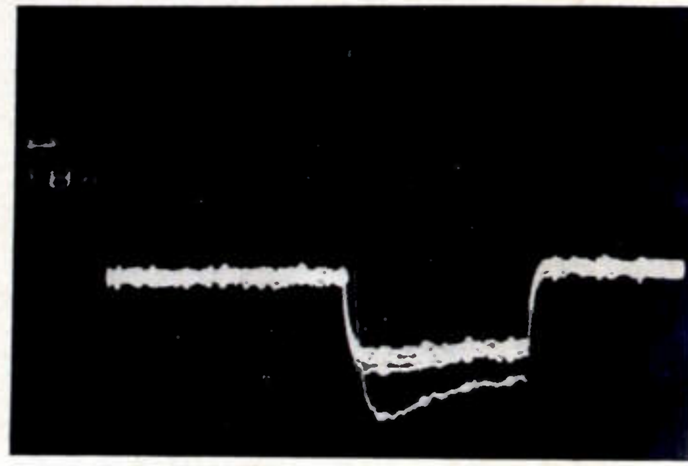

Fig. 7—Photograph of output current from a stationary spot with redistribution. Commercial 5JP5 cathode-ray tube. Beam current, 0.130 microampere.

Fig. 7 is an oscilloscope photograph of the output current versus time, clearly showing the two components referred to above. The experimental technique was to pulse the primary beam on for several intervals instead of for one longer interval, and record all the results on the same photograph. Thus, the output during several hundred microseconds is shown with the positive peak indicating the start of the process. Fig. 8

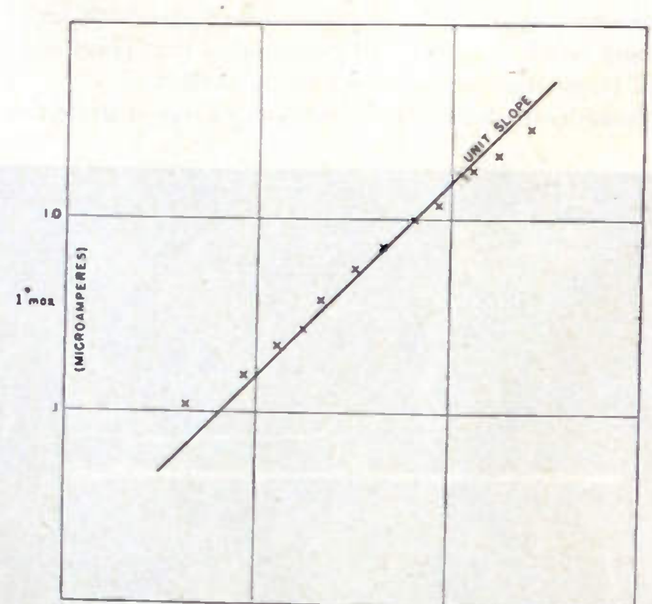

Fig. 8—Positive peak current versus beam current. Stationary spot with redistribution. Crosses show experimental points.

shows the amplitude of positive peaks of such curves as determined experimentally, plotted versus beam current over nearly a 100-to-1 range. Fig. 9 is a plot of the reciprocal of the time constant of the positive component over the same range in beam current. The spot diameter for Figs. 8 and 9 was 0.6 cm representing a

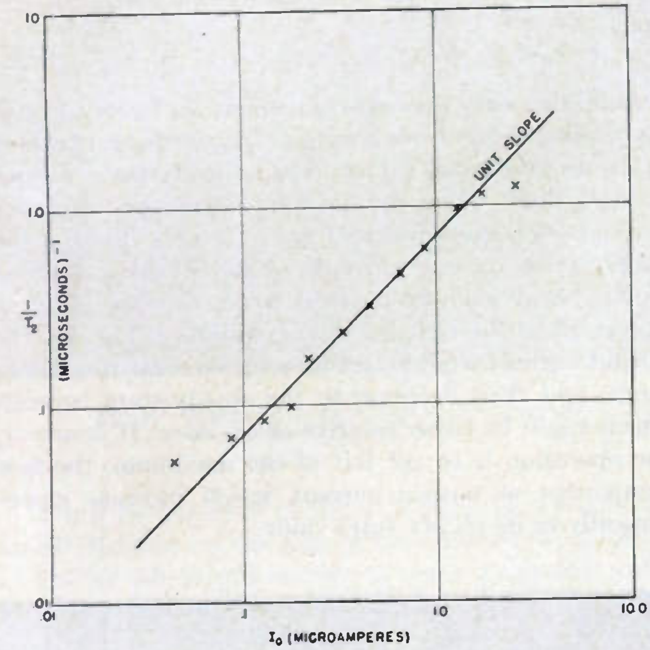

Fig. 9—Reciprocal of time constant of positive component versus beam current. Stationary spot with redistribution. Crosses show experimental points.

purposely defocused beam in order to permit more accurate measurements. Since the N_1 and N_2 factors vary inversely with the capacitance coefficients, increasing C_1 and C_2 by increasing the spot size increases the corresponding time constants.

Part IV. Scanning With Redistribution

The general procedure of Part II may now be repeated using the values for the potentials obtained in Part III, but it must be recognized that this formal process ignores some fundamental physical facts of redistribution. The secondary electrons returning to Region Two fall in part ahead and in part behind the scanning beam, as well as on each side of the line of scan. Thus, when the beam reaches any particular point, it finds an initial potential which is more negative than it was before the beam approached its vicinity and, after the beam has passed by, the point again goes negative in potential since it is again in a Region Two of nearby points. Furthermore, points off the line of scan are bombarded by secondary electrons during the time that the scanning beam moves several times its own width in contrast to the no-redistribution case of Part II in which the steady-state current and potential were reached in the time required for the beam to move just once its own width. Also, the division of secondaries between the different regions of the screen may be affected by the potential of the line of scan, both ahead of and behind the spot.

These complications might be taken into account, at least approximately, but the process involves expressions so unwieldy that a simpler if less accurate analysis will be given here. The final establishment of the limits of validity of the results can only (as in any case) be obtained by resorting to experiment. We proceed then to find the steady state output current by integration over the areas shown in Fig. 10. As in Part II, the current to Region One is

$$I_1 = \int_0^{a_1} \frac{C_1}{a_1}\frac{dV_1}{dt}\,dx = \frac{WC_1}{a_1}\left[V_1\left(t=\frac{a_1}{W}\right) - V_{i1}\right]. \tag{23}$$

As an approximation, the current to Region Two is a_2/a_1 times the current to an area b_2 in height and a_1 in width, or

$$I_2 = \frac{a_2}{a_1}\int_0^{a_1} \frac{C_2}{a_2}\frac{dV_2}{dt}\,dx = \frac{WC_2}{a_1}\left[V_2\left(t=\frac{a_1}{W}\right) - V_{i2}\right]. \tag{24}$$

Introducing the expressions for the potentials from Part III, we have

$$\begin{aligned} I &= I_1 + I_2 \\ &= \frac{W}{a_1}\,[K_1(V_{e1} - V_{i1}) + K_2(V_{e2} - V_{i2})]e^{-\delta I_0 N_1 a_1/W} \\ &\quad - [K_3(V_{e1} - V_{i1}) + K_4(V_{e2} - V_{i2})]e^{-\delta I_0 N_2 a_1/W} \\ &\quad + C_1(V_{e1} - V_{i1}) + C_2(V_{e2} - V_{i2}), \end{aligned} \tag{25}$$

where K_1, K_2, K_3, and K_4 are constants involving C_1, C_2, m_1, m_2, m_3, and m_4. The output current is zero for beam current zero and approaches a constant value

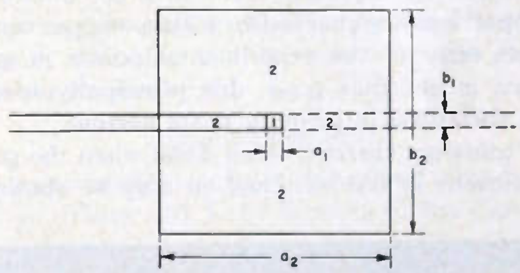

Fig. 10—Instantaneous position of moving spot showing idealized Regions One and Two.

for very high beam current. The first exponential in (25) is relatively close to unity so that the approximation $e^{-x} \doteq 1 - x$ may be used, and, for easy comparison with experiment, (25) written in the abbreviated form

$$I = \alpha\beta(1 - e^{-I_0/\alpha}) - \gamma I_0, \tag{26}$$

in which

$$\begin{aligned} \alpha &= W/\delta a_1 N_2 \\ \beta &= \delta N_2[(K_1+C_1)(V_{e1} - V_{i1}) + (K_2+C_2)(V_{e2} - V_{i2})] \\ \gamma &= \delta N_1[K_1(V_{e1} - V_{i1}) + K_2(V_{e2} - V_{i2})]. \end{aligned} \tag{27}$$

Equation (26) is plotted in Fig. 11 as a function of beam current using values for α, β, and γ which produce a fit with the experimental points. Secondary emission ratio was 3.76 for P-5 phosphor operating at 1,400 volts accelerating potential. The linear dependence of α upon the writing speed has been checked by locating the points at which the output current is zero in the steady

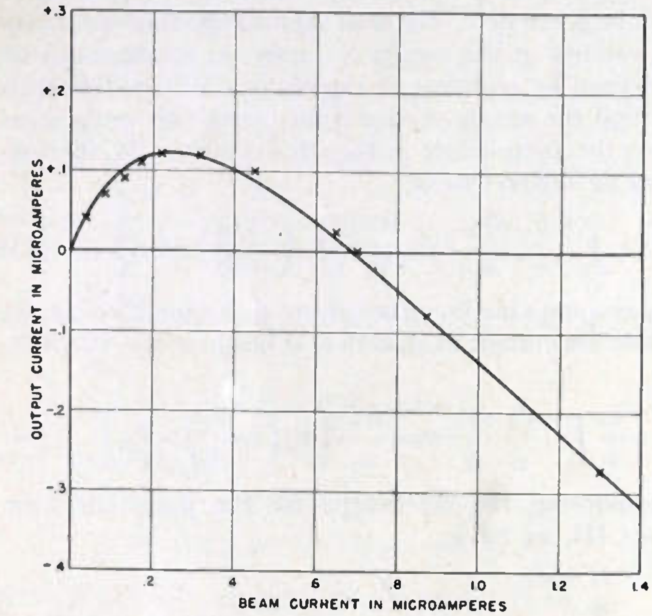

Fig. 11—Steady-state output current versus beam current for scanning beam with redistribution. Writing speed, 0.20 inch per microsecond. Crosses show experimental points. Curve is plot of equation

$$I = 0.325(1 - e^{-I_0/0.190}) - 0.460\, I_0.$$

state. The beam current at these points is directly proportional to the writing speed. For the three writing speeds 0.0247, 0.0817, and 0.200 inch per microsecond, the proportionality checked to within one per cent. The probable error of the experimental points in general, however, is not that good, due principally to uncontrolled variations in primary beam current.

The transient current which flows when the primary beam current is first switched on may be obtained by the method used in Part II with a similar result.

$$I = \frac{W}{a_1}\left[(C_1X_1 + C_2X_2)e^{-\delta I_0 N_1 t} + (C_1Y_1 + C_2Y_2)e^{-\delta I_0 N_2 t} + C_1(V_{e1} - V_{i1}) + C_2(V_{e2} - V_{i2})\right] - \delta I_0\left[1 - \frac{Wt}{a_1}\right]\left[N_1(C_1X_1 + C_2X_2)e^{-\delta I_0 N_1 t} + N_2(C_1Y_1 + C_2Y_2)e^{-\delta I_0 N_2 t}\right]. \tag{28}$$

Considering only the first component, as t goes through its possible values from zero to a_1/W, the output current takes on the same values as the steady-state output current does when the primary beam current goes through its possible values from zero to I_0. Thus, if the steady-state output current is less positive than it would be at a lower beam current, corresponding to operation to the right of the maximum in Fig. 11, the output current would start at zero, increase to a maximum, and then decrease to the steady-state current, which might be either positive or negative. If, however, the operation is to the left of the maximum, the first component of output current would increase monotonically to its steady-state value.

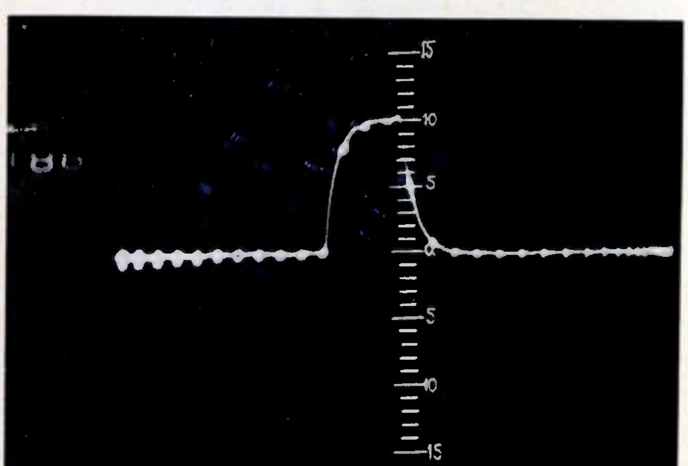

Fig. 12—Photograph of transient output current obtained with commercial 5JP5 cathode-ray tube. Writing speed, 0.150 inch per microsecond. Beam current, 0.130 microampere, 2-microsecond markers.

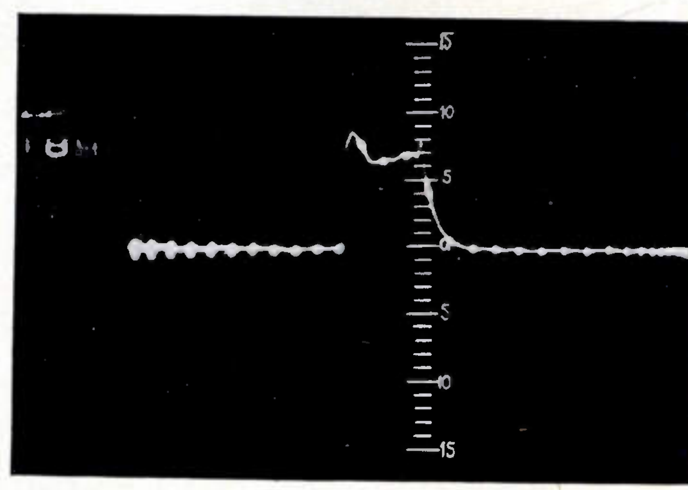

Fig. 13—Same as Fig. 12, except beam current is 0.650 microampere.

Hence, the total output current begins at the peak value of the stationary spot output current and then either increases gradually to its final positive value or else first increases to a maximum and then decreases to either a positive or negative steady-state value depending on the primary beam current. Some examples of these transients as observed experimentally are shown in Figs. 12 through 15.

In practice, the leading edge of the pulsed signal applied to the control grid of the cathode-ray tube approaches fairly closely the action of an ideal switch, but the trailing edge is usually not quite so good. This may give rise to a peculiar effect at the end of the pulse: As the primary beam current decreases towards zero, output current goes through normal variation with I_0, which may involve an increase in output current.

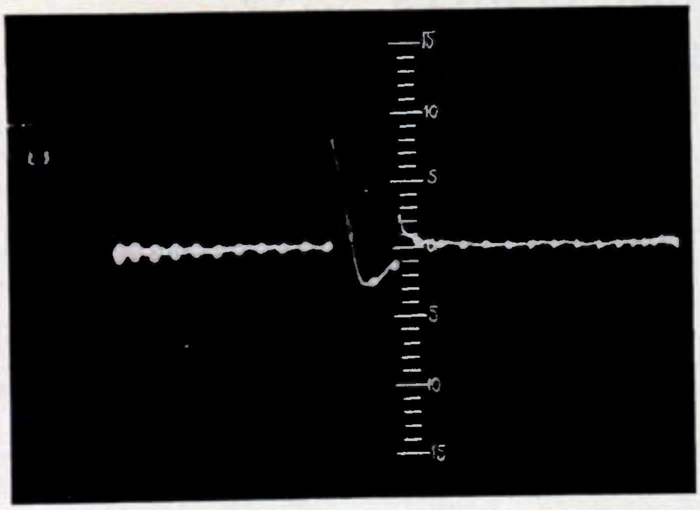

Fig. 14—Same as Fig. 12, except beam current is 0.780 microampere.

It will be observed in Figs. 12 through 15 that the actual distance on the face of the cathode-ray tube required for the output current to reach a constant value is many times a spot diameter. This distance is really a measure of the extent of Region Two along the line of scan which was neglected in this approximate analysis. It appears to be approximately independent of beam current or writing speed in the range so far investigated. No information is yet available on its variation with spot size or other parameters.

Applications

The theory so far developed may be extended as necessary to cover various interesting special cases, although the details will not be given here. Thus the effect of intensity modulation by an ideal square pulse has already been indicated. Modulation by a continuously varying voltage on the control grid, by deflection of the beam, or by variation of the voltage between target and collector may be handled by point-by-point computations if the period of the modulation is long compared to the time required for the beam to move its own width. Otherwise it is necessary to find the solutions of the differential equations for the surface potentials corresponding to the applied function and use these in carrying out the subsequent integration.

Much experimental work remains to be done in checking the validity of the theory presented here. In particular, the effect of variation in initial potentials needs quantitative investigation in view of the complex phenomena which has been neglected in the simplified theory as pointed out at the beginning of Part IV. However, the experimental work so far carried out indicates conformity with the simple theory for the range of parameters investigated.

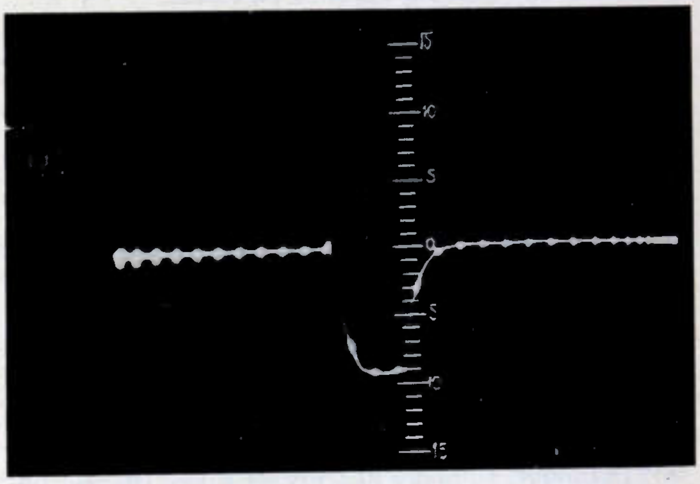

Fig. 15—Same as Fig. 12, except beam current is 1.30 microamperes.

Acknowledgment

The experimental results in this report were obtained by W. A. White and J. E. Scobey, of the Naval Research Laboratory.

CORRECTION

Donald B. Harris, author of the paper, "Product Phase Modulation and Demodulation," which appeared on pages 890–895 of the August, 1950, issue of the Proceedings of the I.R.E., has brought to the attention of the editors the following typographical errors:

1. Equation (1) should read

$$e_i = [\sum F_{an}(t) \cos n\omega t + \sum F_{bn}(t) \sin n\omega t].$$

2. In the line following equation (19), "k" should read "K."

3. Equation (33) should read

$$Q \int_0^{T_Y} I_a dt = I_a T_Y.$$

Gaseous Discharge Super-High-Frequency Noise Sources*

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Summary—The positive column of an electrical discharge through argon it utilized as a source of random fluctuations to provide a standard noise source for microwave frequencies. It is shown that the waveguide may be matched to the discharge with very low reflection over the entire recommended transmission bandwidth of the waveguide without the utilization of tuned elements. Further, when the discharge is not excited, microwave transmission through the guide is possible with very low reflection. The noise temperature is about 15.5 db above 290° K. Performance data are given for frequencies between 2,600 and 26,000 mc per second.

The effect of varying the discharge parameters, geometry, gas, gas pressure and current, upon both the radio-frequency and noise performance is discussed. The factors affecting the occurrence of low-frequency fluctuations in the discharge are considered and the proper operating conditions to avoid their effects on the microwave performance are given.

The hypothesis that the noise temperature corresponds to the electron temperature of the positive column is presented and supporting evidence is offered.

I. Introduction

THE ADVANTAGES of a standard noise source[1] for use in receiver noise measurements have become widely recognized. These advantages result, in part, from the fact that measurements, made using this technique, compare like signals whose energy content is similarly distributed over the frequency band. This makes a detailed knowledge of the frequency pass band of the receiver unnecessary. On the other hand, if a cw (i.e., discrete frequency) signal is employed to measure the integrated noise components, it does become necessary to determine the receiver noise bandwidth, i.e., the region of the frequency spectrum over which the noise components are integrated.

Further advantages result from the fact that noise signals are generated at a low power level—approximately that level best suited for measurement. Thus the standard noise source does not require the extensive shielding and accurately calibrated attenuator common to cw signal generators where the generated power level may be of the order of 100 decibels above that actually required for the noise measurements. These features of the noise source result in simpler measurement techniques and less costly instrumentation.

The temperature-limited diode[2] has been extensively employed as a standard noise source. At frequencies below 300 mc conventional glass envelope tube construction may be used.[3,4] At higher frequencies, the lead inductances and tube capacitances associated with such constructions make the proper circuitry difficult to achieve. The usefulness of the temperature-limited diode has been extended through the ultra-high-frequency region (300–3,000 mc) by constructing the diode as part of the transmission system.[5,6] This method of attack meets with rapidly increasing difficulties[7] as the frequency is further increased into the super-high-frequency region (3,000–30,000 mc).

An electrical discharge through a gas is also known[8,9] to produce microwave noise energy. Mumford[10] has shown that a gaseous discharge, such as that in a commercial fluorescent lamp, is a stable and uniform noise source at microwave frequencies near 4,000 mc, and its available noise power is adequate for receiver measurement purposes. It is the purpose of the present work to extend the use of gaseous discharges for this application to other microwave frequencies, to show how a discharge tube may be matched to a waveguide over the entire recommended transmission bandwidth of the guide, to investigate briefly the properties of discharges in various gases, and to suggest and to provide experimental evidence for the hypothesis that the microwave noise power is a measure of the electron-temperature of the positive column of the discharge.

For this work, the discharge geometry is similar to that of a commercial fluorescent lamp and utilizes an elongated cylindrical envelope in which the positive column occupies 80 or more per cent of the distance between electrodes. However, adjustments are made in the geometry, the kind of gas, the gas pressure, and the discharge current for optimum performance in the present application.

* Decimal classification: R355.913.21. Original manuscript received by the Institute, October 27, 1950; revised manuscript received, February 16, 1951.

† RCA Laboratories Division, Princeton, N. J.

[1] The term "noise source" is used herein to denote a generator of random electrical fluctuations; that is, "fluctuation" noise as distinguished from "impulse" noise. The energy distribution is continuous and uniform throughout the portion of the spectrum of interest.

[2] W. Schottky, "Spontaneous current fluctuations in various conductors," *Ann. Phys.*, vol. 57, p. 541; December, 1918.

[3] J. Moffatt, "A diode noise generator," *Jour. IEE*, (London) pt. IIIA, vol. 93, p. 1335; 1946.

[4] R. W. Slinkman, "Temperature-limited noise diode design," *Sylvania Technologist*, vol. 2, p. 6; October, 1949.

[5] H. Johnson, "A coaxial line diode noise source for U-H-F," *RCA Rev.*, vol. 8, p. 169; March, 1947.

[6] R. Kompfner, et al, "The Transmission line diode as a noise source at centimetre wavelengths," *Jour. IEE* (London) pt. IIIA, vol. 93, p. 1436; 1946.

[7] The authors have constructed experimental models of broadband temperature-limited diodes as sections of ridge waveguide which were operable as calibrated noise sources near 10,000 mc. The constructional difficulties were great and the geometry did not lend itself to a precise calculation of the avilable noise power.

[8] G. C. Southworth, "Microwave radiation from the sun," *Jour. Frank. Inst.*, vol. 239, pp. 285–298; April, 1945.

[9] L. Goldstein and N. L. Cohen, "Radiofrequency conductivity of gas-discharge plasmas in the microwave region," *Phys. Rev.*, vol. 73, p. 83; January, 1948.

[10] W. W. Mumford, "A broad-band microwave noise source," *Bell Sys. Tech. Jour.*, vol. 28, p. 608; October, 1949.

II. The Gaseous Discharge as a Waveguide Termination

A. The Waveguide Mount

It has been found possible to utilize the positive column of a gaseous discharge as a broad-band untuned resistive termination for a waveguide. This has been accomplished in the present work by insertion of the discharge tube diagonally across the waveguide as shown in Fig. 1, for example. This figure shows the discharge tube inserted through the broad faces of the guide and in the plane formed by the vertical and longitudinal axes of the guide. This method of insertion will be referred to as an E-plane insertion. If the acute angle between the tube axis and the longitudinal axis of the guide is not too large, there is but very little reflection from the discharge. Since the attenuation through the discharge is of the order of 20 decibels or greater, an excellent waveguide termination results without the use

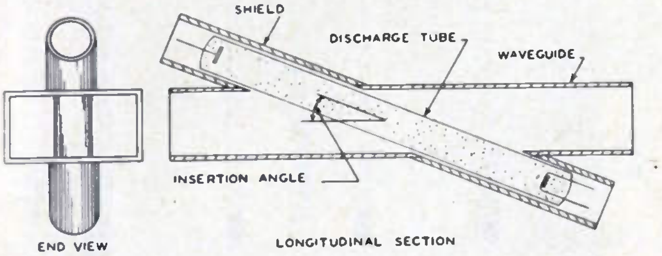

Fig. 1—The E-plane waveguide mount for a discharge tube.

of tuning elements. In particular, with a 10° insertion angle and a discharge through argon, the average voltage standing-wave ratio over the recommended transmission bandwidth of the guide is about 1.07 with a maximum of about 1.1 (this value usually is found near the recommended low frequency limit).

It is also possible to insert the discharge tube through the narrow faces of the guide in the plane formed by the horizontal and longitudinal axes of the guide (H-plane insertion). As a waveguide termination, the performance in this case is generally not quite as good as that described above for the E-plane insertion. It seems probable that similar results would also be obtained for an insertion along a skew diagonal. However, the E-plane insertion is preferred, since the greatest linear dimension of the cut-away portion of the guide wall is parallel to the current flow in the wall, rather than normal to the current flow as in the case of the H-plane insertion. This leads to a minimum reflection when the discharge is turned off and transmission through the waveguide section containing the discharge tube may be accomplished with negligible loss.

The performance of the gaseous discharge as a reflectionless waveguide termination improves as the insertion angle is decreased. This is illustrated by the data of Fig. 2 which show the voltage standing-wave ratios measured over a band of frequencies for insertion angles of 30°, 15°, and 10°. Although these particular data were taken using an H-plane insertion, they serve to illustrate the general statement.

The performance with respect to changes in the diameter of the discharge tube is not particularly critical but our experience indicates that, in order to achieve an efficient design, the ratio of the discharge diameter to

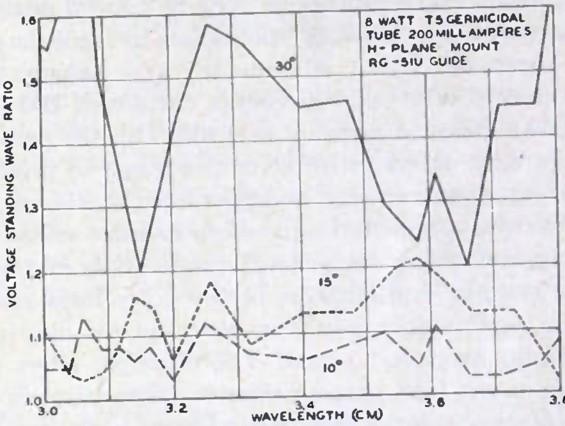

Fig. 2—Illustrative data showing the effect of changing insertion angle.

the inside width of the guide should lie between $\frac{1}{4}$ and $\frac{1}{3}$. This refers to an E-plane insertion and, under these conditions, a 10° insertion possesses a performance adequate for all but the most precise measurements. If the discharge diameter is too small, the tube length must be long (and the insertion angle very small) in order to achieve adequate attenuation. However, if the tube diameter is very large, the reflection due to the cut-away guide wall is increased, and the insertion angle must again be small in order to obtain a gradual transition.

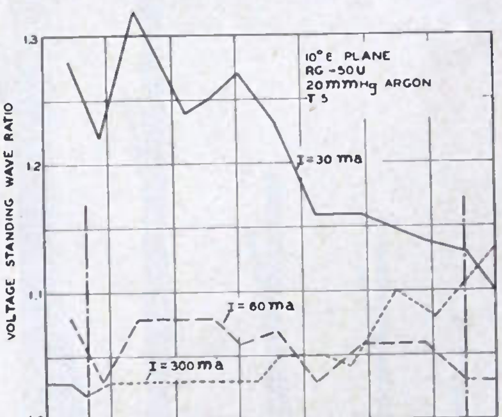

Fig. 3—Typical data showing variation in performance with discharge current

B. Variation of Discharge Parameters

The performance of the above type of termination has been studied to determine the effects of varying the discharge parameters such as the discharge current, the gas pressure, and the kind of gas. In general, the performance is quite insensitive to such variations.

The effect of a variation in discharge current is illustrated by the typical data of Fig. 3 which show the voltage standing-wave ratios measured over a band of frequencies for several values of discharge current. Above a minimum current of about 60 milliamperes in this case, the change in performance is negligible. However, at the lower currents, the attenuation through the discharge may not be entirely adequate.

Similarly, there appears to exist no critical optimum gas pressure. For example, discharges in argon at pressures between 3 and 30 mm Hg give substantially equivalent results (at a discharge current of 200 milliamperes). Below a pressure of 3 mm Hg, the performance rapidly deteriorates. Over the range of pressures tested (up to 30 mm Hg), no upper limit was found.

While the specific data given above have referred to discharges in argon, the general results apply in a qualitative manner to discharges in the other inert gases.[11] Similar performance has been observed for discharges in helium, neon and xenon. The principal effect of a change in the kind of gas seems to be associated with changes in the mean-free-path. Thus, a change to a lighter gas may be compensated for by an increase in gas pressure. In this way, the termination performance figures quoted here for argon may also be readily[12] obtained for other gases.

C. *Typical Results for SHF Band*

The above results have been successfully applied to the construction of untuned waveguide terminations for use as shf noise sources throughout most of the shf band (3,000 to 30,000 mc). A collection of waveguide mounts with a 10° *E*-plane insertion angle is shown in Fig. 4 for various waveguide sizes.

Fig. 4—Collection of waveguide mounts.

Fig. 5 shows a collection of experimental discharge tubes designed to be mounted in the various guide sizes extending from RG-53U ($\frac{1}{4}\times\frac{1}{2}$ inch) to RG-48U ($1\frac{1}{2}\times 3$ inches). These tubes are filled with argon and the pertinent data are given in Table I.

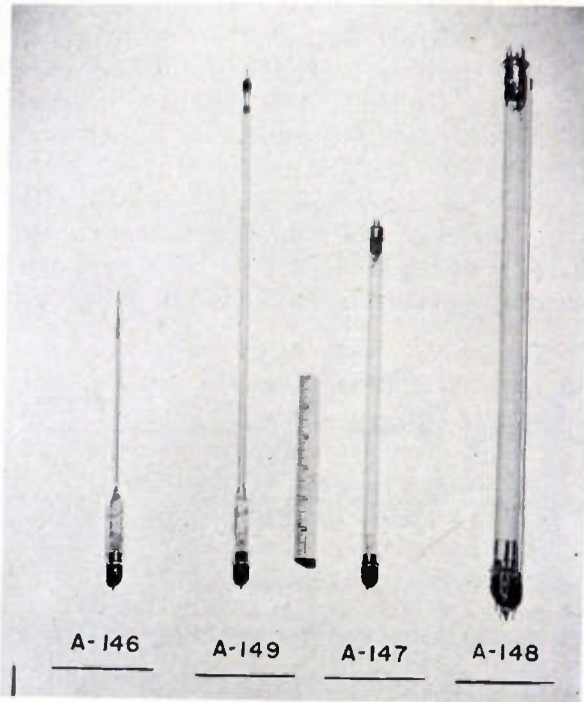

Fig. 5—Collection of experimental discharge tubes.

TABLE I
Experimental Argon Discharge Tube Data

Experimental Designation	*Waveguide Application*	*Nominal Bulb Size**	*Gas Pressure* (mmHg)	*Recommended Operating Current* (ma)	*Power Consumption* (watts)
A 146	RG-53U	T 1½	20	200	15
A 147	RG-50U RG-49U	T 5	20	250	18
A 148	RG-48U	T 8	20	250	22
A 149	RG-52U	T 3	30	200	18

* This designation refers to the nominal diameter (in eighths of an inch) of the envelope along the useful portion of the tube.

Collected voltage standing-wave ratio (VSWR) data showing the performance of these tubes as wave guide terminations is shown in Fig. 6 for various sizes of guide. The tube insertion in each case is a 10° *E*-plane insertion. It is seen that the continuous data extend from 2,600 to 10,000 mc, with an average voltage standing-wave ratio of 1.07. Additional data indicating equal performance is shown between 21,000 and 25,000 mc. The gap in frequency coverage is due to the lack of measuring equipment. The transmission characteristics

[11] These general statements also apply qualitatively to discharges in certain gas mixtures such as the mercury-argon mixture of commercial fluorescent lamps. The performance of certain other mixtures (e.g., hydrogen-neon) is more sensitive to the discharge parameters.

[12] The dc power requirements increase rapidly for the lighter gases. See A. v. Engel and M. Steenbeck, "Electrische Gasentladungen," vol. II, p. 109, J. Springer, Berlin; 1932.

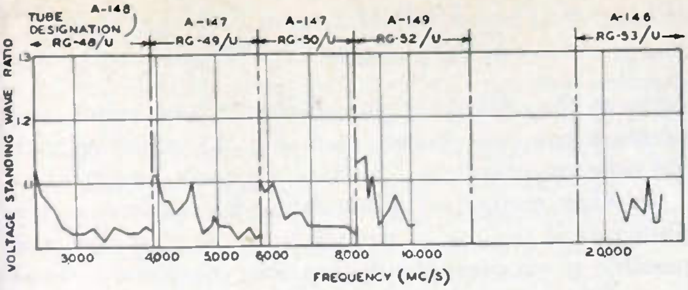

Fig. 6—The performance of argon discharge tubes as waveguide terminations. 10° E-plane insertion.

(the discharge turned off and the guide terminated in a well-matched load) are shown in Fig. 7.

The compactness and simplicity of an assembled unit

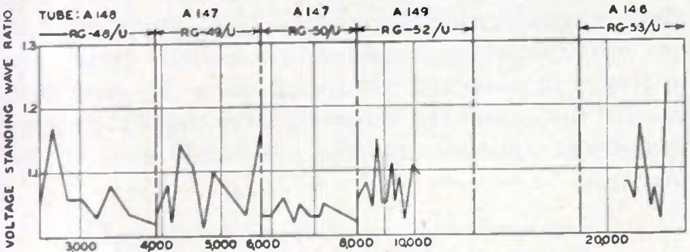

Fig. 7—Transmission characteristics of 10° E-plane waveguide mounts. Waveguide terminated in a matched load beyond mount.

is shown in Fig. 8, which shows a discharge tube (A 147) mounted in an RG-50U guide with the associated power supply mounted beneath the guide.

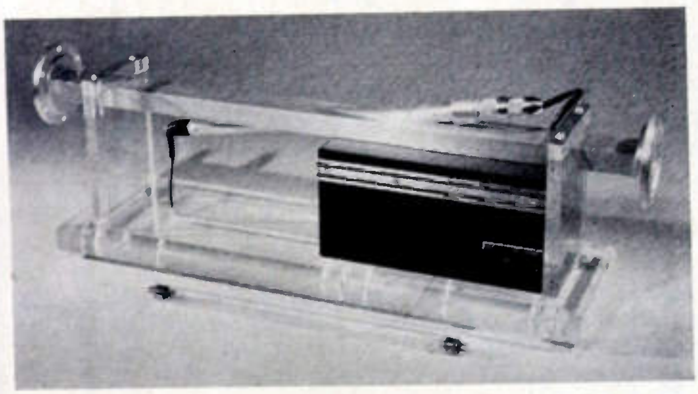

Fig. 8—Assembled unit showing discharge tube, waveguide and associated power supply.

D. Extended Frequency Systems

The performance of the above systems deteriorates rapidly at frequencies lower than that corresponding to about 0.8 of the cutoff wavelength of the guide (this is approximately the lower recommended transmission frequency of the guide). This is presumably due to the rapid increase in guide wavelength at frequencies below this limit. This situation may be alleviated and satisfactory operation obtained at frequencies closer to the cutoff frequency of the standard guide through the use of a ridge[13] as shown in the sketch of Fig. 9.

[13] S. B. Cohn, "Properties of ridge waveguide," Proc. I.R.E., vol. 35, p. 783; August, 1947.

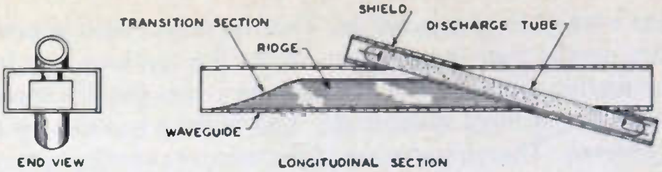

Fig. 9—Waveguide mount employing ridge to improve low-frequency performance.

It is also possible to mount the discharge tube on the top of a ridge as shown in Fig. 10. Through the utilization of the wide transmission bandwidth of a ridge waveguide, a very wide frequency range may be covered with one mount. For example, the A-149 tube mounted on the top of a 0.300×0.217 inch ridge in a standard RG-52U guide ($\frac{1}{2}$×1 inch) gave an average standing-

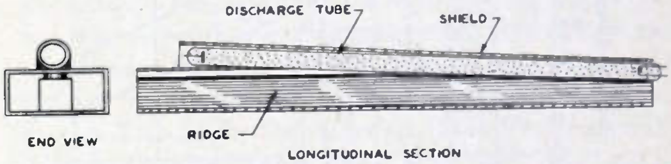

Fig. 10—Very wide-band ridge waveguide mount.

wave ratio of 1.12 over the measured range of 3 to 7.5 cm. Thus, the possibility of utilizing a single tube and mount over a bandwidth comparable to that covered by 3 standard guides is clearly demonstrated.

This ridge guide technique may also be used to obtain improved matching performance. For example, a 20-inch T5 discharge tube mounted on the top of a 0.75 ×0.94 inch ridge in RG-48U guide gave an average VSWR of only 1.03 over the recommended transmission bandwidth and at 0.9 of the cutoff wavelength the VSWR had only risen to 1.08.

III. Fluctuation Phenomena

Electrical discharges through gases are often accompanied by oscillations or fluctuations, and the present case is no exception. The present fluctuations occur at an audio frequency rate, i.e., a few thousand cycles per second. Their presence may affect the instantaneous match presented by the discharge to the guide.

Many microwave mixers allow considerable local oscillator energy to travel towards the antenna. If a gaseous discharge has replaced the antenna, the fluctuations in this discharge, since they modulate the match presented by the discharge, return more or less local oscillator energy to the crystal and thus modulate the crystal excitation. This appears in the receiver output as a low-frequency modulation of the noise. There is no direct evidence that the noise output of the gaseous discharge itself has been modulated.

Thus, the fluctuations may be observed as a modulation of the crystal excitation as well as at the terminals of the gaseous discharge tube. It has been found that there is not necessarily a one-to-one correspondence between such observations. That is, under certain conditions the microwave effects of the fluctuations may be negligible although the fluctuations across the discharge tube are of large magnitude. The inverse has not been observed. This suppression of the microwave effects may be accomplished in a number of ways which tend to increase plasma density, for example by an 1) increase in discharge current, a (2) decrease in tube diameter, an (3) increase in gas pressure, or by (4) the use of an inert gas of greater atomic weight.

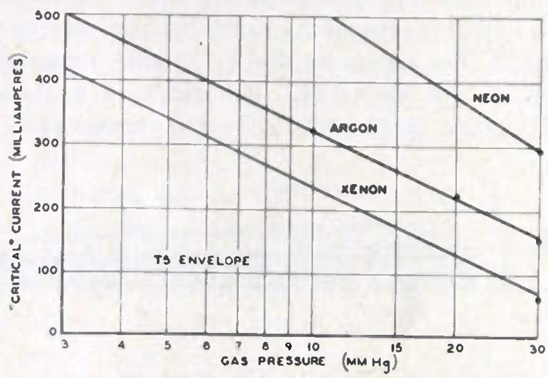

Fig. 11—"Critical" current versus gas pressure.

For a given tube, then, there exists a "critical" current above which the microwave effects of the fluctuations are suppressed. These critical currents have been measured over a wide variety of conditions and the experimental data have been surprisingly consistent. Fig. 11 shows the critical current plotted against the log of the pressure for discharges through various gases in a T5 envelope. A linear relation between the critical current and the log of the pressure appears to exist. Note that the critical current decreases with increasing pressure and with increasing atomic weight of the gas.

A change in tube diameter results in a change in slope of the lines of Fig. 11. This is shown in Fig. 12 for discharges through argon for various discharge tube

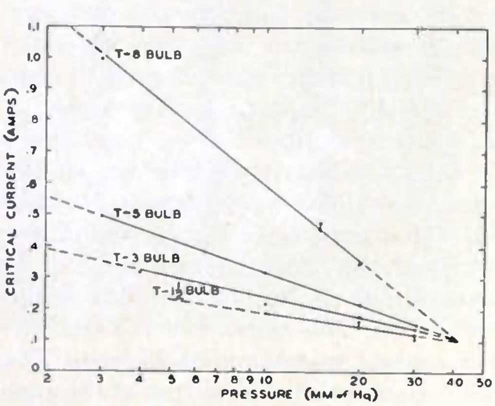

Fig. 12—Critical current in argon discharges versus gas pressure for various discharge tube diameters.

diameters. The data for argon have been collected and the empirical relation

$$\frac{I - 0.11}{0.13 + 0.56R^2} = \log_{10}\frac{42}{P}$$

between the critical current, discharge tube radius, and pressure has been found, as in Fig. 13, together with the experimental data. The data are quite consistent.

Another method of suppressing fluctuations in the discharge is the use of gas mixtures. In this case it is possible to suppress the fluctuations completely. That is, the fluctuations no longer exist across the tube terminals.[14] An example of such a tube is the commercial fluorescent lamp containing argon and mercury. To secure reliable operation, i.e., to insure the presence of an adequate density of vaporized mercury, the operating temperature should be somewhat higher than that given by standard operating conditions. Other possible gas mixtures which have been successfully tested are hydrogen in neon and nitrogen in neon. We note that in all of these cases the ionization potential of the added component approximates the metastable level of the inert gas.

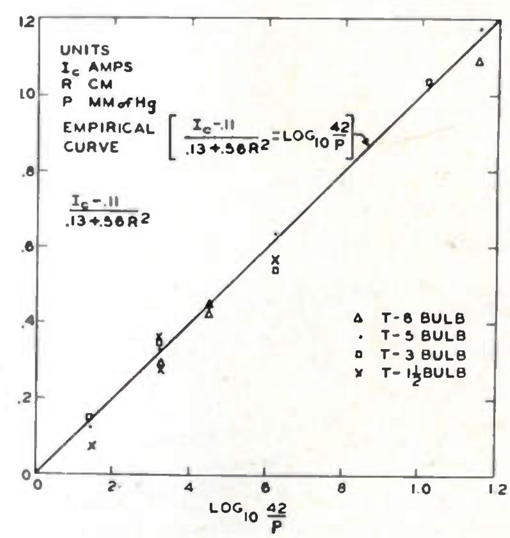

Fig. 13—Correlation between empirical curve and experimental points relating critical current, tube interior radius, and gas pressure for argon.

IV. Noise Aspects

The operation of a gaseous discharge waveguide termination as a noise source will be described in terms of its apparent noise temperature. That is, the available noise power from such a waveguide termination is a number of times greater than that from the waveguide terminated in a resistive load at a room temperature of 290° K. The discharge, therefore, may be considered to act as a resistive termination at a temperature that is a number of times greater than room temperature. This ratio will commonly be expressed in decibels, since it is proportional to a power ratio.

[14] It may, however, be necessary to suppress fluctuations due to a multipath instability in the anode region by other means to obtain complete freedom from fluctuations.

It is suggested that the noise temperature exhibited by a gaseous discharge employed as above is a measure of the electron temperature of the positive column of the discharge.[15] On this basis, one would anticipate noise temperatures on the order of those shown in Fig. 14 which was taken from Engel and Steenbeck.[16] It will appear that the measured noise temperatures approximate these calculated electron temperatures.

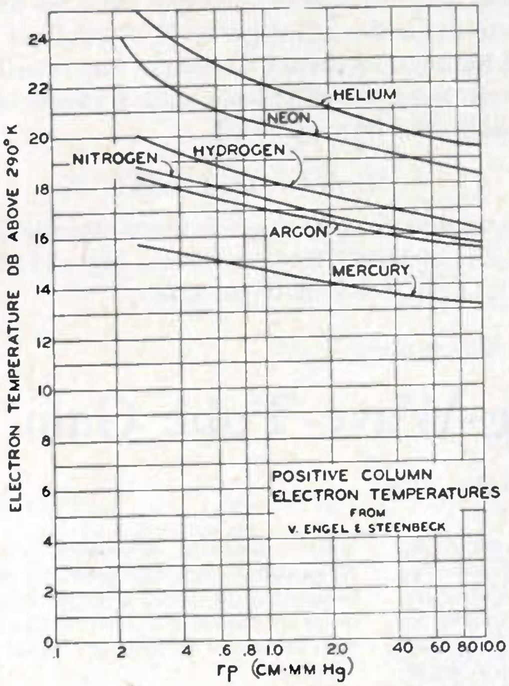

Fig. 14—Positive column electron temperatures.

Consider first the experimental noise measurements shown in Fig. 15. These measurements were made at 4,000 mc and the reference level of the left-hand scale refers to the noise temperature of a germicidal tube.[17] The right-hand scale is an absolute scale against which the calculated and measured electron temperatures were plotted. The reference level of the left-hand scale was placed at an absolute level of 15.5 db on the basis of the published data of Mumford.[10] For the most part, the variation of noise temperature with a change in pressure varies qualitatively in the manner predicted by theory. A serious deviation from this statement occurs for the very low pressure neon tubes. While these tubes were difficult to match, a sufficient number of tubes are included to believe that this was the actual state of affairs in this experiment.

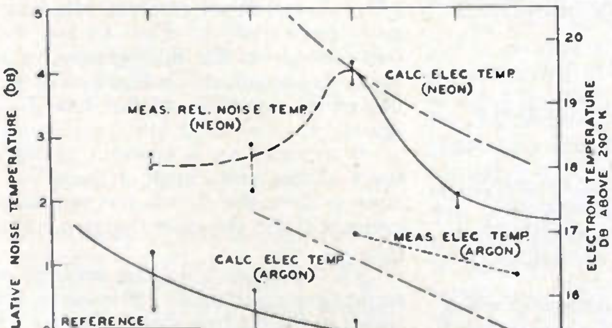

Fig. 15—Relative noise temperatures measured at 4,000 mc as a function of gas pressure for argon and neon. Calculated data for electron temperatures are also shown.

It is noted that the noise temperatures of the neon tubes lie above those for argon approximately the amount predicted by an electron-temperature hypothesis. A single helium tube (not shown on the figure) of 3-mm Hg pressure was also measured. Its relative noise level was +5.2 db in good agreement with theory. Discharges in xenon give a noise temperature approximately 2 db below the reference level. Thus, insofar as relative measurements are concerned, the relative measured noise temperatures respond to a change in the kind of gas and to a change in the gas pressure in a manner that is in qualitative agreement with that predicted by an electron-temperature hypothesis.

TABLE II

Noise Temperature Measurements

Frequency (mc)	*Tube Designation*	*Calculated Electron Temp.* (db above 290°K)	*Measured Noise Temp.* (db above 290°K)	*Signal Generator*
2,930	A148	15.3	15.5	"A"
2,930	A147	15.4	15.5	"A"
4,000	A147	15.4	15.5	"A"
5,990	A147	15.4	16.6	"B"
6,670	A147	15.4	17.4	"B"
6,930	A147	15.4	16.0	"B"
9,500	A147	15.4	17.3	"B"
9,500	A149	15.4	17.3	"B"
24,000	A146	16.3	16.0	"C"

On the other hand, if the absolute level is established, as above, by reference to the published data of Mumford the absolute noise temperatures lie up to about a decibel below the calculated (and also the measured) electron temperatures.

It should be pointed out that the inert gas discharge tubes exhibit no dependence of the noise temperature upon the operating temperatute such as that found[10] in commercial fluorescent (germicidal) lamps, due to the change in mercury vapor pressure with temperature. The noise temperature is similarly independent of the magnitude of the discharge current.

Absolute noise measurements have been attempted at various frequencies throughout the shf region. Standard signal generators were used as a calibrated standard. The results of these measurements are shown in Table II

[15] As this paper was being written, the work of P. Parzen and L. Goldstein deriving an analytical expression for the noise power was called to our attention. Their theoretical analysis shows that the noise power can be separated into two parts, one corresponding to the electron temperature and the other, generally negligible under the present conditions, due to the dc discharge current. See P. Parzen and L. Goldstein, "Current fluctuations in a d-c discharge plasma," *Bull. Amer. Phys. Soc.*, vol. 25, p. 40, no. 3; April 27, 1950.

[16] See page 86 of footnote reference 12.

[17] This is an argon-mercury tube similar to a fluorescent lamp but without a fluorescent coating.

The accuracy of the above measurements is probably of the order of ±2db (roughly the accuracy of the calibration of the standard signal generators). These measurements provide evidence that the noise temperature is substantially uniform over the shf region and, within the accuracy of measurement, the measured noise temperatures show fair agreement with the electron temperatures. Also within the accuracy of measurement, agreement with published data is found at those frequencies for which such data exist.[10,18]

V. Conclusions

The above work has demonstrated that the gaseous discharge type of noise source introduced by Mumford is useful throughout the shf region. By mounting the discharge tube diagonally into the waveguide, it has been shown that the discharge provides an excellent untuned waveguide termination which is matched over the entire transmission bandwidth of the guide. The hypothesis that the noise temperature of the discharge corresponds to the positive column electron temperature has been suggested and experimental evidence offered in its support. Through the use of discharges through pure gases, there is no dependence of the noise temperature upon the ambient temperature. A study of the effects of fluctuations in the discharge has revealed the conditions under which the microwave effects of these fluctuations may be suppressed.

[18] Recent unpublished measurements by J. E. Sees and H. Corbett at the Naval Research Laboratory have indicated a noise temperature of 15.5±0.25 db at 6,975 mc for our argon-filled discharge tube bearing the experimental designation A-147.

VI. Acknowledgment

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The Calculation of Traveling-Wave-Tube Gain*

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Summary—Essential information for calculation of traveling-wave-tube gain is summarized and condensed in this paper. The important relations are documented, presented in a concise form for simplified computation, and developed as a nomograph. The conclusions have been found to be in agreement with measurements on six different tube designs.

As a RESULT of recent developments in traveling-wave-tube theory, it is now possible to calculate the performance of such a tube with some accuracy. Because of the amount and complexity of the written material on traveling-wave tubes, however, it is sometimes difficult to separate the relations which are of basic importance to traveling-wave-tube operation from those presented for completeness and pedagogical purposes. Also some of the work is not complete in giving actual working relationships which are valid in the light of subsequent disclosures. This paper is intended to summarize the available information and make it useful to those who are more interested in using the results than in understanding the logic by which they were obtained.

The most valid available relationships relating to gain and power output, taken principally from the publications of Pierce, are grouped and documented in Fig. 1. The main purpose of this block diagram is to provide a guide to the literature for the individual interested in pursuing any particular aspect of the problem. At the same time it may serve as a guide in calculation. It can be seen that the process of calculation of even such a fundamental thing as gain is rather complex and involves a large number of parameters.

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The attached nomograph for calculating traveling-wave-tube gain has been tested on tubes made in England, Stanford University, and at the Bell Telephone Laboratories, for a variety of conditions. The comparisons, summarized at the end of the paper, are remarkably good and are well within the accuracy that can be expected, considering the lack of precision of knowledge of the actual electron beam diameter, and impedances of the wave propagating circuit.

In the following discussion some of the relations are combined and simplified and put in terms of more basic circuit parameters for simplified calculation. (Symbols are defined in the Glossary.)

Gain

The fundamental equation for gain is

$$G = (A + BCN)\text{db}, \qquad (1)^{1}$$

where A is the loss in initiating the amplified wave, and BCN is the asymptotic gain.

N is the number of active wavelengths in the tube. Thus, BC is the asymptotic gain per wavelength.

C is defined as the gain parameter which depends on the circuit and electron beam impedances.

$$C^3 = K\frac{I_0}{4V_0}, \qquad (2)^{2}$$

where for the ideal helical surface

$$K = \frac{E^2}{2\beta^2 P} = \frac{1}{2}\frac{\beta}{\beta_0}\left(\frac{\gamma}{\beta}\right)^4 F^3(\gamma a). \qquad (3)$$

Now a practical circuit may differ by a significant amount from the idealized helical surface. In a practical case this may be calculated, measured, or deduced on a basis of performance. We will therefore include a constant K_2/K which is equal to the ratio of the impedance parameter of the actual circuit, to that of an idealized helical surface. Therefore,

$$C = \left(\frac{\beta}{\beta_0}\right)^{1/3}\left(\frac{\gamma}{\beta}\right)^{4/3} \cdot F(\gamma a)\left(\frac{K_2}{K}\right)^{1/3}\left(\frac{I_0}{8V_0}\right)^{1/3} \cong 3.98\left(\frac{K_2}{K}\right)^{1/3} F_1F_2\frac{(I_0)^{1/3}}{(V_0)^{1/2}}, \qquad (4)$$

where F_1 and F_2 are graphical relationships given respectively by Figs. 3.4 and 3.5 of reference (3) in the Bibliography, or perhaps more accurately by Fig. 5 of reference (4) where essentially $(F_1F_2)^3/2$ is plotted against γa.

B accounts for a constant multiplier, space charge and circuit attenuation. In order to determine B, it is first necessary to evaluate C and the space charge parameter Q.

Q is dependent in a rather involved way on the circuit parameters and coupling of the circuit to the electron stream. For the ideal helical surface and assuming a hollow cylindrical beam, Q_1 is given in Fig. 8.12 of reference (3) or Fig. 1 of reference (4). The relation for solid beams is given in Fig. 4 of reference (4). Q is by definition inversely proportional to the circuit impedance parameter K, i.e.,

$$Q = \frac{\beta_e}{2\omega C_1 K} \qquad (5)^{3}$$

and therefore for a practical circuit

$$Q_2 = Q_1\frac{K}{K_2}, \qquad (6)$$

[1] See equation (52) of reference (1) or equation (2.38) of reference (3) in the Bibliography.
[2] See equation (2.43) of reference (3) or equation (19) of reference (1) in the Bibliography.

[3] See equation (7.15) of reference (3) in the Bibliography.

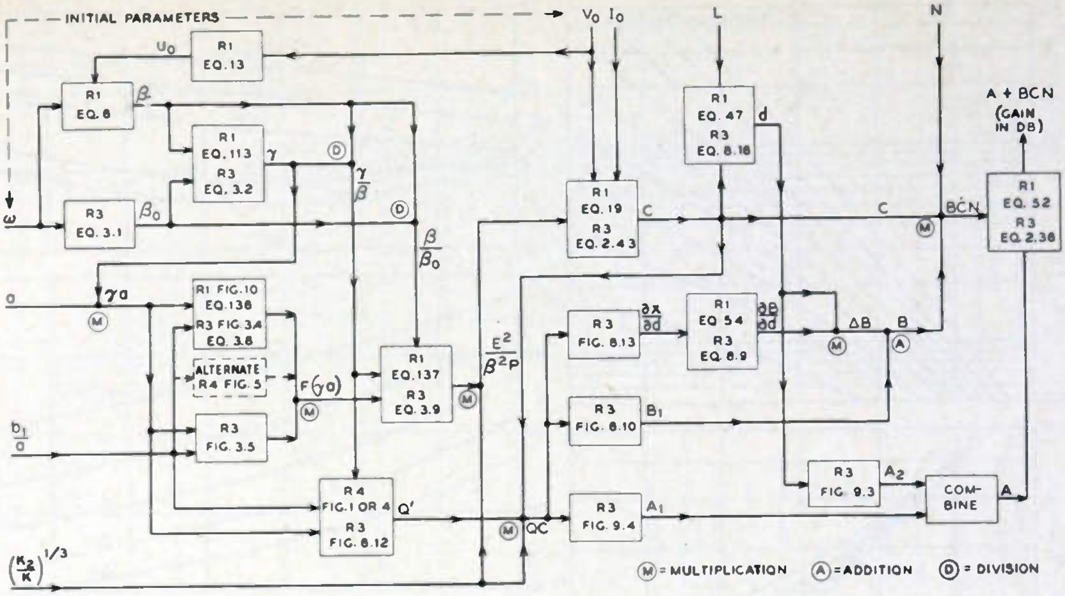

Fig. 1—Documentation of essential traveling-wave-tube relations.

where the subscripts 1 and 2 indicate the values of Q for the helical surface and the actual circuit, respectively.

Now B is a function of both the space charge factor QC and the attenuation parameter d.

$$d = \frac{L}{54.5\,C}. \qquad (7)^4$$

The effect of QC on B may be considered first and attenuation used as a correction. In this case, taken B_1 as the value of B for zero attenuation.[5] For small attenuations d may be accounted for by taking

$$B = B_1 + 54.5\frac{\partial x_1}{\partial d}d, \qquad (8)$$

where $\partial x_1/\partial d$ is given as a function of QC in Fig. 8.13 of reference (3). B may be used directly in (1), or if we substitute (8) in (1) and substitute (7) for d, we get

$$G = A + B_1CN + \frac{\partial x_1}{\partial d}LN. \qquad (9)$$

Thus, attenuation may be accounted for by subtracting a fraction of the total attenuation from the lossless gain.

If QC is <0.15 and attenuation is large, it may be best first to obtain B_2 (the value of B for zero space charge) from Fig. 8.5 of reference (3) and to correct for space charge later. In this case

[4] See equation (47) of reference (1) or equation (8.15) of reference (3) in the Bibliography.
[5] See Fig. 8.10 of reference (3) in the Bibliography.

$$B = B_2 + 54.5\frac{\partial x_1}{\partial(QC)}QC \qquad (10)$$

where $\partial x_1/\partial(QC)$ is given as a function of d in Fig. 8.14 of reference (3).

The effect on B of space charge and attenuation taken together has been computed by Quate, and the curves in Figs. 2 and 3 are drawn from his unpublished data.

If attenuation is not applied uniformly and is not greater than 100 C db per wavelength, it may be accounted for to a first approximation by averaging the attenuation over the length of the tube. If the attenuation is greater than this, the effects of the discontinuities can be estimated from Fig. 9.5 or Fig. 9.8 of reference (3).

A is the initial loss at the input boundary and is given as a function of space charge parameter QC in Fig. 9.4 of reference (3) and as a function of the loss parameter d in Fig. 9.3 of reference (3). In the absence of a relation for combined d and QC, A may be estimated with sufficient accuracy for most purposes by interpolation from these curves. The combined effect has been computed,[6] however, and is presented in Figs. 2 and 3 of this paper. In the case of discontinuous attenuation, the amount of loss also depends on its position as indicated in Fig. 9.6 of reference (3) which may simply be taken as a design criterion that the loss should be as small as possible up to a distance corresponding to $CN>0.2$ from the input end of the circuit.

[6] Unpublished data by C. F. Quate.

Summary of Relationships

Fig. 1 of this paper is a block diagram summarizing the relations and source material that may be used in calculating traveling-wave-tube properties.

In Fig. 2, the essential relations are combined and condensed to apply specifically to *solid-beam* cylindrical structures and uniform attenuation, in a concise form and utilizing such approximations as are legitimate.

In Fig. 3 the data of Fig. 2 are plotted in nomograph form. In using this, the attenuation is either averaged over the helix length or else, if lumped, must be accounted for as a subsequent correction according to Fig. 9.8 of reference (3).

In the last two figures the general relationships are taken from the works of Pierce (1, 2, 3) and the graphs for Q and F_1F_2 (grids 2 and 8) from Fletcher (4). The curves for A and B (Fig. 2) and grid 12, (Fig. 3) are plotted from unpublished information furnished by Quate.

In the absence of more complete analytical work this is probably as well as can be done and limits the resultant accuracy to the order of a few db. The actual circuit impedance parameter and the effective beam diameter are usually not sufficiently well known to expect greater accuracy than this provides. Comparison of the nomograph with a number of tubes described in the literature gives a remarkably good correlation, as can be seen from the comparisons in Table I.

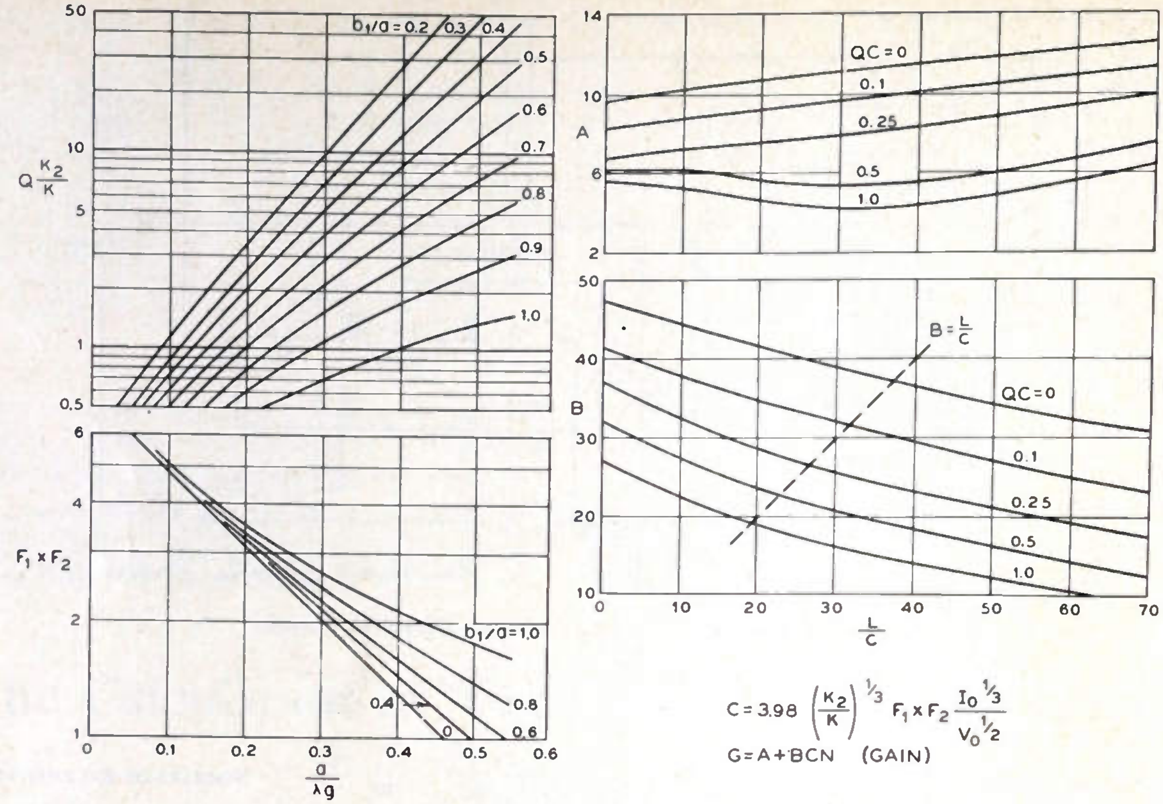

Fig. 2—Summary of relations necessary to calculation of traveling-wave-tube gain.

Glossary of Symbols

β = a phase constant $=\omega/v=2\pi/\lambda g$
β_e = electron phase constant
$\beta_0=\omega/c=2\pi/\lambda a$
γ = radial propagation constant $=\sqrt{\beta^2-\beta_0^2}$
λ_a = free-space wavelength
λ_g = circuit wavelength
η = the ratio of charge to mass of the electron $=1.759\times10^{11}$ coulomb/kg
ω = radian frequency
A = ratio of power levels in increasing wave component and total initial power at the beginning of a traveling-wave tube, measured in db
a = circuit radius
b = beam radius
B = a factor related to the increase of the increasing wave per wavelength
c = the velocity of light $=3\times10^8$ meters per second
C = a parameter for gain per unit length
d = a factor proportional to attenuation
E = electric field acting on the beam in the direction of propagation
$F_1=F_1(\gamma a)$ = a function relating (γa) and K_1
$F_2=F_2(\gamma a)$ = a factor indicating the effect of beam diameter on K_1
G = gain in db
I_0 = the dc beam current, amperes
K = circuit impedance for idealized helical surface $=E^2/2\beta^2P$
K_2 = actual circuit impedance
L = the attenuation in db per circuit wavelength
N = the length of circuit in wavelengths
P = the transmitted power at any point, watts
Q = a parameter related to the beam and circuit coupling
V_0 = The dc beam voltage, volts
x_1 = a factor proportional to gain per wavelength

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(1) J. R. Pierce, "Theory of the beam-type traveling-wave tube," Proc. I.R.E., vol. 35, pp. 111–123; February, 1947.
(2) J. R. Pierce, "Effect of passive modes in traveling-wave tubes," Proc. I.R.E., vol. 36, pp. 993–997; August, 1948.
(3) J. R. Pierce, "Traveling-Wave Tubes," D. Van Nostrand Co., Inc., New York, N. Y.; 1950.
(4) R. C. Fletcher, "Helix parameters used in traveling-wave tube theory," Proc. I.R.E., vol. 38, pp. 413–417; April, 1950.
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TABLE I

Tube	Frequency (Mc)	V_0 (Volts)	I_0 (ma)	$\frac{a}{\lambda_g}$	$\frac{b}{a}$	$\frac{K_2}{K}$	L	N	G (Measured)	G (From Nomograph)
BTL Tube #1	4,000	1,600	40	0.406	0.55	0.56	0.88	50	38	38
BTL Tube #2	4,000	1,600	40	0.192	0.55	0.56	(1.0+lump)	20	39	43
BTL Tube #3 (By Hollenberg)	200	56	49	0.286	0.55	0.56	(lumped)	10	37	38
BTL Tube #4 (By Little)	48,000	1,000	0.87	1.072	0.85 to 0.9	0.56	0.3	91	3	0 to 4.5
Stanford Univ. Tube[1]	3,000	720	0.15	0.26	0.36	0.56	0.2	60	21	23
British Tube (Reference (5))	4,300	1,400	1.5	0.253	0.55	0.56	0.52	34.4	20	20.5

[1] From unpublished report by C. F. Quate.

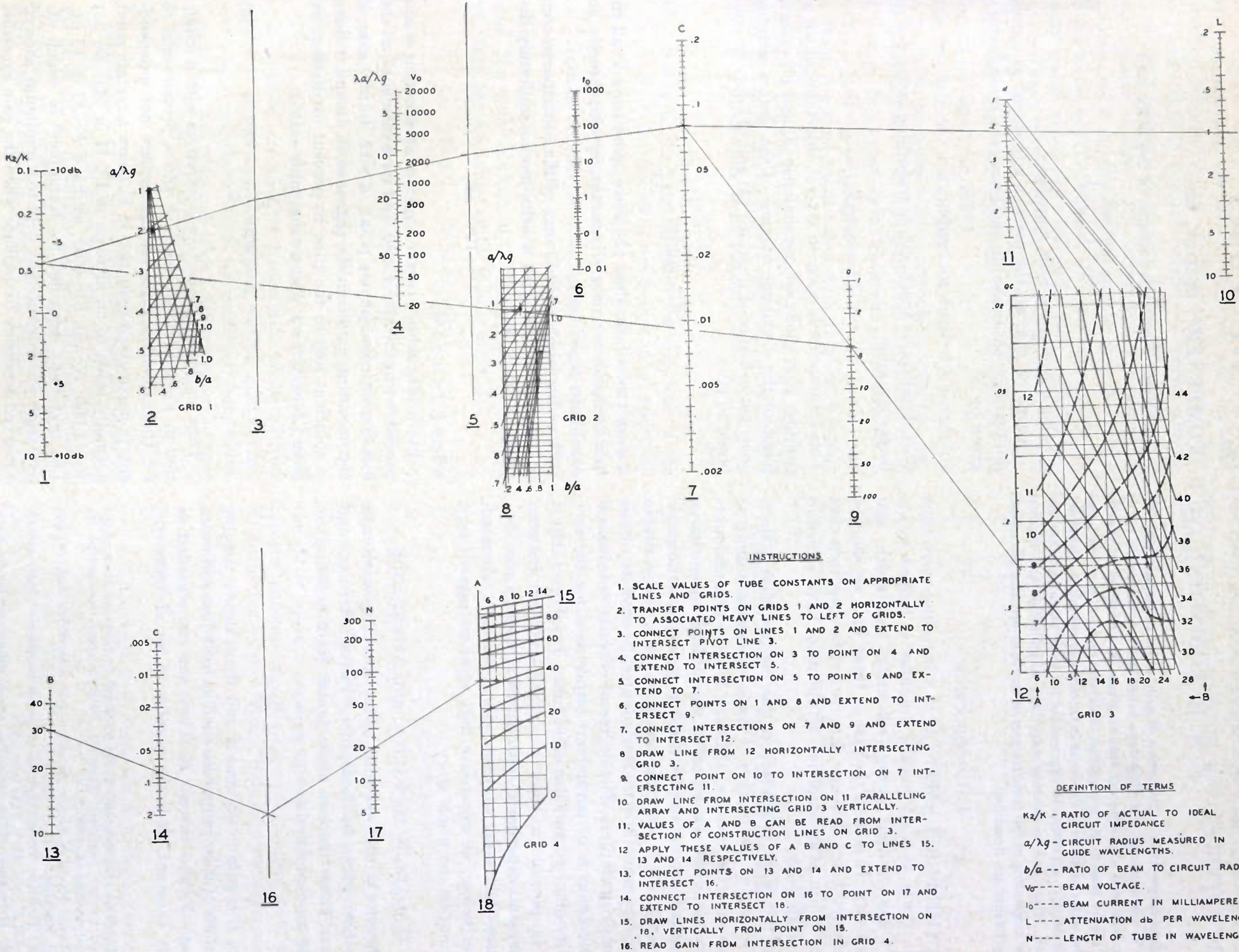

Fig. 3—Nomograph of traveling-wave-tube relations.

Noise in Traveling-Wave Tubes*

F. N. H. ROBINSON†, AND R. KOMPFNER†, FELLOW, IRE

Summary—After a discussion of gain, noise factor and attenuation in conventional traveling-wave tubes (twt), a dispersive twt is described and experimental results given. Examination of results and theory lead to the recognition of several types of partition noise as the cause of much of the noise in traveling-wave tubes. These are analyzed and approximate expressions for their magnitude are obtained. Noise due to electron-wave type amplification and thermal noise originating in the resistive part of the helix are also discussed. Experiments are described.

I. Introduction

The performance of a traveling-wave tube (twt) as a microwave amplifier is judged by its gain, output power, efficiency, bandwidth, and sensitivity. It is with the last of these that we shall be concerned here, although it must not be forgotten that to be of use as an amplifier, a tube must have a certain minimum gain.

As this work progressed it was realized that little is known about the process of space charge smoothing at microwave frequencies, although long beams undoubtedly appear to be smoothed. A study of the space-charge smoothing process itself was felt to be somewhat outside the scope of the work on low-noise twt's, and a simple model of a smooth beam was adopted to serve as a basis for our calculations. The degree of effective smoothness of the beam is of paramount importance for the sensitivity of a twt, and also other tubes. It is hoped that it will be possible to show that the agreement between theory and experiment is sufficiently good to give confidence that some understanding has been obtained. Nevertheless, we are aware that what we present here may not be the whole story.

II. Gain, Noise Factor[1] and Attenuation

Theories of the operation of the traveling-wave tube have been given by Kompfner,[2] Pierce,[3] Bernier,[4] and others. These theories agree (apart from notation) in giving an expression for the gain:

$$G = -A + BCN \tag{1}$$

where G is gain in db, and A and B are slowly varying functions of tube properties such as helix loss and beam voltage V_0 and current i_0. They may be regarded as nearly constant. N is the number of wavelengths corresponding to the active length of the helix, and

$$C = \left(\frac{i_0 Z}{8V_0}\right)^{1/3}, \tag{2}$$

Z being the effective helix impedance.

Assuming that all the noise is introduced as current fluctuations (shotnoise) in the beam, the noise factor is given by

$$F = 80\Gamma^2 V_0 C. \tag{3}$$

Γ^2 is the effective space-charge smoothing factor defined by the equation

$$\overline{i_n^2} = 2ei_0\Gamma^2\Delta f.$$

This assumption is open to several objections which we consider later; however, we shall use (3) as a provisional basis for the discussion of some of the factors affecting noise performance. Equations (3) and (1) may be combined to give

$$F = 80\Gamma^2 V_0 \frac{A+G}{BN}, \tag{4}$$

a more useful form than (3), since apart from Γ^2 all the factors are either nearly constant, or else directly accessible to experiment.

Over the whole useful range of the other parameters $B \sim 40$. For a tube with a helix having uniformly distributed loss L db

$$A \sim 10 + \alpha L,$$

where $1/3 < \alpha < 2/3$.

For a tube with a center region of attenuation of appreciable length, $20 < A < 30$, and for a tube with all the loss concentrated at one point, $A \sim 12$. In any case, if the operating gain is in the region of 20 db and if there is sufficient loss in the tube to prevent oscillation, we have $(A+G)/B \approx 1$, and we may now write

$$F \approx 80\Gamma^2 \frac{V_0}{N}. \tag{4a}$$

Thus on this simple picture F improves with a reduction in V_0 and an increase in N and is proportional to Γ^2. Unfortunately these parameters are not independent; in particular a reduction in V_0 below a certain point normally leads to an increase in Γ^2. If in (4a) we put $V_0 = 1{,}000$, $N \sim 50$, $\Gamma^2 \sim 0.02$, we have $F \sim 32$ or 15 db. The first twt's described by Kompfner[2] in which attenuation was intentionally kept to a minimum, showed a marked tendency to oscillate when the gain exceeded unity, although by carefully terminating the input and ouptut he was able to obtain $F = 11$ db, with 11-db gain in a tube 60 c/m long at 3,000 mc.

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[1] H. T. Friis, "Noise figures of radio receivers," (correction), Proc. I.R.E., vol. 32, p. 419; July, 1944.

[2] R. Kompfner, "The traveling-wave tube: centimetre-wave amplifier," *Wireless Eng.*, vol. 24, pp. 255–267; September, 1947.

[3] J. R. Pierce, "Theory of the beam-type traveling-wave tube," Proc. I.R.E., vol. 35, p. 111–123; February, 1947.

[4] J. Bernier, *L'Onde Electrique*, vol. 27, pp. 231–243; June, 1947.

III. Dispersive Traveling-Wave Tubes

It was not until the advent of tubes with high attenuation deliberately introduced in the helix described by Pierce[5] that it became possible to make reliably stable twt's having a large gain. These tubes, however, were reputed to have noise factors in the region of 25 to 30 db, and it appeared to the authors that this might be due to the effect of the loss. In attempting to design a tube which would be stable, give a gain of 20 db, and not need more than 15-db loss in the helix, one of us was led to the dispersive twt (described in a report presented at the IRE Electron Tube Conference at Cornell University in 1948.[6]

In this tube use was made of the dispersive properties of the helix to restrict the electronic bandwidth, i.e., the bandwidth over which beam wave interaction could take place, to a value for which it was easily possible to obtain a good match in and out of the tube. In this way feedback is reduced and tubes with 20-db stable gain were made.

A helix is dispersive (i.e., the velocity of the wave supported by the helix varies with frequency fairly rapidly) when the radius of the helix a is such that

$$ka \leqq 1.2, \tag{5}$$

where $k=2\pi/\lambda$, the axial propagation constant of the helix. At 3,000 mc (which is the frequency at which all the work was done), a helix of 3-mm diameter supporting a wave with a phase velocity corresponding to 1,600-volt electrons satisfies this condition and the bandwidth may be reduced to 200 mc or less. It is a simple matter to obtain a good match to the external circuit over this band. If α is the fraction of the power reflected at each end, then the maximum stable gain is given by

$$G \leqq L - 20 \log_{10} \alpha \text{ db}. \tag{6}$$

If $\alpha=\frac{1}{4}$, $L=13$, the tube will have a stable gain of up to 25 db. In fact it was found possible to use tubes with an insertion loss of only 11.5 db (which was entirely due to loss in the wire of the helix) and to obtain a stable gain of 30 db.

Curves measured for such a tube are shown in Fig. 1, and it will be seen that the bandwidth calculated from the beam voltage-frequency curve and the gain-voltage curve agree excellently with the measured bandwidth. This result was obtained with a beam current of only 150μa. The noise factor measured was 17 db.

IV. Partition Noise

Examination of (4) shows that the observed improvement in F could not have been due to the reduction of the attenuation, drastic as this was (this could only account for an improvement of, at the most, 2 db). Similarly the spectacular decrease in beam current—by nearly a factor of 100—made possible in the dispersive twt could not by itself account for the noise factor improvement, since the current does not appear in equation (4).

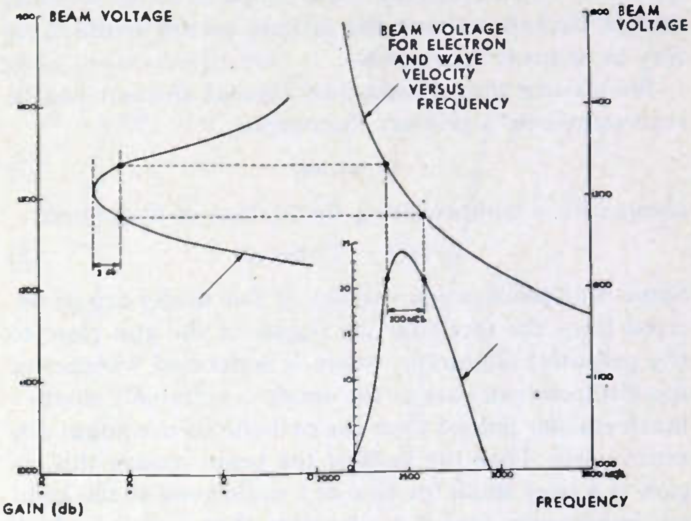

Fig. 1—Graphical construction of frequency response curve of a dispersive traveling-wave tube based on measured beam-voltage-versus frequency curve.

However, this drastic reduction of beam current is essential in obtaining a high electron gun and collector efficiency. It was comparatively simple to project a beam of, say, 200μa at 1,600 volts through a helix 3 mm internal diameter and 50 cm long without losing more than 1μa on the way. The beam was generated by a Pierce-type gun and held together by a long *uniform* magnetic field of 200 gauss.

This suggested that the noise improvement was due to a reduction of partition noise of one kind or another, a hypothesis already put forward earlier.[2]

V. Space-Charge Smoothing

Space-charge smoothing at low and medium frequencies is well understood, and has been described in a number of classic papers.[7–9] The theory rests on the assumption that the transit time of an electron crossing the diode space is small compared with a cycle of the measuring frequency. Rack's theory[9] indeed goes somewhat further by postulating that only the *differences* in transit time due to the Maxwellian spread of initial velocities need be small. It can be shown, however, that in guns used in twt's none of these conditions can be justified.[10,11]

[5] J. R. Pierce, and L. M. Field, "Traveling-wave tubes," Proc. I.R.E., vol. 35, pp. 108–111; February, 1947

[6] F. N. H. Robinson, "TWT's with dispersive helices," *Wireless Eng.*, in publication.

[7] B. J. Thompson, D. O. North, and W. A. Harris, "Fluctuation in space-charge-limited currents at moderately high frequencies," *RCA Rev.*, vol. 9, pp. 269–286; January, 1940.

[8] W. Schottky and E. Spenke, *Wiss. Veroff. Siemens-Werken*, vol. 16, p. 1; 1937.

[9] A. J. Rack, "Effect of space-charge and transit-time on the shot noise in diodes," *Bell. Sys. Tech. Jour.*, vol. 17, pp. 592–620; October, 1938.

[10] D. K. C. MacDonald, "Transit time phenomena in electron streams II," *Phil. Mag.*, vol. 41, pp. 863–872; September, 1950.

[11] Treatments based on Rack's theory such as that outlined by J. R. Pierce at the IRE Conference, Princeton, N. J., 1949, are open to the same objection.

So far no theory of how a smooth beam of electron is produced at frequencies where the transit time in the gun is several cycles is available. However, it is an experimental fact that smoothing exists at these frequencies. Therefore, in lieu of a complete theory we shall assume henceforth that the process occurs in the same way as at lower frequencies.

We assume that a smoothed current element can be represented by a primary fluctuation

$$di \sin \omega t, \tag{7}$$

along with a compensating fluctuation of magnitude

$$-di(1 - \Gamma) \sin \omega t. \tag{8}$$

Some confidence in the validity of this model can be derived from the fact that the region in the gun close to the potential minimum, where it is decided whether or not electrons can pass to the anode is extremely small—much smaller indeed than the cathode to minimum distance itself. Thus the bulk of the beam crosses this region in a very small fraction of a cycle even at the highest frequencies, and it is plausible that a model which assumes that excess electrons are accompanied by "deficit" electrons is valid as at lower frequencies.

Further we assume that while the primary fluctuation is confined to a small element dA of the beam cross section A, the compensating fluctuation is spread over the whole beam. This can be shown to be true in certain circumstances; it is a consequence of the sideways initial velocities of the electrons at the cathode, since the direction and magnitude of such sideways components are not correlated with the component in the beam direction. The concept of spatial noncorrelation of initial and compensating fluctuations is not new; it has in fact been used in the past to explain partition noise at lower frequencies, e.g., in the literature.[7]

The assumption of sinusoidal fluctuations can be justified only by considering the shot noise in a very narrow band df. Thus we write

$$\tfrac{1}{2}\overline{(di)^2} = 2eI_0 df \frac{dA}{A}, \tag{9}$$

where I_0 is the total beam current.

VI. Sources of Partition Noise

Equation (4) would be correct if the beam were of infinitesimal cross section and if all electrons had the same axial velocity. In fact, however, a beam has a finite cross section and the axial velocities of electrons are by no means uniform.

In this section we discuss a number of sources of partition noise, assuming that they are independent processes, so that if several are present together this over-all effect can be expressed by a summation. Thus, if the initial degree of disorder of the beam is Γ_0^2 and processes $m, n \cdots k$ produce separately disorders Γ_m^2, Γ_n^2, $\cdots$, Γ_K^2, the effective overall degree of disorder is

$$\Gamma^2 = \Gamma_0^2 + \Gamma_m^2 + \Gamma_n^2 + \cdots + \Gamma_k^2. \tag{10}$$

This assumption obviously implies that both Γ^2 and each partial Γ_m^2 are small compared with unity.

We now discuss partition noise due to direct interception of part of the beam current and nonuniform coupling of the beam to the helix. These are a consequence of the finite size of the beam and may be distinguished from the following cases of partition noise due to the nonuniform axial velocities, due to nonparallel trajectories and nonuniformity of space potential within the beam.

A. Direct Interception

Partition noise due to this cause is well known at low frequencies,[7] and it is an experimental fact that it can also occur at microwave frequencies.

By way of an introduction to the more sophisticated types of partition noise we give in outline a derivation of the noise due to interception of a fraction α of the beam, using the concept of an initial elementary fluctuation current $|i|$ and its associated compensating current $-|i|(1-\Gamma)$ which latter is spread over the whole beam. We are interested in the noise in the remainder $(1-\alpha)$ of the beam. Consider the following currents:

Initial fluctuation in $(1-\alpha)$:

$$\sqrt{1-\alpha}\,|i|. \tag{11}$$

Compensating fluctuation in whole beam:

$$-\sqrt{1-\alpha}\,|i|(1-\Gamma). \tag{12}$$

Compensating fluctuation in $(1-\alpha)$:

$$-(1-\alpha)\sqrt{1+\alpha}\,|i|(1-\Gamma). \tag{13}$$

Initial fluctuation in α:

$$\sqrt{\alpha}\,|i|. \tag{14}$$

Compensating fluctuation in the whole beam:

$$-\sqrt{\alpha}\,|i|(1-\Gamma). \tag{15}$$

Compensating fluctuation in $(1-\alpha)$:

$$-(1-\alpha)\sqrt{\alpha}\,|i|(1-\Gamma). \tag{16}$$

Equations (11) and (13) are phase related being parts of one particular fluctuation; the phase of (16) is completely random with respect to (11) or (13), hence the total fluctuation in $(1-\alpha)$ is obtained by adding (11) to (13), squaring and adding to the square of (16):

$$\overline{i_{1-\alpha}^2} = \overline{i^2}\{[\sqrt{1-\alpha}\,[1-(1-\alpha)(1-\Gamma)]]^2 + [-(1-\alpha)\sqrt{\alpha}\,(1-\Gamma)]^2\}.$$

The current fluctuation in $(1-\alpha)$ is therefore a fraction

$$\alpha(1-\Gamma_0^2)+\Gamma_0^2$$

of the completely random fluctuations, and can thus be identified with the effective Γ^2 of that portion. If $\Gamma_0^2 \sim 0$

$$\Gamma_{eff}^2 \approx \alpha \tag{16a}$$

and it is plain that to get effective values of Γ^2 of the order of 0.01 less than 1 per cent of the beam must be intercepted.

B. Nonuniform Coupling (Induced Partition Noise)

Microwave currents can be induced in circuits as a consequence of free electron currents in their vicinity. There is a corresponding "induced" partition noise caused by nonuniform coupling between the circuit and individual elements of the beam.

In traveling-wave tubes the axial field due to the power flow in the helix varies across the inside of the helix according to

$$E(r) = E(0)I_0(kr). \quad (17)$$

$E(0)$ is the field in the axis, and I_0 is the modified Bessel function. The coupling between beam and helix is therefore not constant over the whole of the beam and the voltage induced in the helix by an initial localized fluctuation di_s is not exactly compensated by the voltage induced by the compensating fluctuation di_c which is spread over the whole beam. By virtue of (17) we can express the voltage induced by a fluctuation i at radius r by

$$V(r) = i\phi(r) = i\phi(0)I_0(kr). \quad (18)$$

If we consider an annulus of radius r, thickness dr, a fraction $i_0(2rdr/b^2)$ of the beam current i_0 (assumed of uniform density and radius b) will flow through it. The fluctuation associated with this current is

$$di_s = \left(\frac{2rdr}{b^2}\right)^{1/2}(2ei_0\Delta f)^{1/2} \quad (19)$$

and the compensating fluctuation

$$di_c = -(1-\Gamma_0)di_s. \quad (20)$$

The noise voltage induced in the helix by di_s is $dV_s = \phi(r)di_s$ and the voltage induced by di_c, which is spread over the whole beam, is $dV_c = \bar{\phi}di_c$, where

$$\bar{\phi} = \frac{1}{\pi b^2}\int_0^b \phi(r)\cdot 2\pi r dr = \phi_0\frac{2}{kb}I_1(kb) \quad (21)$$

and is the average value of ϕ across the beam. Thus the net noise voltage associated with this current element is

$$dV = (\phi(r) - (1-\Gamma_0)\bar{\phi})di_s \quad (22)$$

and the mean square noise voltage induced by the whole beam is

$$\overline{V_n^2} = \sum (dV)^2,$$

giving

$$\overline{V_n^2} = \int_0^b (\phi(r) - (1-\Gamma_0)\bar{\phi})^2\cdot 2ei_0\Delta f\cdot\frac{2r}{b^2}dr. \quad (23)$$

If we define the effective Γ_e^2 by the relation

$$\overline{V_n^2} = \Gamma_e^2\,\overline{\bar{\phi}^2}\cdot 2ei_0\Delta f \quad (24)$$

evaluating (23), we obtain

$$\Gamma_e^2 = \Gamma_0^2 + \Gamma_i^2 \quad (25)$$

$$\Gamma_i^2 = \left(\frac{kb}{2}\right)^2\left\{\left(\frac{I_0(kb)}{I_1(kb)}\right)^2 - 1\right\} - 1, \quad (26)$$

(25) is formally in agreement with equation (10). A plot of Γ_i^2 is given in Fig. 2. It will be seen that for $kb<1$ this source of noise is unimportant. (This corresponds to a beam diameter of 2 mm at 3,000 mc and 900 volts.) However, for a 6-mm beam at 1,200 volts, $kb\sim 2.6$ and the contribution to Γ^2 form this source is 0.12, which is by no means negligible.

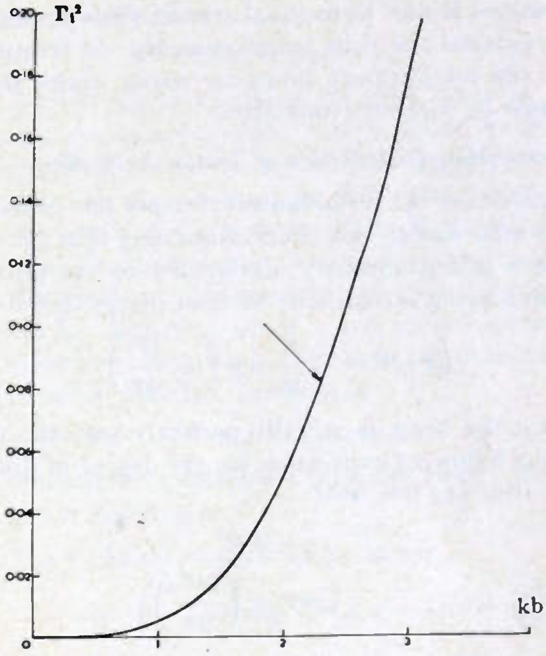

Fig. 2—Induced partition noise as a function of normalized beam radius.

This result holds for a beam symmetrically disposed about the axis, but we can easily make an estimate of the effect due to a smaller beam off the axis by taking as a model a beam occupying only a sector of an annular beam (see Fig. 3). The induced partition noise due to

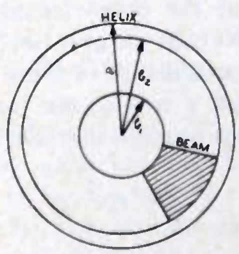

Fig. 3—Model adopted for analysis of induced partition noise in an eccentric beam.

such a beam is then obtained merely by changing the limits of integration in (23) to b_1 and b_2, and we obtain

$$\Gamma_i^2 = \Gamma_{b_2}^2 - \Gamma_{b_1}^2, \quad (27)$$

where $\Gamma_{b_2}^2$ and $\Gamma_{b_1}^2$ are the values relevant to symmetrical beams of radius b_2 and b_1. This gives a result of the same order of magnitude as for a symmetrical beam of radius b_2 and we may say that if the beam is entirely within a radius to where $kb<1$ the contribution to the noise from this source will be negligible.

It is of interest to note that a similar source of partition noise is to be expected in klystrons as a result of the variation of the beam coupling factor (β) across the gap, and that the same numerical considerations apply there.

C. Disordering of the Beam due to Transit-Time Noncorrelation

Another new type of partition noise that arises at microwaves is due to unequal transit time of electrons which entered the tube simultaneously. At frequencies where the total transit times are many cycles this effect must be carefully considered.

1. Maxwellian Distribution of Initial Velocities

MacDonald[12] in a fundamental paper has considered the disorder due to this effect. Assuming that the beam has been instantaneously accelerated to the potential V_0 after leaving the cathode with an energy distribution

$$i(V)dV = i_0 \frac{e}{k\theta} \exp\left(-\frac{eV}{k\theta}\right)$$

and that the beam is initially perfectly smooth, he derives the following expression for the degree of disorder after a time t of free drift:

$$\Gamma_\tau^2 = \frac{(\omega\tau)^2}{(\omega\tau)^2 + \left(\frac{2eV_0}{k\theta}\right)^2}. \tag{28}$$

Under conditions prevailing in twt's, this effect is negligible; if $V_0 = 500$; $\omega t = 2\pi \times 30$, then

$$\Gamma_\tau^2 = 4 \times 10^{-4}.$$

2. Velocity Spread Arising from Nonparallel Trajectories Due to Electron-Optical Defects

In any real traveling-wave tube electron trajectories are bound to be nonparallel for a number of reasons, such as misalignment of the tube in the magnetic field, nonhomogeneity of the magnetic field, and inevitable electron optical effects in the gun (see Section VIII *B*).

To estimate the magnitude of these effects, we assume that the small angle ϵ an electron makes with the axis is proportional to its original distance from the axis, i.e.,

$$\epsilon = \epsilon_0 \frac{r}{b},$$

where ϵ_0 is the angle made by electrons in the edge of the beam. The electrons are also assumed to have identical initial velocities v_0, thus the axial velocity of an electron is given by

$$v = v_0 \cos\epsilon \sim v_0(1 - \tfrac{1}{2}\epsilon^2). \tag{29}$$

In the presence of a uniform axial magnetic field it does not matter whether the electrons rotate as a whole about the axis, or whether they travel in helices around their own individual axes. We now consider an element of the beam which originally passed through the annulus $2\pi r_1 dr_1$. A time t later and a distance x along the tube the fluctuation in this element is

$$di_s = \left(2ei_0\Delta f \frac{2r_1dr_1}{b^2}\right)^{1/2} \sin\left(\omega t - \frac{x}{v}\right),$$

accompanied by a compensating fluctuation which was originally spread over the whole beam

$$di_c = -(1-\Gamma_0)\left(2ei_0\Delta f \frac{2r_1dr_1}{b^2}\right)^{1/2} \int_0^b \sin\left(\omega t - \frac{x}{v}\right)\frac{2rdr}{b^2}.$$

Evaluating this expression, we have

$$di_c = -di_s(1-\Gamma_0)\frac{2}{\theta}\sin\frac{\theta}{2}\sin\omega\left(t - \frac{x}{v_0} - \frac{x}{v_0}\frac{\epsilon_0^2}{4}\right)$$

where

$$\theta = \frac{\omega x}{v_0}\frac{\epsilon_0^2}{2}.$$

The resultant elementary fluctuation is

$$di = di_s + di_c,$$

and the total mean square fluctuation is

$$\overline{i_n^2} = \sum_{r_1} di^2.$$

When these operations are carried out we obtain

$$\overline{i_n^2} = 2ei_0\Delta f\left[1 - (1-\Gamma_0^2)\left(\frac{2}{\theta}\right)^2 \sin^2\frac{\theta}{2}\right]. \tag{30}$$

We note in passing that whatever the initial value of Γ_0^2 the beam will be completely random after $2/\epsilon_0^2$ cycles. In practice, however, we shall be interested in beams that are far from completely random. It is then permissible to write

$$\left(\frac{2}{\theta}\right)^2 \sin^2\frac{\theta}{2} \sim 1 - \frac{\theta^2}{12}$$

so that

$$\overline{i_{\text{total}}^2} = 2ei_0\Delta f\left[\Gamma_0^2 + \frac{\theta^2}{12}\right] \tag{31}$$

if we assume that $\Gamma_0^2 \ll 1$.

It would be premature to identify $\theta^2/12$ with our corresponding Γ_{eff}^2, since at the input of the twt where it matters most, θ is smaller than further on. We could split this partition noise into two parts: $\Gamma_1^2 = \theta_0^2/12$ (where θ_0^2 is the appropriate phase shift between the origin of the non-parallel electrons and the input end of the twt) and Γ_2^2 which takes account of the increase in Γ^2 as the beam passes down the tube.

In practical tubes θ_0 is small and we can ignore Γ_1^2, taking N as equal to the number of cycles electrons spend in the helix.

We now require the value of Γ^2 of the beam I_0 due to an equivalent noise current constant with distance which will give the same voltage at the twt output as the *actual* noise current which increases with distance.

It can be shown that an alternating current i constant with distance causes a voltage at the output (i.e., after N_1 cycles)

$$V_N = i\frac{Z}{2C}f_1(N_1C)$$

$f_1(N_1C)$ is a function of N_1 and C.

[12] D. K. C. MacDonald, "Transit-time deterioration of space-charge smoothing," *Phil. Mag.*, vol. 40, pp. 561–568; May, 1949.

Similarly, it can be shown that a current of the form $i' \cdot N$ causes a voltage

$$V_N' = i'N_1 \frac{Z}{4\pi C^2 N_1} f_2(N_1 C).$$

For a tube with sufficient gain

$$f_1 \approx f_2$$

and equating V_N and V_N' we have

$$i = i' \frac{1}{2\pi c}.$$

Identifying $\overline{i^2}$ with $2ei_0\Gamma_{eff}^2\Delta f$ and

$$\overline{(i')^2} \text{ with } 2ei_0\Delta f \frac{\theta^2}{12N^2}$$

we obtain finally

$$\Gamma_{eff}^2 = \frac{\epsilon_0^4}{48C^2}. \tag{32}$$

In practice C lies between 0.1 and 0.01. If $C=0.02$

$$\Gamma^2 \sim 50\epsilon_0^4.$$

Thus if Γ^2 is to be less than 0.005, ϵ_0 must be less than 0.1 radian, $\sim 5°$.

3. Depression of Beam Potential Due to its Own Space Charge

Partition noise from this cause arises in a way very similar to the one discussed in the previous section. If V_0 is the potential at the axis of the beam, the potential at a radius r can be written, in a first approximation,

$$V_r = V_0 + 9 \times 10^{11}\left(\frac{r}{b}\right)^2 \frac{I_0}{v_0}.$$

v_0 is the velocity of the axial electron. In practical units we have

$$V_r = V_0 + 1.50 \cdot 10^4 \cdot I_0 \times V_0^{-1/2} \cdot \left(\frac{r}{b}\right)^2.$$

We have assumed that the actual potential difference due to space charge is small compared with the mean beam potential axial and so we can write for the axial velocity at r

$$v_r = v_0\left[1 + 7.5 \times 10^3 I_0 V_0^{-3/2}\left(\frac{r}{b}\right)^2\right].$$

This is similar to equation (29) with the exception that now the peripheral electrons are going faster than the axial ones. It is clear that this cannot make any difference to the partition noise and we can thus use the result obtained in the previous section, replacing

$$\frac{1}{2}\epsilon_0^2 \quad \text{by} \quad 7.5 \cdot 10^3 I_0 V_0^{-3/2}.$$

This gives

$$\Gamma_{eff}^2 \approx 5 \cdot 10^6 I_0 V_0^{2-3} C^{-2}.$$

Putting $I_0 = 10^{-3}$ amps, $V_0 = 500$ volt, $C=0.02$ we obtain

$$\Gamma_{eff}^2 = 10^{-4}$$

This source of partition noise is then quite negligible at beam voltages normally used.

VII. Amplification of Thermal ("Johnson") Noise Originating in the Resistance of the Helix

The distributed loss in the helix is described by a positive attenuation coefficient α, and the "electronic" gain by a negative attenuation coefficient γ.

The net gain is then

$$G = \exp\left[-2(\alpha - \gamma)l\right]$$

where l is the active length of the tube. This we know is a gross over-simplification, but is bound to give an answer at least correct to an order of magnitude.

Further, we assume that the helix is matched at its input end into a pure resistance at temperature T_0; thus the available noise power at the input is given by KT_0B. The helix itself (or whichever part of the structure is responsible for the attenuation) is at a temperature T.

An element of length dx of the helix emits an amount of noise power in either direction of

$$dP = KT\Delta f . 2\alpha dx.$$

At the output (a distance l from the input) the total noise power arriving is

$$\begin{aligned} P_{out} &= KT_0\Delta f \exp\left[-2(\alpha - \gamma)l\right] \\ &\quad + \int_{x=0}^{x=l} dP \exp\left[-2(\alpha-\gamma)(l-x)\right] \\ &= KT_0\Delta f G\left\{1 - \frac{T}{T_0}\frac{\alpha}{\alpha-\gamma}\left(1-\frac{1}{G}\right)\right\}. \end{aligned}$$

The expression inside the brackets is the noise factor of the traveling-wave tube due to attenuation in the helix.

In any real tube G is usually very much larger than unity, so that $1/G$ can be neglected. Also $|\alpha-\gamma|$ is normally of the order of $|\alpha|$. Hence the "thermal" noise factor is approximately

$$F_T = 1 + \frac{T}{T_0}, \tag{33}$$

and if $T=T_0$, $F_T=2$ (or 3 db). It sometimes happens that the helix runs quite hot, so that if for instance $T=3T_0$, the noise factor rises to $F_T=4$, (or 6 db).

In a real twt the bulk of the noise power output comes from the shot-noise. If the noise factor due to shot noise is F_s, the over-all noise factor F is given by

$$F = F_T + F_s.$$

Suppose $F_s=40$ (or 16 db) and $F_T=2$ initially; then $F=42$ and we should observe an increase in the observed noise factor, on heating up the helix to $3T_0$, by a factor 44/42, that is, 0.2 db. This is near the limit, of what is observable, and it is of interest to note that Rogers and Chick[13] observed no increase in noise factor when heating up a helix in a twt to 650° above its normal value. These authors believe that a change of 0.5 db could have been detected.

[13] S.T.C. Ltd., Valve Laboratory Report No. G 90. July, 1949.

This, then, is not an important source of noise,—at least while noise factors due to shot noise only are well above 10.

A. Electron Wave Interaction

In the two previous sections (VI 2 and VI 3) we have considered electron beams with axial electron velocities graded across the cross section, and derived the partition noise due to the ensuing transit time disordering of primary and compensating fluctuation.

In his original paper on the electron-wave or two-beam tube, Haeff[14] also describes a tube in which a signal is amplified by the action of a beam having graded axial electron velocities, but does not give a theory of this particular effect.

However, shot noise in the beam will be amplified by means of this type of electron-wave interaction, and we shall give an estimate of this effect below.

The maximum difference between the velocity of the axial and peripheral electrons in the cases discussed in section VI 2 is

$$2\delta = v_0 \tfrac{1}{2} \epsilon_0^2,$$

using Haeff's notation. Haeff gives an expression for the rate of gain of current modulation in the two-beam tube

$$\gamma \frac{v}{\omega_1} = \pm \sqrt{\left(\frac{\delta\omega}{v\omega_1}\right)^2 + 1 \pm \sqrt{4\left(\frac{\delta\omega}{v\omega_1}\right)^2 + 1}},$$

where γ = the gain parameter, v = the mean velocity, and $\omega_1 = (e\rho/m\epsilon)^{1/2}$, the plasma frequency. When $(\delta/v) \ll (\omega_1/\omega)$, which will normally be the case in practical twt's

$$\gamma \approx \frac{\delta\omega}{v^2}.$$

We substitute for δ and have for the gain after distance z, or N cycles drift,

$$\exp(\gamma z) = \exp\left(\frac{\pi}{2}\epsilon_0^2 N\right).$$

On the assumption that the exponent is much smaller than unity, we can write for the gain

$$1 + \frac{\pi}{2}\epsilon_0^2 N.$$

Using the same argument as in section VI 2 we can derive an equivalent Γ^2 for this effect; it should be noted that it will be proportional to the original Γ_0^2, namely,

$$\Gamma_{eff}^2 = \Gamma_0^2 \frac{\epsilon_0^4}{32C^2}. \tag{34}$$

Inasmuch as Γ_0^2 is commonly much smaller than unity, Γ_{eff}^2 will be orders of magnitude smaller still, hence this effect is unimportant. The case of space-charge depression of potential is still less important, as we have seen that the velocity spread due to this effect is considerably smaller than that due to non-parallel electron trajectories.

[14] A. E. Haeff, "The electron-wave tube," Proc. I.R.E., vol. 37, pp. 4–10; January, 1949.

VIII. Experimental

A. Experiments have been carried out on a large number of tubes of varying dimensions and construction. The experimental method and the laboratory setup, however, have been the same throughout. The apparatus is shown in Fig. 4.

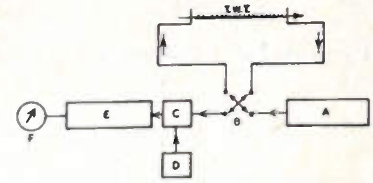

Fig. 4—Block diagram of apparatus used in measuring noise factor.

The entire traveling-wave tube with its waveguide input and output were within the field of the large solenoid which can be moved in any direction so that the magnetic field and the helix are aligned.

- A = either A_1, a pulsed signal generator with calibrated attenuator or A_2, a cw signal[15] generator, or A a coaxial line noise diode made in these labs. for this purpose;
- B = a waveguide switch[15] which makes it possible to put the twt either in or out of circuit;
- C and D are a broadband mixer[15] and local oscillator, respectively;
- E = an IF amplifier[15] 8 mc wide centered on 45 mc;
- F = an output meter connected to the linear detector of E for use in noise measurements, or else a crt for use in gain measurement.

The gain of the twt is measured by using a substitution method using the waveguide switch B, the pulsed signal generator S, and a cathode-ray tube display F. The noise factor was measured either by a signal substitution method with the cw signal generator or with the coaxial noise diode. Both these methods agreed within ½ db.

B. From the gun we desire a narrow well-defined beam of electrons moving in trajectories parallel to the axis.

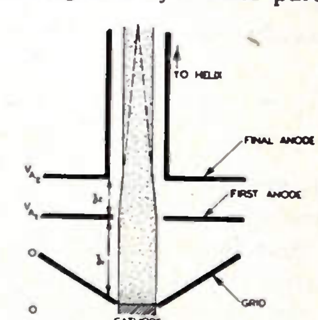

Fig. 5—Basic structure of traveling-wave tube gun.

This, of course, is an ideal, and in practice we can only approach it by degrees and by making some compromises.

A "triode" gun illustrated in Figs. 5 and 6 was

[15] We are indebted to C. Baron of R.R.D.E. for the loan of the C. W. Sig. Gen. to D. C. Rogers of S.T.C. for the idea of the waveguide switch and to C. P. Fogg and N. Houlding of T.R.E. for the loan of the mixer and IF amplifier.

adopted. The region of the gun between the cathode and the first anode A is a conventional "Pierce" gun giving a parallel beam of electrons. The gun dimensions are

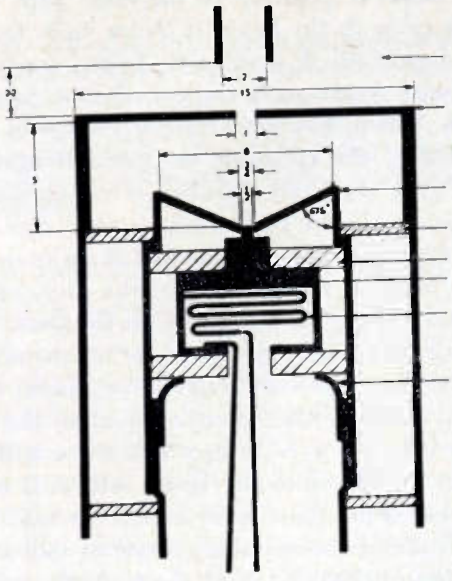

Fig. 6—Constructional details of traveling-wave-tube gun.

such that the lens at the first anode is converging and the lens at the second anode, diverging so that the beam finally emerges parallel. This is achieved when the ratio of the distance z_1 of the cathode to first anode is related to z_2 the distance from first to second anode and the ratio α of the anode potentials by

$$\frac{z_2}{z_1} = \frac{15}{4} \frac{(1-\alpha)^2}{\alpha(5-\alpha)}. \tag{35}$$

In particular, if $\alpha = 0.5$, $z_2/z_1 = 0.42$.

This gun has several advantages. Its optics vary only slowly with V_{A_1} allowing control of the beam current; and the region of the gun in which electrons travel at an angle to the axis occurs at a point of relatively high potential and is short. It therefore sets up little spiralling of the electrons.

To conclude a discussion of the optics of the tube; the magnetic field was produced by a solenoid 12 inches in diameter and 36 inches long giving a field uniform within 2 per cent over a distance of 18 inches on the axis. Keep in mind that a relatively small nonuniformity of field, say of a few per cent, extending over a small section of the twt would produce spiralling in an otherwise parallel stream of electrons such as to generate partition noise, as in section VI 2, sufficient to swamp all other sources of noise.

The earliest tubes made in which any attention was paid to the noise factor worked at approximately 1,600 volts, had helices 3 mm in diameter and were about 35 cm long, with a gain of 17–20 db, and noise figures of around 17 db. The guns in these tubes, although based on the considerations outlined in section B, were rather crudely constructed but sufficed to put 98 per cent of the current through to the collector. Guided by (4), we reduced the working voltage to 400 and the helix diameter to 2 mm. A gun using a 1-mm diameter cathode was used and gains of 25 db were easily obtained but the noise figure, instead of being reduced by the expected factor 8 times (9 db), was only reduced by 1 or 2 db, although isolated tubes had noise figures of 14 db. Tubes of similar dimensions but working at 625 volts and 850 volts were then made with identical results. Seeking for an explanation of this, we made a tube to work at 850 volts with a 2-mm helix but only a $\frac{1}{2}$-mm diameter cathode, in the expectation that if induced partition noise were the cause, this would reduce the noise. This tube has a noise figure of 12 db and we felt that this was evidence of the existence of induced partition noise in the other tubes. However, the detailed calculation of this effect (which is presented in section VI B) showed that this effect could not have been of importance in any of the tubes so far investigated. The real reason for the improvement in noise figure was that in reducing the cathode size to $\frac{1}{2}$ mm (0.020 inch) we had not only been forced to greater accuracy in the alignment of the gun parts, but also that the use of a smaller cathode makes the gun itself a much shorter structure. For a given over-all diameter, the structure approximates more closely to the ideal. As a result the beam produced by the gun consists more nearly of parallel trajectories, with a consequent reduction in noise. It was a common experience with earlier tubes that the percentage of the beam transmitted to the collector varied up and down a few per cent, as the magnetic field was increased, but this was disregarded as it was felt that as long as a value of magnetic field could be found for which 99 per cent of the current was transmitted, this effect was irrelevant. The calculation in Section VI 2 shows that of all the sources of partition noise that due to nonparallel trajectories is most important.

More detailed results for a series of 4 tubes follows:

Tube 1. (See Fig. 7): *Helix*: 2 mm inside diameter 48 T.P.I. 0.010″ tungsten wire 29.5 cm long supported in a

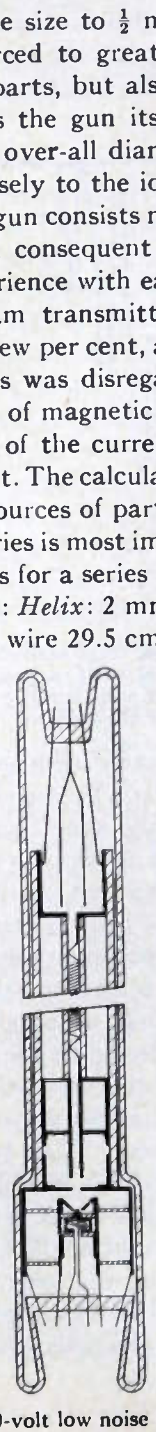

Fig. 7—Sketch of 850-volt low noise traveling-wave tube.

precision bore Pyrex glass tube. Loss, 18 db. Voltage: 850 (lower than the value 1,500 expected from the dimensions, due to the effect of the glass). $ka=1.2$; $v/c=0.058$; $N=50$; *Gun* 0.020 inch diameter cathode (see Fig. 7). $kb=0.3$. *Performance*: (i) Gain v. $i^{1/3}$ linear up to 22-db gain; (ii) Noise factor 11 db measured at a gain of 14 db with 167μa total current. 165μa collector current. (This figure obtained after subtracting 3 db for 2nd channel noise.)

The noise factor was measured at 14-db gain in order to ensure that the result was not invalidated by feedback effects. There was however no apparent change in noise factor up to 20-db gain. The percentage of the beam current collected is shown as a function of the axial magnetic field in Fig. 8 curve 1 for a total current of 167μa. It will be seen that there is a smooth rise with magnetic field and this is taken to indicate that beam consists of electrons moving very nearly parallel to the axis (see section VI 2). The measurements were taken in a field of 215 gauss.

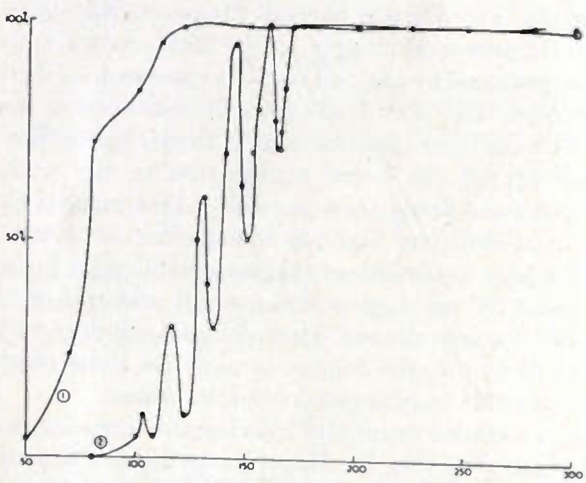

Fig. 8—Percentage of the total cathode current reaching the collector as a function of axial magnetic field. (1) For tube with good noise factor and well-aligned gun. (2) For tube with poor noise factor and badly aligned gun.

By reducing the heater volts and maintaining the current and gain constant by varying A_1 voltage, it was possible to increase the noise factor by 13 db. The current at this point did not appear to be entirely temperature limited, but it was not possible to go further as A_1 was then at the final helix potential. However this indicated a space-charge smoothing coefficient at least as small as 0.05. If (4) is used to calculate the effective Γ^2 from the results quoted, we obtain 0.013.

Tube 2 was intended to be similar to tube 1, and had a similar gain v. $i^{1/3}$ characteristic, but it was not possible to get better than 90 per cent of the total beam current to the collector. With 160μa collector current out of 180μa total current, the gain as for tube 1 was 14 db, but the noise figure was 19 db, corresponding to a Γ^2 of 0.07. If 10 per cent of the beam were being intercepted before reaching the helix, (16a) leads us to expect $\Gamma^2=0.11$, but if the interception is taking place at some part way along the helix, a value of 0.07 is not unreasonable. On opening this tube the cathode was found to be 0.015 inch off center.

Tube 3 was similar to 1, except that the construction adopted made it possible to dispense with the short drift tube prior to the helix; it had a noise factor of 12 db. However, tube 4, similar to 3, had a performance identical with 3, except in respect of noise figure, which was 17 db. When, however, the percentage of the beam transmitted to the collector was plotted against axial magnetic field, the curve 2 in Fig. 8 was obtained. Such a variation could arise if the beam consisted of spiralling electrons due to a poor gun. Then, if there is an aperture of slightly smaller size than the helix anywhere in the tube, there will be low transmission for these values of the field for which this aperture is at an antinode. As H is increased there will eventually come a point where the maximum width of the beam is equal to the aperture size; from that value of H upwards there will be complete transmission, but the beam will still consist of spiralling electrons that cause noise. We may estimate the magnitude of the spiralling angle as follows:

The radius of the orbit of an electron which is making a helical path at an angle ϵ_0 with the axis is

$$r = 3.4 \sin \epsilon_0 \frac{\sqrt{V}}{H}.$$

When the orbit of an electron on the outside of the beam (of radius b) always just misses the aperture (of radius d) we have

$$\epsilon_0 \sim \frac{(d-b)H}{3.4\sqrt{V}}.$$

Fig. 8 shows that this occurs for $H \sim 170$ gauss, approximately. If we identify b with the beam radius (0.25 mm) and put d equal to the helix radius (1 mm), then using (31). The contribution to Γ^2 from this source is 0.03. Assume that the initial Γ_0^2 of the beam is the same as in tube 1, and we have $\Gamma^2=0.013+0.04=0.053$, which corresponds to the observed noise factor of 17 db very closely. However, as Γ^2 depends on ϵ_0^4 and is therevery sensitive to what we identify with b and d, the most we can say that is that the theory gives the right order of magnitude.

IX. Conclusions

It has been shown that at 3,000 mc traveling-wave tubes with useful gain and noise figures of no more than 11 db can be made. In the course of this development some insight into the mechanism of noise generation in long beams has been gained. However, the authors feel that before any further progress can be made, a deeper understanding of the nature of the reduction of shot noise at these frequencies is needed.

X. Acknowledgments

The authors wish to thank the Chief of the Royal Naval Scientific Service for permission to publish this paper, and to acknowledge F. Ketley's able assistance in the construction of the tubes.

A Self-Balancing Microwave Power Measuring Bridge*

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Summary—The basic techniques of microwave power measurement are discussed, together with principles of self-balancing bridges for that application. A method of measuring the power change by means of a nonlinear "slave" bridge is presented. Temperature compensation problems are discussed as applied to the meter under consideration. The complete power meter is considered along with operating procedure.

I. Microwave Power Measurement

BRIDGE METHODS for microwave power measurement are popular because of their simplicity and high sensitivity. The power to be measured, or a known fraction of it, is propagated to a bolometer mount where, depending on the mount efficiency, this microwave power is dissipated in the bolometer element. A bolometer element is basically a temperature sensitive resistor whose temperature is now a function of the power dissipated in it. The change in bolometer resistance could be interpreted as an equivalent power, but for the sake of keeping the proper mount termination it is desirable to keep the bolometer resistance constant. This bolometer can be made one arm of a bridge that is initially balanced with external power in the absence of microwave energy. As microwave power changes the bolometer resistance and unbalances the bridge the external power can be removed to restore balance. The change in external power is proportional to the microwave power dissipated. The above operation can be carried out either manually or automatically by means of self-balancing bridges. Bolometer elements are of two varieties; barretters, having positive temperature coefficients of resistivity, and thermistors, having negative coefficients. Many technical articles[1–3] are available describing principles and techniques involved.

II. Self-Balancing Bridges

Alternating-current self-balancing can be accomplished by making the power-measuring bridge a feedback, or amplitude stabilizing network of an oscillator.[4] If a bridge utilizing one bolometer arm is examined as a four-terminal nonlinear network, a unique relationship exists between the output or error voltage and input or applied voltage, for a single ambient temperature.

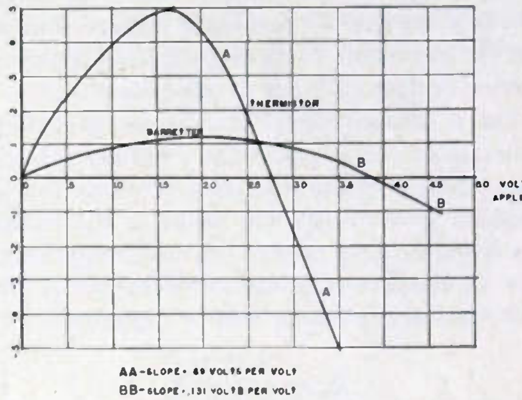

Fig. 1—Typical 200-ohm barretter and thermistor bridge characteristics. The thermistor bridge has a greater sensitivity at balance.

Fig. 1 shows two typical curves. With no applied voltage the bridge is unbalanced, but as greater powers are dissipated in the power sensitive element, its resistance changes bringing the bridge to balance. The nonlinear element must be capable of balancing the bridge at some power level. Since thermistors decrease their resistance with increasing power, balance is approached with the thermistor resistance above that of the comparison arm. The converse is true for a barretter element. If the ambient temperature rises or if external power is dissipated in the bolometer, the balance point moves closer to the origin and the slope at the balance point decreases. The slope at the balance point is referred to as the bridge sensitivity and is an important factor in improving the self-balancing action.[5] Elements having negative differential resistance regions are capable of infinite sensitivities, while any element that has only positive differential resistance regions is limited to a maximum sensitivity of 1/2, for equal arm bridges.

The bridge can be thought of as a nonlinear feedback network in which the transfer function varies with the amplitude of the applied voltage. If the bridge is made to introduce sufficient positive feedback between the output and input of a selective amplifier, then the entire circuit will oscillate and build up to an amplitude which makes the loop gain unity. If the linear amplification is "A" then the nonlinear feedback factor "β" will adjust itself so that

$$\beta = 1/A.$$

Graphically, this corresponds to superimposing a line of slope $1/A$ and passing through the origin on the bridge characteristic. The intersection of the bridge and

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[1] C. G. Montgomery, "Technique of Microwave Measurements," vol. 11, Chap. 3, McGraw-Hill Book Co., New York, N. Y.; 1947.

[2] H. J. Carlin and J. A. Blass, "Direct reading d.c. bridge for microwave power measurements, Paper 48-47, *Trans. AIEE;* 1948.

[3] E. M. Hickin, "Bolometers for vhf power measurement," *Wireless Eng.*, vol. 23, p. 308; November, 1946.

[4] L. A. Meacham, "The bridge-stabilized oscillator," Proc. I.R.E., vol. 26, pp. 1278–1294; October, 1938.

[5] G. N. Patchett, "The characteristic of lamps as applied to the non-linear bridge, used as the indicator in voltage stabilizers," *Jour. IEE.*, vol. 93, Part III, p. 305; September, 1946.

amplifier characteristics determines the stable amplitude of voltage across the bridge as well as the error voltage. There is only one feedback phase position that will result in proper operation at stable amplitudes. Operation must be at a point where an increase in amplitude decreases the feedback voltage, and tends to restore the amplitude to its original level. Barretter and thermistor bridges stabilize in opposite phase. It can be seen that a greater bridge sensitivity at balance will result in a circuit less sensitive to amplifier gain changes.

As an example of the balancing that can be expected, an amplifier gain of 400 will result in the bridge error voltage being 0.25 per cent of the bridge applied voltage, and for an equal arm bridge of resistance R ohms per arm the nonlinear element will be (Appendix I).

$$R_0 = \frac{A \mp 2}{A \pm 2} R, \tag{1}$$

where R_0 is the nonlinear resistance. If R is 200 ohms, then for a thermistor

$$R_0 = 200 \frac{400 + 2}{400 - 2} = 200\ (1.010).$$

There is generally a minimum error voltage to which a bridge can be balanced. Stray capacities may exist but can generally be balanced out. However, the error can contain harmonics of the operating frequency that are generated in the nonlinear arm. These harmonics can not be balanced out and their magnitudes depend on the thermal response of the bolometer. If the resistance is capable of partially following the instantaneous self-balancing power variations, which are of a double frequency nature, there will be a small voltage drop consisting of the double frequency resistance variation multiplying the fundamental current. The resulting sum frequency is third harmonic generated internal to the arm. The same reasoning can be extended for the case when dc is injected into the self-balancing bridge, for the error then contains a distinct second harmonic in addition to the third.

It is therefore important that the amplifier be selective and capable of rejecting the harmonic error and balancing at the fundamental frequency. Although it is best to operate at frequencies high enough so that the bolometer cannot follow self-balancing power frequency, it appears to be more practical to operate at lower frequencies. A compromise must be made based on the over-all shunt capacity of the bolometer mount. For the meter described frequencies between 10 and 15 kc have been used.

III. Direct Reading of Power

Having obtained self-balancing action for the bolometer bridge, the next problem is to measure and interpret the change in self-balancing power as external microwave power is introduced. If the initial power in the bolometer is P_0 for a particular ambient temperature, then as external microwave power is introduced (P_e) the per-unit change in voltage across, or current through, the bolometer is (Appendix II).

$$\frac{\Delta V}{V_0} = \frac{\Delta I}{I_0} = 1 - (1 - P_e/P_0)^{1/2}. \tag{2}$$

For external power up to 25 per cent of P_0, the change in voltage or current is linear with P_e according to

$$\frac{\Delta V}{V_0} = \frac{\Delta I}{I_0} = \frac{P_e}{2P_0}. \tag{2a}$$

Thus the maximum usable per-unit change is approximately 12 per cent if linearity is to be preserved. Greater sensitivity can be obtained if the variation of P_e is equal to P_0, but linearity of indication is sacrificed. In order to use this greater sensitivity condition, bias power such as dc must be injected into the self-balancing bridge. As dc is injected, P_0 decreases since it represents the maximum self-balancing component of power. It is important that the dc injected be exceedingly stable if low powers are to be measured and the amplitude stability of the self-balancing bridge is not to be impaired. This scheme has been applied to certain commercial power meters.[6,7]

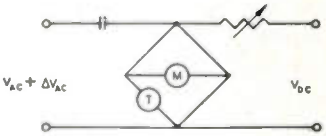

Fig. 2—The slave bridge is initially balanced with a combination of ac and dc power. A change in ac power is measured by the dc unbalance of the bridge.

Irrespective of the P_0 level the direct-reading problem is one of measuring the change in voltage across the bolometer. A differential type of voltmeter would be desirable. A nonlinear bridge operating at its balance point has been suggested[1,8,9] since a nonlinear bridge will produce an error voltage which is proportional to the change in applied voltage about the balance condition. This detection bridge can be paralleled with the self-balancing bridge. The detector bridge is referred to as a "slave" bridge. Thermistors should always be used because of their high sensitivity. Although an ac voltmeter can be used as the slave error detector, the minimum balance point previously referred to will offset the power indicator zero. The detection can be made simpler if the slave bridge is balanced partly with ac and partly with dc (Fig. 2). A dc microammeter can then be made the detector, but here again the injected dc must be

[6] Instruction Manual for 123A Wattmeter Bridge, Sperry Co., Great Neck, L. I., N. Y.

[7] Instruction Manual for 430A Microwave Power Meter, Hewlett-Packard Co. Palo Alto, Calif.

[8] C. C. Bath and H. Goldberg, "Self-balancing thermistor bridge," *Proc. Nat. Electronics Conf.*, p. 47; 1947.

[9] Suggested by E. W. Houghton, Bell Telephone Laboratories. See, E. W. Houghton, Wattmeter Circuit, Patent 2,449,072, Serial No. 737,948.

exceedingly stable. It is shown in Appendix III that the slave can be designed for maximum sensitivity. A typical maximum sensitivity is over 1.80 microamperes per millivolt. In terms of the power equation, a 200-ohm element with an initial power level equal to 10 milliwatts would have a per-unit voltage change of 0.005, if 100 microwatts of external power were introduced. The change in voltage that will appear across the slave is 10 millivolts and it will result in a detector reading of 18 microamperes. For laboratory set-ups where more sensitive detectors are available, the sensitivity can be increased.[10]

IV. Temperature Compensation

Since the bolometer is indifferent to whether its increase in temperature was a result of microwave power or of a change in ambient temperature, such temperature changes will result in a shift of the zero reference point and will give a corresponding error in reading. If a bolometer is exposed to various temperatures while in a self-balancing bridge, a relation between the power necessary to keep the element at 200 ohms (or any fixed resistance) and the temperature can be obtained. These

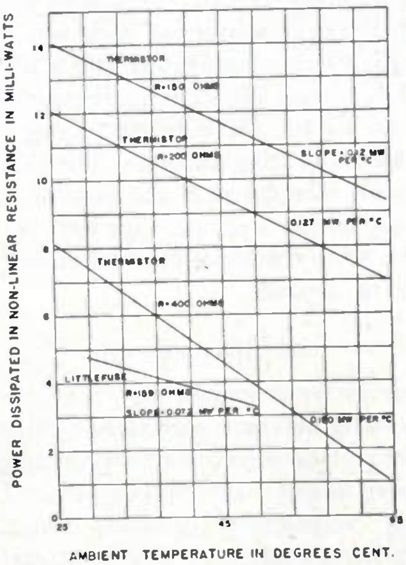

Fig. 3—Curves relating the power required to keep the bolometer at a constant resistance as ambient temperature increases are shown for various bolometers.

constant resistance characteristics can be approximated by straight lines (Fig. 3) in the room temperature region[9] and when extrapolated to zero, the intersection is the theoretical operating element temperature. For the popular thermistor and barretters used at 200 ohms, a round figure of 100 microwatts per degree centigrade can be assumed for the slope. Therefore, if ambient temperature were to change by one degree during the course of a power measurement, the resultant error would be 100 microwatts. The ultimate limit in power measurement depends a great deal on the degree of temperature compensation or stability present. For the meter under consideration in which a fraction of the self-balancing bridge power is applied to the slave bridge, compensation can be obtained if the change in power appearing across the slave, as a result of the self-balancing power changing, is exactly equal to the change in power necessary to keep the slave in balance for that

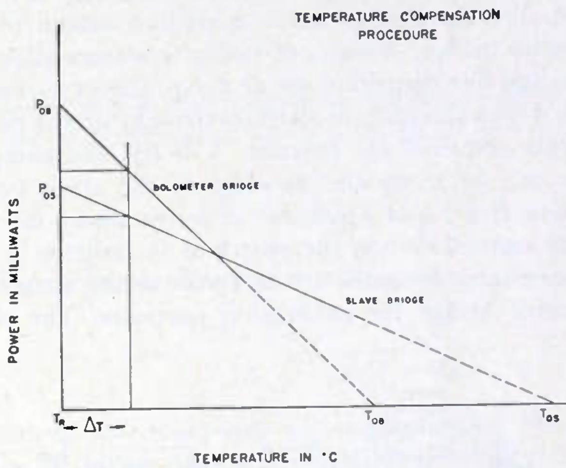

Fig. 4—For ideal compensation, a fraction of the power change in the bolometer bridge due to a temperature increment ΔT, when applied to the slave, will keep the slave in balance.

same temperature change. In Fig. 4 a constant resistance characteristic for a bolometer bridge is indicated together with a desirable slave characteristic. For a change in temperature of ΔT, there is a change of bolometer bridge power. The slave must have applied to it less of a power change in order to keep it in balance (Appendix IV). However, that was the original condition of balance in which a fraction of the a.c. power was used to balance the bridge. Experience indicates that compensation[11] can generally be obtained for moderate temperature changes. In any case, a certain degree of compensation is present for any type of bolometer bridge in parallel with the slave. Another secondary temperature effect that might be considered is the sensitivity of the slave bridge. It is seen that all non-linear bridges decrease sensitivity with increasing ambient temperature and is therefore in the proper direction to partially compensate for a reduced P_0 level.

V. The Practical Circuit

The circuit of the microwave power meter is shown in Fig. 5. Attempts were made to eliminate precision components. At present, the design is flexible since no particular bolometer was the design center.

The self-balancing bridge consists of a twin triode, parallel T tuned voltage amplifier driving a power amplifier stage which is transformer coupled to the 200 ohm bridge. The bridge error voltage is returned to the cathode of the input stage in order to take full ad-

[10] By changing the P_0 level, the sensitivity can be increased but full scale calibration becomes more complicated.

[11] Less than 50-microwatt error in a 30° C temperature change was observed.

vantage of the T selectivity. At the points X, X^1 a reversing switch controls the phase for use with either a barretter or thermistor. The bridge has been standardized at 200 ohms. There is provision for inserting dc into the self-balancing bridge for calibration purposes. When no bridge calibrating power is required, the calibrating power is returned to ground through a 200-ohm resistor and so a constant regulator tube load is maintained. The slave is driven in a similar fashion to the bolometer bridge. A series of two attenuators allow for course and fine control of the ac drive. Direct current is injected into the slave at all times to make up the deficit of power required for balance. A 0–100 microampere meter can be made the detector of the slave bridge (position B-B^1) and a resistor in series allows for sensitivity control. When the switch is in position A the microammeter becomes a 0-to-1-volt meter across the bolometer bridge for calibration purposes. The slave detector terminals are available for an external movement. The power supply is conventional and the meter shown has a double VR tube regulation arrangement.

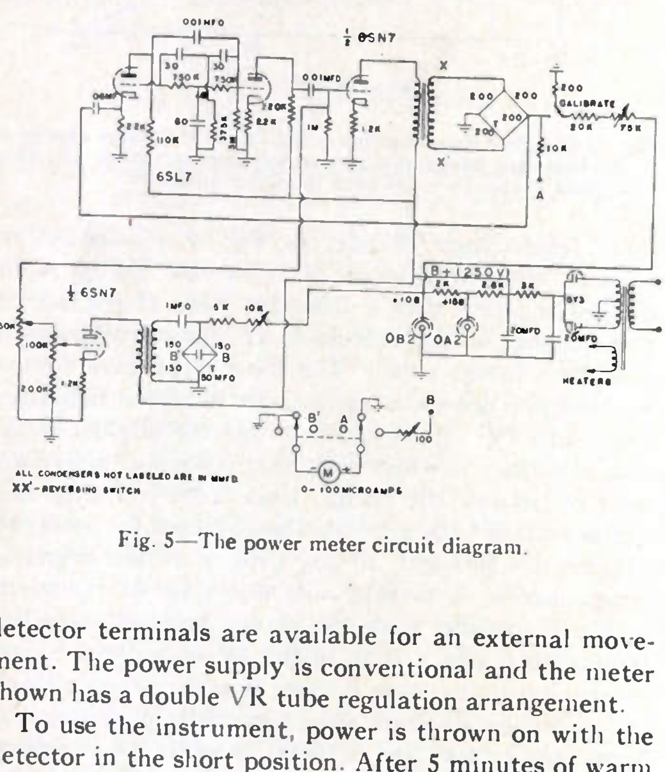

Fig. 5—The power meter circuit diagram.

To use the instrument, power is thrown on with the detector in the short position. After 5 minutes of warm up, the major drift has taken place and the switch is thrown to the measure power position. The ac and dc powers are adjusted[12] for zero indication and the switch is then thrown to the calibrate position. When the "calibrate" switch is closed, the voltage across the bolometer bridge is adjusted for a power dissipation per arm of 1 milliwatt. Going back to the power measuring position, sensitivity can be adjusted until calibrating power corresponds to full scale or any portion desired. The meter readings are linear with power (Fig. 6).

[12] As discussed previously, temperature compensation and sensitivity considerations dictate the proportion of a.c. and d.c. powers to be used.

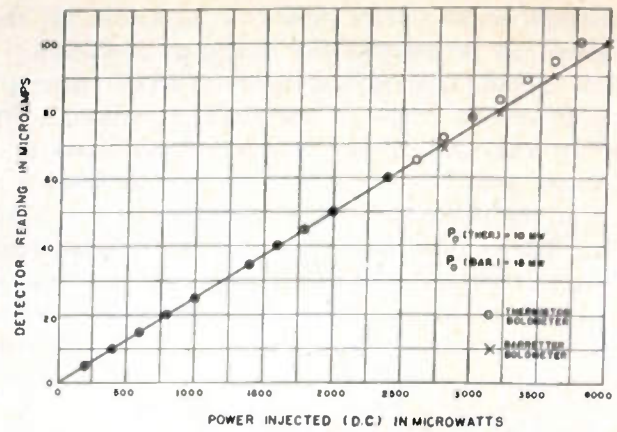

Fig. 6—The meter indication as dc power is injected into the bolometer. The linearity depends upon the bolometer's initial power level.

Zero drift can come about through drift in the regulator tube's voltage or variations in the slave bridge driver's gain. The self-balancing bridge is insensitive to voltage variation and in a test where the voltage was varied from 220 to 320 volts, the power equivalent error was 10 microwatts. The chief instability lies in the VR tube which if necessary can be replaced with a battery.

Although 500 microwatts full scale sensitivity can be obtained with 10 mw. bolometers and a 100-μa detector for laboratory setups a more sensitive movement can be connected on to the slave bridge, using the internal meter just for calibrating purposes. The calibration level setting can be eliminated if the regulator tube is assumed to stay within 1 per cent for long periods of time and only the ac to the slave need be varied if a simpler version is to be desired.

VI. Conclusions

The power meter described may require slight modification, to accommodate special bolometers. As indicated, the circuit does have certain advantages in that it reads power linearly, has no complicated calibration circuits and voltmeters, contains some degree of temperature compensation, accommodates either barretter or thermistor, and is physically small.

Acknowledgment

The authors wish to extend their thanks to T. A. Peterson, W. Kurylo, H. Zablocki, and G. Badoyannis, who assisted in the development of the instrument and principles involved, and to the United States Air Force, Watson Laboratories, who sponsored this work under contract No. AF29(099)-33.

Appendix I

If the nonlinear self-balancing bridge has adjusted itself so that

$$\frac{V_{\text{applied}}}{V_{\text{error}}} = \pm A,$$

where A is the amplifier gain and the sign accommodates the gain phase shift, then

$$\frac{R_0}{R+R_0}-\frac{1}{2}=\frac{1}{\pm A},$$

where R_0 is the nonlinear arm. Simplifying and reducing there results

$$\frac{R_0}{R}=\frac{-2\pm A}{2\pm A}=\frac{A\mp 2}{A\pm 2}. \tag{3}$$

The lower sign is taken for thermistor bridges, and the upper sign for barretters. Large values of A allow the simplification by binomial series expansion as

$$R_0=R[(1\mp 2/A)(1\mp 2/A)\cdots],$$

or

$$R_0=R[1\mp 4/A].$$

Hence a gain of 400 and a 200 ohm bridge result in the nonlinear arm balancing to ± 2 ohms or 1 per cent.

Appendix II

If a component of bolometer current due to the self-balancing arrangement is I_0, when external power is introduced (P_e) it will decrease the current by ΔI so as to keep the total power in the bolometer a constant. This can be expressed as:

$$I_0^2-(I_0-\Delta I)^2R=P_e.$$

Since RI_0^2 is the initial self-balancing portion of power (P_0) then

$$\frac{\Delta I}{I_0}=1-\left(1-\frac{P_e}{P_0}\right)^{1/2}=\frac{\Delta V}{V_0}. \tag{4}$$

By binomial expansion and using the first two terms of the series there results

$$\frac{\Delta I}{I_0}=\frac{\Delta V}{V_0}=1-1+1/2\left(\frac{P_e}{P_0}\right)\cdots\cong\frac{P_e}{2P_0}. \tag{4a}$$

Appendix III

Consider a nonlinear bridge as in Fig. 2 balanced with both ac and dc powers initially, and with a dc microammeter as a detector. A change in ac voltage, or power, will unbalance the bridge dc-wise and so measure the change in ac voltage. For an unbalanced bridge of impedance R, the current through the detector (R_M) is

$$i_d=\frac{\Delta R}{4R}\cdot\frac{V_D}{R+R_M},$$

where ΔR is the resistance change in one arm and V_D represents the dc voltage across the bridge. The voltage V_D is insufficient for balance since a voltage k is required and the remainder is made up by V_A, the ac voltage from the self-balancing bridge. Initially at balance

$$V_D^2+V_A^2=k^2.$$

If the ac voltage across the bridge changes by an amount ΔV_A, the change in ac power for each arm of the bridge is

$$\frac{2\Delta V_A\cdot V_A}{4R}=\frac{\Delta V_A\cdot V_A}{2R}.$$

Since one arm of the bridge is nonlinear to power, it will change its resistance at s ohms per watt[13] or

$$\Delta R=\frac{s\Delta V_A\cdot V_A}{2R}.$$

The resulting detector current is

$$i_d=\frac{s\Delta V_A\cdot V_A}{2R}\cdot\frac{V_D}{4R(R+R_M)}.$$

However, the dc voltage was dependent on the initial value of the ac voltage so that the detector current is

$$i_d=\frac{s\Delta V_A\cdot V_A}{2R}\cdot\frac{(k^2-V_A^2)^{1/2}}{4R(R+R_m)}.$$

The maximum sensitivity of i_d to a change in V_A occurs when the bridge is initially balanced with equal ac and dc powers. For a given percentage change in ac power, the maximum sensitivity occurs when the ac power is initially twice the dc power.

Appendix IV

Referring to Fig. 4, the constant resistance characteristic for the bolometer bridge can be represented by

$$P_B=\frac{-P_{0B}}{T_{0B}}\Delta t+P_{0B},$$

where P_{0B} is the power necessary for balance at the reference temperature T_r, and T_{0B} is the theoretical bolometer temperature. Similarly for the slave

$$P_S=\frac{-P_{0s}}{T_{0s}}\Delta t+P_{0s}.$$

Only a fraction y of the voltmeter self-balancing power appears across the slave and that power change must equal the change necessary to keep the slave in balance. Thus,

$$\frac{P_{0s}}{T_{0s}}=y\frac{P_{0B}}{T_{0B}}$$

is the condition that must be satisfied for temperature compensation with the assumed idealized characteristics.

[13] The above assumes no negative resistance effects which tend to increase the power dissipated in the nonlinear element.

Radio-Frequency Current Distributions on Aircraft Structures*

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Summary—A method for the experimental determination of the distribution of radio-frequency current on the metal surface of an aircraft mounting a transmitting antenna is described. Plots of the amplitude and phase of the current excited on various types of air frames by typical medium-high-frequency antennas are presented and compared qualitatively with the results obtained from quasi-static arguments.

Introduction

OVER A LARGE PORTION of the frequency range of greatest interest in aircraft radio systems, analysis of the characteristics of suitable airborne antennas is extremely difficult. The reason lies in the complexity of the radiating structure. Whenever the major dimensions of the air frame (wingspan, length of fuselage, and the like) are comparable with the wavelength, the airborne transmitting antenna excites a system of "parasitic" currents on the air frame which contribute significantly to both the antenna impedance and the radiation pattern. In antenna analysis, it is customary to consider the current distribution on the conducting surface as being fundamental to the study of characteristics of the antenna. The required description of the current distribution is obtained either by solution of the appropriate integral equation, or by treating the analysis as a boundary value problem. The analytical complexity of these methods usually restricts the application of the former to one-dimensional problems, and the latter to geometries appropriate to a co-ordinate system in which the wave equation is separable. Neither method is suited to the study of aircraft antenna problems in the "resonant range." This paper is devoted to the description of an experimental technique for the measurement of the current density distribution on such structures and a presentation of some of the results obtained.

Method of Measurement

Although current-density distribution measurements have been made on full-scale aircraft,[1] it is generally more convenient to employ scale models. The theory of scale models provides the assurance that the surface current density distribution (on highly conducting structures) is invariant to the reduction of all the linear dimensions of the structure by a constant factor, providing that the frequency of the driving source is increased by the same factor.[2] Models used in experiments described here were covered with aluminum or copper. Scale factors of 43 to 1 and 72 to 1 were used.

The surface current density is most readily determined from an exploration of the magnetic field intensity at the surface of the conductor, since

$$[\hat{n}, \vec{H}] = -\vec{J}$$

where $\vec{H}$ is the magnetic vector, $\vec{J}$ is the surface current density vector, and $\hat{n}$ is the unit vector normal to the surface. The required magnetic field data are obtained with an exploring loop. The amplitude of the loop pickup is proportional to the amplitude of the current density. The time phase of the loop output is uniquely related to the phase of the surface current, and the direction of current flow is determined by the intersection of the conductor surface with the plane of the loop when the loop is oriented for maximum response.

In practice, a number of precautions must be observed. To preserve accuracy, the loop must be made as small as is practical. True electrical balance must be achieved for the loop and its load to avoid spurious responses. Care must be taken to ensure that the spacing between the loop and the surface of the antenna under test is kept constant, and as small as practicable. In these experiments a shielded circular loop, approximately 1 centimeter in diameter, was used. It was constructed of rigid coaxial line with an outer diameter of approximately 1 millimeter, and was mounted in a polystyrene frame for convenience in manipulation and as an aid in maintaining constant spacing to the antenna conductor.

This construction made it possible to achieve a high degree of balance (a 60-db difference in output between maximum orientation and minimum orientation) and a very good rejection of spurious modes. (For equal incident field intensities, pickup in the electric dipole mode was 90 db below maximum pickup in the loop mode.)

The output signal was detected by a calibrated crystal detector mounted in a shielded cartridge in the probe handle, and brought to the auxiliary measuring

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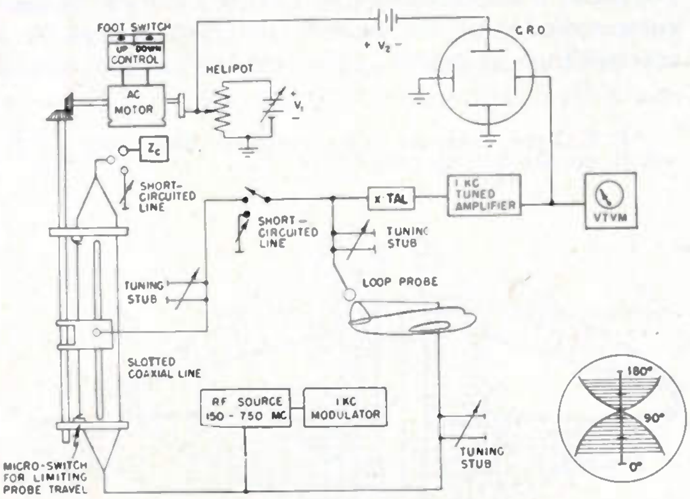
[1] K. Pippel and H. Baerner, "Strom- und Spannungs-Verteilung auf Hochfrequent Schwingenden Flugzeugen," Forschungsbericht Nr. 915 Flugfunk-Forschungsinstitute, Oberpfaffenhofen, Germany; February, 1938. USAF translation No. 511, Air Technical Service Command.

[2] G. Sinclair, "Theory of models of electromagnetic systems," Proc. I.R.E., vol. 36, pp. 1364–1369; November, 1948.

apparatus on a suitable length of RG-59/U coaxial cable. The aircraft model under test was supported on a pile of Styrofoam blocks in an open area. The probe was operated manually, and a large number of tests demonstrated that the proximity of the operator had a negligible effect on the density distributions studied.

Relative phase measurements were made by mixing the radio-frequency output of the loop probe with the output of a reference source in a crystal detector. The reference signal was obtained from a movable probe on a terminated coaxial line driven by the same generator as the antenna under test. Relative phase could thus be obtained from the position of the probe on the reference line when it had been so positioned that the detector output was a minimum. For operating convenience certain refinements were added. These are illustrated by the block diagram of the measuring setup (see Fig. 1). The position of the reference probe was determined by a motor drive which was operated by foot switches. A multiturn potentiometer was geared to the probe



(a) (b)

Fig. 1—Functional diagram of the setup used in measuring surface current amplitude and phase using scale models. (a) Block diagram. (b) Cathode-ray oscilloscope presentation of phase.

carriage and biased by an adjustable dc source (V_1 in Fig. 1) in such a way that the voltage on the movable arm was directly related to the probe position. This dc signal supplied the vertical deflection for a cathode-ray oscillograph with a screen calibrated from 0° to 180° along the vertical axis. Adjustment of a series bias voltage (V_2 in Fig. 1) together with V_1 permits setting the 0° and 180° positions of the cathode-ray-tube beam at the appropriate positions of the reference probe (a half wavelength apart). The amplified output of the crystal mixer is applied to the horizontal deflection plates of the cathode-ray tube. The cathode-ray tube pattern which results from motion of the reference probe through a half wavelength is indicated in the lower right corner of the figure. Phase angles in the third and fourth quadrants are determined by the same method after rotating the loop probe 180°.

Results

In interpreting the measured data which follow, certain fundamental notions regarding current density distributions should be kept in mind. First, it may be recalled that the surface current density on a long, isolated circular cylinder, whose diameter is small compared with the operating wavelength, is proportional to the total current and is inversely proportional to the diameter of the conductor. Therefore, a given observed surface current density on a circular cylinder of moderate diameter implies a correspondingly greater *total* current than if the same surface density were observed on a conductor of much smaller diameter. Second, if the cross section of a long, isolated cylindrical conductor is not circular, the surface current density around the periphery of the cross section will not be uniform, but will tend to be greatest where the radius of curvature is smallest. Lettowsky[3] has shown that for such cases, and assuming a high conductivity and high frequency and that the maximum cross-sectional dimension is small compared to the wavelength, the surface current density distribution around the cross section is very nearly the same as would be the distribution of static charge on the same conductor. This fact can be used to calculate the current density distributions appropriate to particular cross sections. Fig. 2 shows the density distribution calculated on this basis for a conductor of

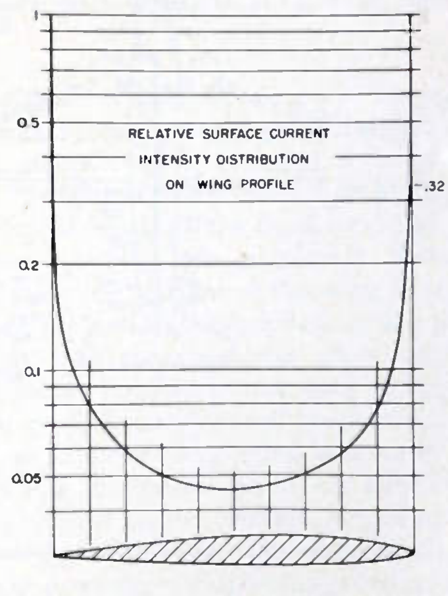


Fig. 2—Current density distribution on a cylinder of air-foil cross section calculated from quasi-static formulas.

air-foil cross section. It is apparent that the current density at the leading and trailing edges of the wing will be much greater than the average density at any particular cross section.

The problem of representing the measured data on a two-dimensional diagram is not a simple one. Two

[3] F. Lettowsky, "Skineffekt in Zylindrichen Leitern mit elliptischem Querschnitt bei hohen Frequenzen," *Arch. Elektrotech.*, vol. 35, p. 643; 1941.

methods have been employed in presenting the data which follow. These have been referred to as "peripheral" and "contour" plots. Where the cross-sectional dimensions of the air-frame components are small compared to the longitudinal dimensions and to the wavelength, the following assumptions are reasonable ones: The current flow is parallel to the major axes of the air-frame components around the periphery of any cross section, the time phase of the current is constant, and the amplitude distribution of the current density is the same as in the quasi-static case, as discussed above. Under these conditions, all of the important properties of the surface current distribution as well as the amplitude and phase of the equivalent axial total currents are derivable from amplitude and phase data taken at one reference point on each cross section. For convenience, the reference line is taken as the leading or trailing edge of the wings and control surfaces, and at the top or bottom, or along the sides, of the fuselage. Plots of such data are referred to in this paper as "peripheral" plots. The "contour" plots yield considerably more detailed information. In these an appropriate number of the current flow lines are traced out in their entirety. Contours of constant surface current amplitude and of constant time phase are plotted to complete the picture. Such plots permit a closer study of the "fine structure" of the current density distribution and help in evaluating validity of assumptions implicit in the use of "peripheral" plots.

Fig. 3 shows a peripheral plot of the current distribution on a C-47 (DC-3) aircraft at a full-scale frequency of 6 Mc. At this frequency the inclined wire is $0.35\ \lambda$ long, while the length of the fuselage is about $0.55\ \lambda$. The data show that the current on the inclined wire is approximately sinusoidally distributed, and differs only slightly in phase along its length. This is in accordance with the usual behavior for a thin wire antenna. The current on the fuselage is quite different. The current density has an approximately uniform amplitude distribution around the periphery of the fuselage cross section, but exhibits a marked phase shift as the point of observation progresses from the top to the bottom of the fuselage. The authors have no adequate explanation of this effect. A clue may lie in the results of Carter's analysis of the current induced in a circular cylinder by an incident plane wave.[4] Carter shows that the phase of the surface current density induced in a circular cylinder by an incident plane wave lags by a marked angle at points on the "shadow" side. Although

[4] P. S. Carter, "Antenna arrays around cylinders," Proc. I.R.E., vol. 31, pp. 671–692; December, 1943.

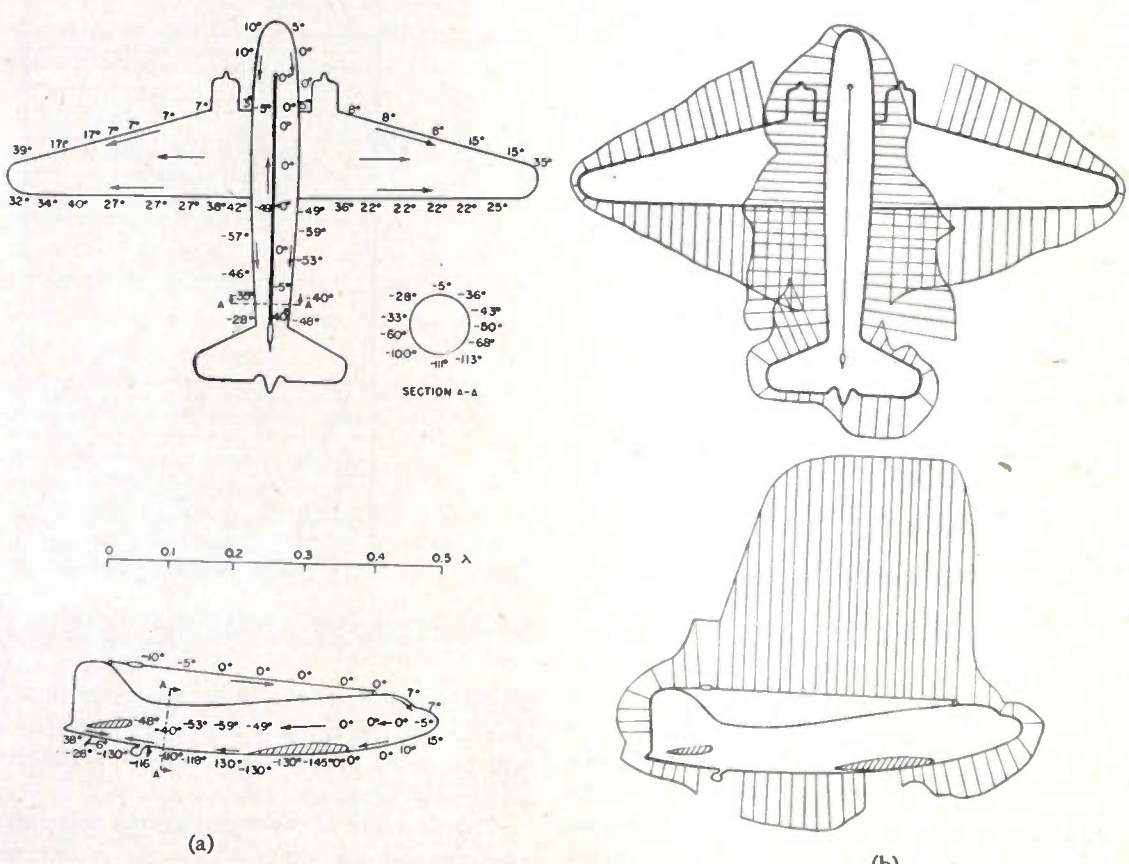
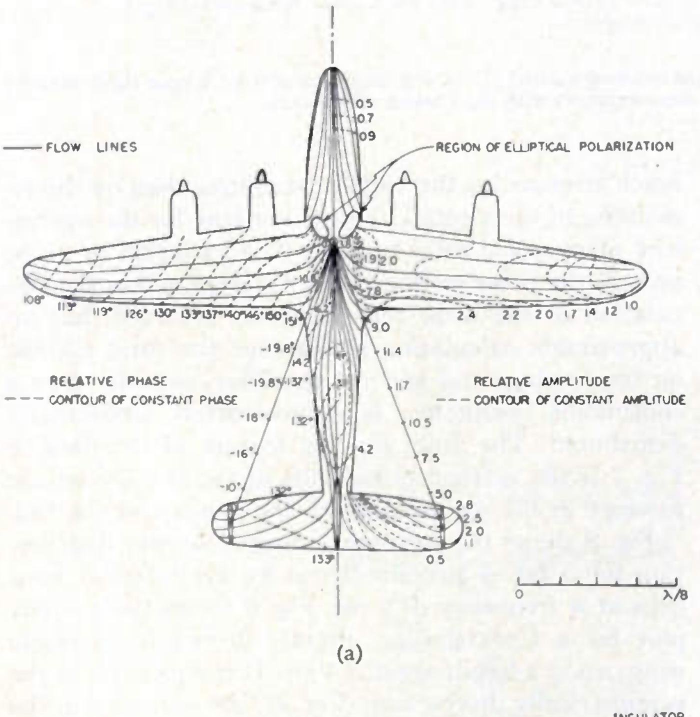


Fig. 3—Peripheral surface current amplitude and phase data measured on a scale model of a C-47 aircraft excited by a fixed inclined-wire antenna at a full-scale frequency of 6 Mc. The lengths of the lines drawn perpendicular to the periphery of the air frame indicate the relative amplitude of the surface current density at those points. The arrows show the direction of the current at a particular instant of the radio-frequency cycle. (a) Phase distribution. (b) Amplitude distribution.

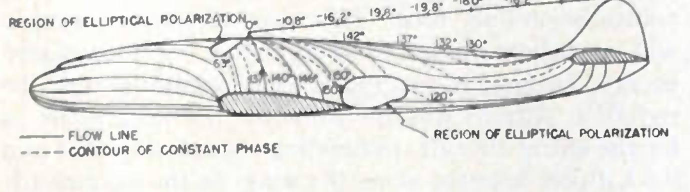
his results apply to cylinders with a larger electrical diameter than the fuselage in this case, the close proximity of the driven wire may serve to accentuate the phase-lag effect he describes. If an equivalent filamentary total current, assumed to be flowing along the axis of the fuselage in the opposite direction from the current on the wire, is calculated from the density measurements at section A-A, it is found to have an amplitude of 0.72 times that of the antenna current and a phase of $-60°$. In calculating the equivalent filamentary total current, the error in surface density data due to the use of an exploring loop of finite size must be taken into account. A derivation of the expression for the correction factor is given in Appendix A.

Figs. 4 and 5 show the contour plots of surface current distribution on a Constellation aircraft excited by an inclined-V antenna at a full-scale frequency of 7 mc. At this frequency the wing span is just equal to a half wavelength. Since the system is both electrically and mechanically symmetrical about the center line of the fuselage only one half of contours are shown in Figs.

A study of Figs. 4 and 5 reveals several interesting points concerning the current distribution. From the upper parts of Figs. 4 and 5 it is seen that on the wings,

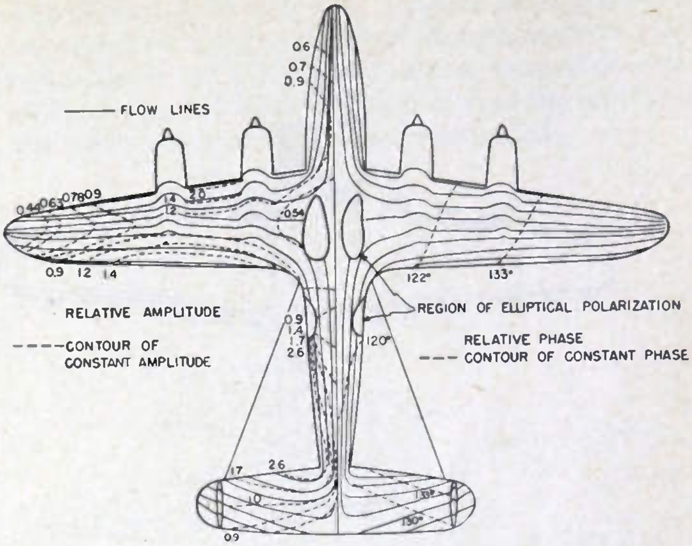


(a)

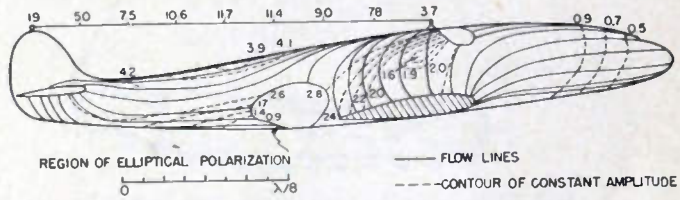


(b)

Fig. 4—Contour plots of the surface current distributions on a Constellation aircraft excited by a fixed inclined-V antenna at a full-scale frequency of 7 mc. (a) Top view. (b) Side view.



(a)



(b)

Fig. 5—Other views of data shown in Fig. 4. (a) Bottom view. (b) Side view.

where the chord is small compared to the span and to the wavelength, the current flow is nearly parallel to the wing axis, except near the junction of the wing and the fuselage. The shape of the constant amplitude contours on the wing shows that the cross-sectional amplitude distribution resembles the quasi-static distribution of Fig. 2 on the outer portion of the wing. Near the wing root and on the horizontal stabilizer, where the chord is comparable with the span, the quasi-static approximations are not justified by the data of Figs. 4 and 5. The quasi-static assumptions of longitudinal flow, constant phase at any cross section and static amplitude distribution at any cross section, are nowhere applicable to the current on the fuselage, according to Figs. 4 and 5. Another feature of these data which should be noted is the presence of elliptically polarized current on certain portions of the structure. The contours surrounding the "regions of elliptical polarization" shown in Figs. 4 and 5 are those on which the ellipticity is 2 to 1. Near these regions, then, the assumption of "stationary lines of flow" which is implied in the usual applications of the integral-equation method of antenna analysis is not adequate.[5]

Fig. 6 shows a comparison of the peripheral distributions for an asymmetrical shunt-fed wing structure on a

[5] J. Aharoni, "Antennae," Clarendon Press, Oxford, England, pp. 127 ff.; 1946.

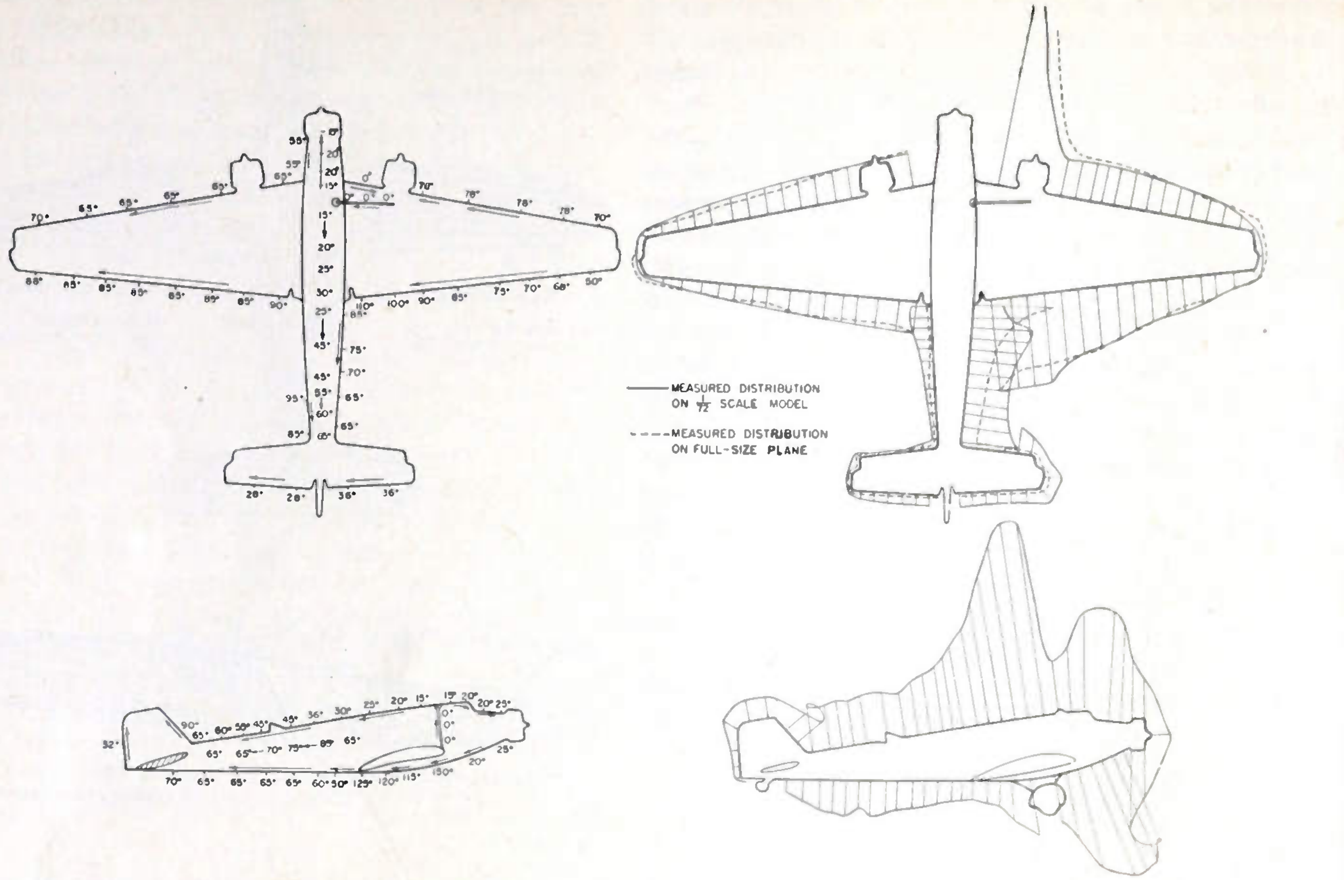


Fig. 6—Peripheral plots of surface current density distribution on a shunt-fed wing excited JU-52 at a frequency of 6 mc. Upper right drawing shows a comparison of the results of model measurements with data taken at full scale.

JU-52 aircraft at a full-scale frequency of 6 mc. The solid curves refer to data on a 1/72-scale model obtained by the method described here, while the dashed curves refer to data obtained by Pippel and Baerner[1] on a full-scale aircraft. The discrepancies between the two sets of data can be adequately explained on the basis of the difference in the size of the exploring loops employed in the two experiments.

Fig. 7 shows the peripheral current density distribution for a DC-6 aircraft excited by a tail-cap antenna at a frequency of 6 mc. In studying these results, it should be noted that the apparent reversal in phase of current flowing onto the upper surface of the fuselage from the vertical stabilizer is not an actual phase reversal. Rather, it results from the convention used in defining phase. Obviously, the phase of a current depends on which direction of flow is taken to be positive. In a system in which currents are flowing in 3 mutually perpendicular directions, it is logical to define the positive directions of current flow as being in the directions of the positive axes of, say, a right-handed Cartesian co-ordinate system. This has been done for the data given in this paper. In Fig. 7 the arrow indicating the positive direction of current flow is directed downward, hence the apparent reversal in phase of what is in fact a continuous current flow at the junction of fuselage and tail.

In Fig. 7 it is evident that the current density is much stronger on the vertical stabilizer than on the remainder of the aircraft. This is not true for the equivalent filamentary total current. The variation in phase around the cross section of the fuselage makes the calculation of the total current rather involved, but an approximate calculation shows that the total current on the fuselage and vertical stabilizer, considered as a continuous conductor, is approximately sinusoidally distributed. The most striking feature of the data of Fig. 7 is the marked phase shift of the current on the fuselage in the region between the wings and the tail.

Fig. 8 shows the peripheral current density distribution for a DC-4 aircraft driven by symmetrical wing caps at a frequency of 6 mc. Fig. 9 shows the contour plot for a Constellation aircraft driven by a single wing cap at a frequency of 3.1 mc. It is seen that, in the symmetrically driven case (Fig. 8), the currents on the two sides of the fuselage are of the antisymmetric, or transmission-line, form. This type of current system will contribute to stored energy, but not to radiated energy. In fact, it was found experimentally that the radiation pattern was the same for the wing alone as for the entire aircraft, resembling very closely that of a 0.6 λ dipole oriented along the wing. In the asymmetrically driven case (Fig. 9), the current divides between the wing and fuselage at their junction in a manner that depends on the shape and dimensions of the air frame. In general, both current systems contribute significantly to the far-zone field. If the fuselage and wing currents are in time phase, as was the case for the data shown in Fig. 9, the resulting radiation pattern is that of a dipole with its axis in the horizontal plane but skewed with respect to the wing axis. A simple consideration of the resolution of the total current vectors shows that the null to the left of the aircraft is shifted toward the nose while that on the right is shifted toward the tail. This conclusion is supported by experimental data.

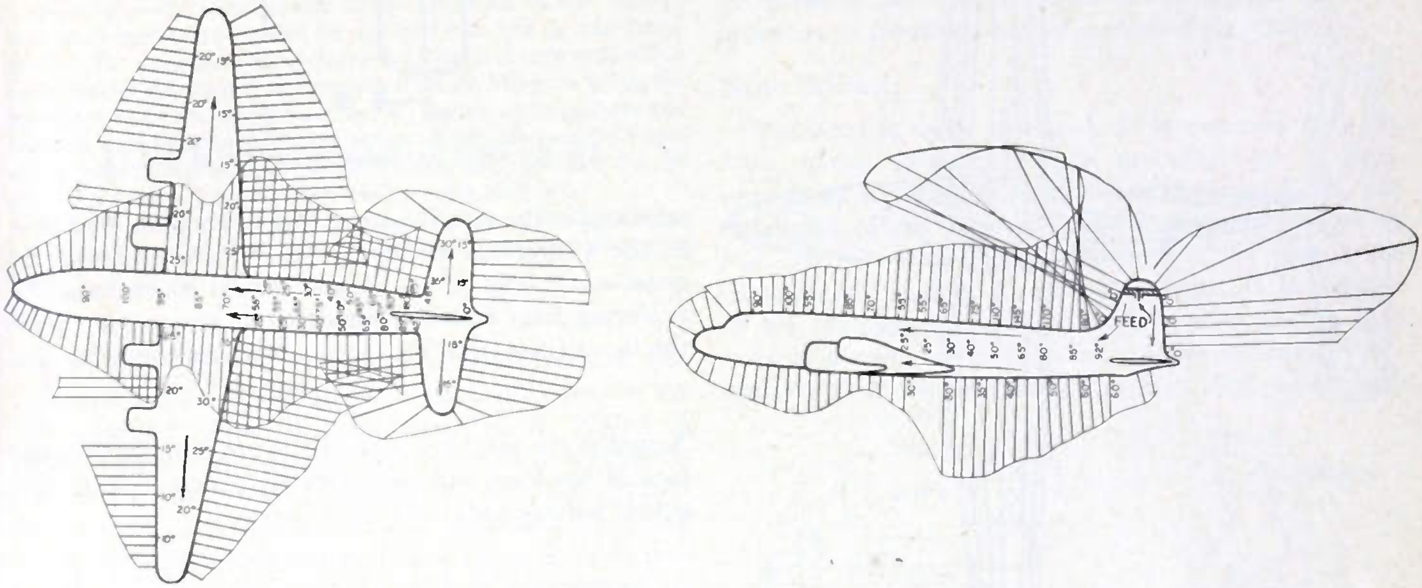


Fig. 7—Peripheral plots of surface current density distribution on a DC-6 aircraft excited by a tail-cap antenna at a full-scale frequency of 6 mc.

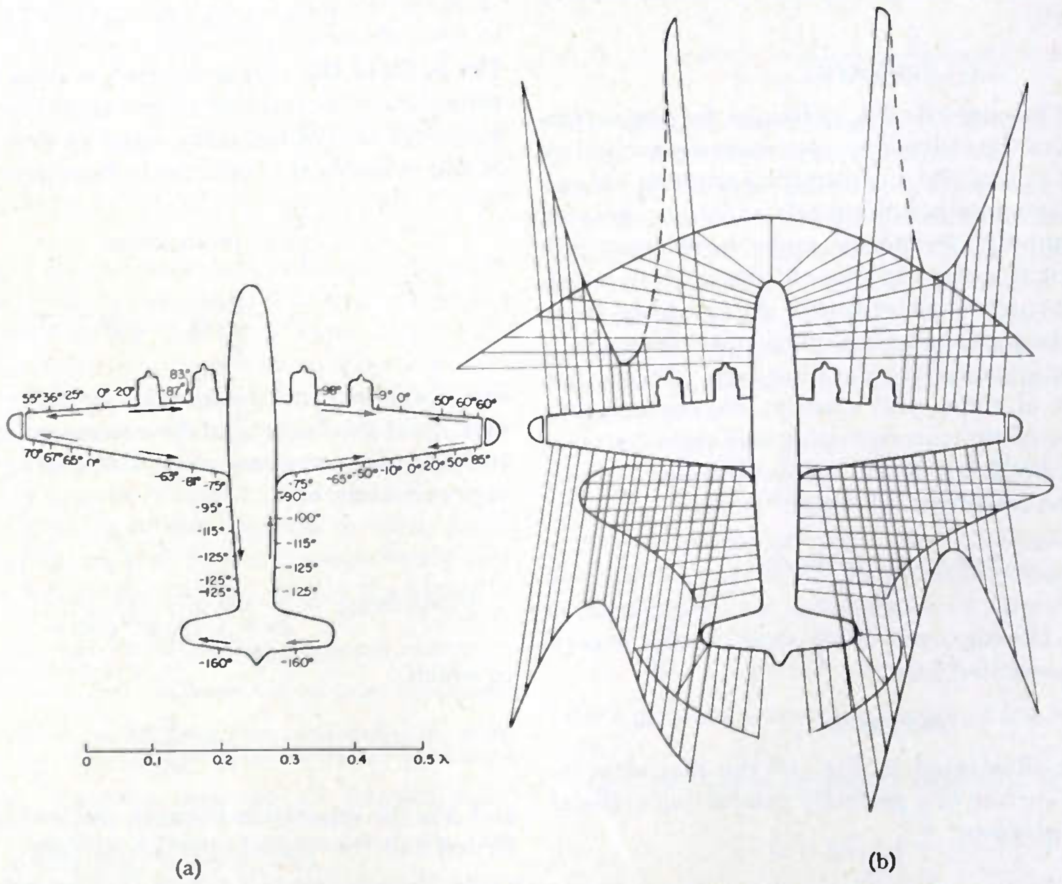


Fig. 8—Peripheral plots of surface current density distribution on a DC-4 aircraft excited by symmetrically located wing caps at a full-scale frequency of 6 mc. (a) Phase measurement. (b) Amplitude measurement.

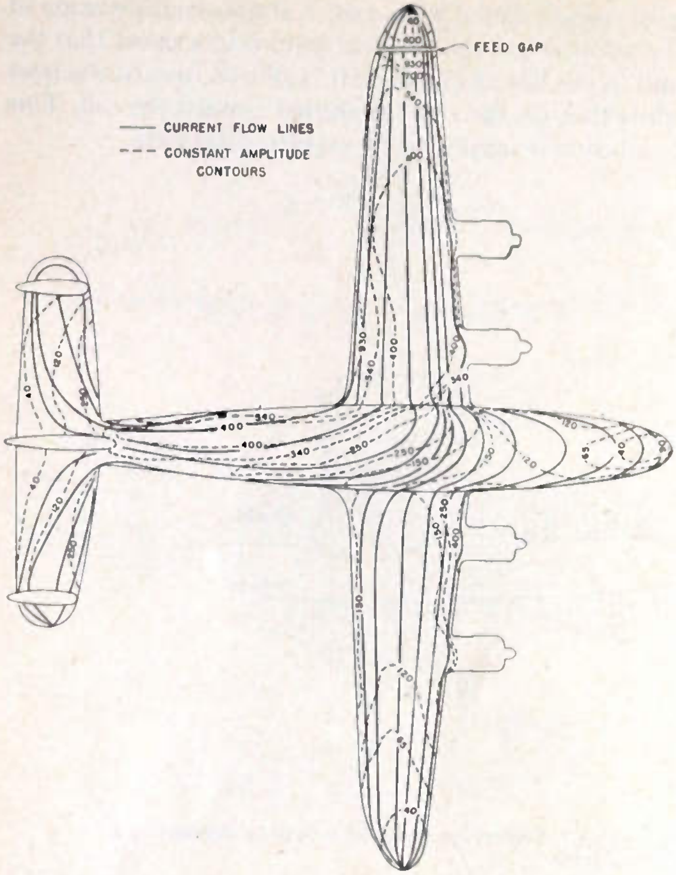


Fig. 9—Contour plot of surface current distribution on a Constellation aircraft excited by a single wing cap at a full-scale frequency of 3.1 mc.

Summary

This paper has described a technique for the experimental study of the radio-frequency currents excited on the skin of an aircraft by a transmiting antenna. Measured results for a number of aircraft antennas operated in the medium-high-frequency range have been presented. Except to point out some of the ways in which the observed current distributions depart from those which would be predicted from quasi-static arguments, no attempt to interpret the data has been offered. The interpretation of these, and similar, current distributions in terms of the corresponding radiation patterns and input impedances is an involved subject which must await the conclusion of further work.

Acknowledgment

The authors wish to acknowledge the contribution of Mrs. Robert Dressler, who took most of the experimental data presented here.

Appendix

In the case illustrated in Fig. 10 the magnetic intensity at the surface of a perfectly conducting cylinder of radius a is given by

$$H = \frac{I_t}{2\pi a}$$

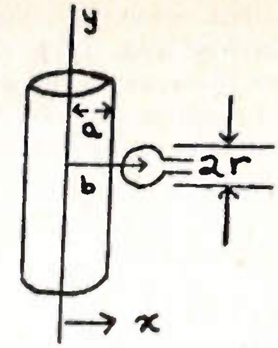


Fig. 10—Perfectly conducting cylinder of radius a.

where I_t is the total current, which is assumed to flow in the y direction and to be distributed according to quasi-static laws. If the magnetic field which links the exploring loop had this intensity at every point over the face of the loop, the induced voltage would be

$$V_1 = -j\omega\mu H\pi r^2.$$

Actually, the magnetic intensity is not uniform over the face of the loop, but decreases with increasing x. The actual voltage induced in the loop is

$$V_2 = -j\frac{\omega\mu I_t}{\pi}\int_{(b-r)}^{(b+r)} \frac{\sqrt{(r^2-b^2)+2bx-x^2}\,dx}{x}.$$

The integration yields

$$V_2 = -j\omega\mu I_t(b - \sqrt{b^2-r^2}).$$

The ratio of the actual induced voltage to the voltage which would be induced in the same loop by a uniform magnetic field of intensity equal to that at the surface of the cylindrical conductor is therefore

$$\frac{V_2}{V_1} = \text{probe factor}$$

$$= \frac{2a}{r^2}(b - \sqrt{b^2-r^2}).$$

For a symmetrical Joukowsky airfoil[6] of chord l and maximum thickness t_{max}, the corresponding "probe factor" for measurements along the leading edge is given approximately by

$$\text{Probe factor} = \frac{\left(2+\frac{1}{\epsilon}\right)r}{2r + \sqrt{l}(\sqrt{l\epsilon^2+\Delta+2r} - \sqrt{l\epsilon^2+\Delta})}$$

in which

$$\epsilon = \frac{4t_{max}}{3l\sqrt{3}}$$

and Δ is the separation between the leading edge of the air foil and the nearest point on the loop.

[6] L. A. Pipes, "Applied Mathematics for Engineers and Physicists," McGraw-Hill Book Co., New York, N. Y., p. 515; 1946.

Transfer Properties of Single- and Coupled-Circuit Stages With and Without Feedback*

JOACHIM W. MUEHLNER†, MEMBER, IRE

Summary—Universal curves of the reciprocal of the complex response function are studied for systems with one or two tuned circuits. The normalized parabola for two circuits is presented with a net of loci for the origin of co-ordinates within it. The ratio Q_1/Q_2, the amount of fixed detuning, the amount of coupling, and amplitude and phase of any additional feedback independently displace the origin with respect to its position in case of two identical, uncoupled circuits without feedback. Complex cases may be easily analyzed.

Different methods incorporating regeneration or feedback loops are discussed which allow practical bandwidth control within wide limits.

I. Introduction

To REPRESENT the transfer admittance of two coupled tuned circuits by a parabolic locus in the complex plane, was proposed by H. Kafka in 1934.[1] The author subsequently published a comprehensive analysis of tuned-circuit filters with and without feedback loops.[2] The normalized transfer admittance loci for one, two, and three circuits had then been presented in their most general form. Other publications on the subject followed, more recently those by Chang,[3] and Harrison and Mather.[4]

The author feels that the subject should be presented once more. If variables and constants are properly normalized, a very simple and comprehensive view of both coupled- and uncoupled-circuit behavior is obtained which greatly facilitates a quick and clear analysis of the general case. Contributing to this fact in the case of two circuits is a new presentation of the normalized parabola which differs slightly from former proposals.

It is suggested that a change of circuit data does not result in a move of the parabola with respect to the origin of co-ordinates, but vice versa. Thus, a net of co-ordinates for the origin may be drawn within the parabola (Fig. 3), one family for constant ratio Q_1/Q_2, the other one for a constant amount of fixed detuning. Any additional shift of the origin due to circuit coupling and/or feedback may be readily incorporated.

The reciprocal of the system response function E_i/E_o will be considered, E_i and E_o being, respectively. the complex input and output voltage. It is in the nature of the presentation of E_i/E_o in the complex plane that the influence of feedback can be conveniently studied.

Basic Relations

Reference is made to Fig. 1. It is assumed that the filter under investigation is preceded by a high-impedance amplifier tube. The complex ratio E_g/E_o, the reciprocal of the internal amplification, is plotted in Fig. 1(b). In other words, Fig. 1(b) is the plot of the complex grid voltage, E_g, for unity output. If, in the case of an additional feedback circuit, part of the grid voltage is supplied from the output, the feedback vector appears as a constant vector in Fig. 1(b), with phase

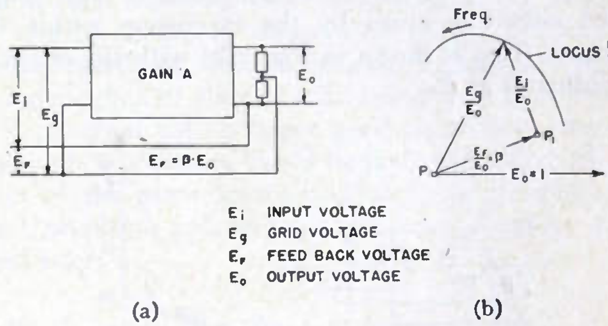


Fig. 1—(a) Feedback amplifier. (b) Reciprocal of the amplification, locus in the complex plane.

and magnitude depending on the special feedback conditions. With feedback, the origin of co-ordinates with respect to the external amplification is at P_1. The locus itself and its frequency scale is independent of feedback conditions. The influence of feedback phase and amplitude on the system response function is obvious from the position of the origin P_1 with respect to the locus and can be studied conveniently. Self-oscillation will occur if, by properly choosing β, the origin is made part of the locus.

II. One Tuned Circuit With and Without Feedback

The reciprocal of the system response function for the circuit shown in Fig. 2a *without* feedback is given by:

$$\frac{E_i}{E_o} = -\frac{1}{g_m \cdot Q}\sqrt{\frac{C}{L}}\left[1 + jQ\left(\frac{\omega}{\omega_0} - \frac{\omega_0}{\omega}\right)\right], \quad (1)$$

where g_m is the transconductance of the filter-preceding tube.

The normalized frequency deviation Ω is defined by:

$$\Omega = Q(\omega)\cdot\left(\frac{\omega}{\omega_0} - \frac{\omega_0}{\omega}\right) \quad (2)$$

Ω varies approximately linearly with the frequency

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[1] H. Kafka, "Ein Beitrag zur Theorie der zweikreisigen Bandfilter fuer Zwischenfrequenzstufen," *Hochfrequenztechnik and El. Akustik*, vol. 44, p. 125; 1934.

[2] J. Muehlner, "Bandfilter ohne und mit Rueckkopplung," *Hochfrequenztechnik and El. Akustik*, vol. 54, pp. 80–93; 1939.

[3] S. Chang, "Parabolic loci for two tuned coupled circuits, Proc. I.R.E., vol. 36, pp. 1384–1388; November, 1948.

[4] A. E. Harrison and N. W. Mather, "Graphical analysis of tuned coupled circuits," Proc. I.R.E., vol. 37, pp. 1016–1021; September, 1949.

deviation from the resonance frequency ω_0. For $Q \geqq 100$ Q may be considered frequency independent within the frequency band of conventional interest and $(\omega/\omega_0)-(\omega_0/\omega)$ may be substituted by 2δ. In this case we obtain:

$$\Omega \cong 2Q \cdot \delta. \tag{3}$$

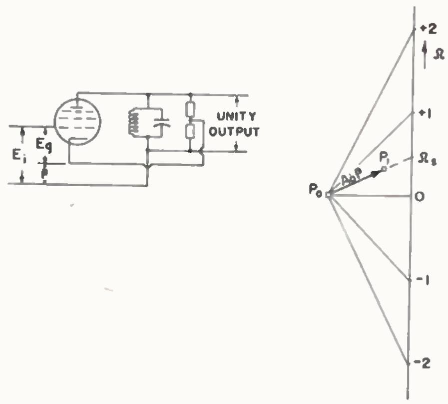
Since $-(1/g_mQ)\sqrt{C/L}$ represents the reciprocal of the amplification A_0 at frequency ω_0 (identical to $\Omega=0$), the following substitution may be made:

$$-\frac{1}{g_mQ}\sqrt{\frac{C}{L}} = \frac{1}{A_0}.$$

With this substitution and with (2) (1), (4) may be written:

$$\frac{E_i}{E_o} = \frac{E_g}{E_o} = \frac{1}{A_0}[1 + j\Omega]. \tag{4}$$

If (4) is plotted in the complex plane, the normalized straight-line locus which is applicable to any single, tuned circuit is given by the expression within the brackets. This is shown in Fig. 2(b) with the origin of co-ordinates at P_0.



(a) (b)

Fig. 2—(a) One-circuit stage with voltage feedback. (b) Normalized straight-line locus with feedback vector.

With feedback as shown in Fig. 2(a), a constant fraction β of E_o, which in the general case is complex, is added to E_i to form E_g. Fig. 1(b) shows that:

$$\beta + \frac{E_i}{E_o} = \frac{E_g}{E_o}. \tag{5}$$

Thus (4) changes into

$$\frac{E_i}{E_o} = \frac{1}{A_0}[1 + j\Omega - A_0\beta]. \tag{6}$$

The effect of the feedback is obvious from Fig. 2(b). The origin of co-ordinates located at P_0 without feedback shifts to P_1. The real component of the feedback vector, $\beta_r \cdot A_0$, affects filter selectivity and amplitude whereas the imaginary component, $\beta_i \cdot A_0$, changes the resonance frequency of the overall system by $\Delta\Omega = \beta_i \cdot A_0$. Increasing feedback voltage of like phase would result in oscillations of frequency Ω_s. If the feedback voltage is shifted by 90 degrees with regard to E_o, only detuning results and the tube acts in this case as a reactance.

III. Two Tuned Circuits Without Coupling, Without and With Feedback

Without Feedback

For an amplifier with two single-circuit stages, the system response function E_i/E_o is obtained by multiplying the single-stage functions ((1) with (3)).

$$\frac{E_i}{E_o} = \frac{1}{g_{m_1}g_{m_2}Q_1Q_2}\sqrt{\frac{C_1C_2}{L_1L_2}}[(1 + 2jQ_1\delta_1)(1 + 2jQ_2\delta_2)]. \tag{7}$$

Q_1 and Q_2 may be referred to a quantity Q of arbitrary value.

$$Q/Q_1 = a; \qquad Q/Q_2 = b. \tag{8}$$

Furthermore, we substitute

$$2Q\delta_1 = \Omega_1(\omega); \qquad 2Q\delta_2 = \Omega_2(\omega)$$

$$-\frac{1}{g_{m_1}Q_1}\sqrt{\frac{C_1}{L_1}} = \frac{1}{A_{10}}; \qquad -\frac{1}{g_{m_2}Q_2}\sqrt{\frac{C_2}{L_2}} = \frac{1}{A_{20}}. \tag{9}$$

With (8) and (9), (7) may be changed into:

$$\frac{E_i}{E_o} = \frac{1}{a \cdot b \cdot A_{10} \cdot A_{20}}[(a + j\Omega_1)(b + j\Omega_2)]. \tag{10}$$

In the special case where

$$a = 1, \quad b = 1, \quad Q_1 = Q_2 = Q$$

and

$$\Omega_1 =. \Omega_2 = \Omega, \qquad A_{10} = A_{20} = A_0,$$

(10) is reduced to

$$\frac{E_i}{E_o} = \frac{1}{A_0^2}[1 + j\Omega]^2. \tag{11}$$

Equation (11) corresponds to (4) for only one circuit.

The bracket term of (11) represents the normalized parabolic locus in the complex plane as it may be obtained by squaring all vectors of the straight-line locus (Fig. 2(b)). Reference is made to Fig. 3.

In the case of two equal circuits as defined for (11), the origin of co-ordinates is located at P_0 in unity distance from the parabola vertex. It will be shown that by allowing a complex displacement of the origin of co-ordinates from P_0, any case in which two circuits are involved, coupled or uncoupled, with or without feedback, can be presented in the same, normalized parabola.

Assuming that in the general case of two separate tuned circuits as given by (10), the origin of co-ordinates is displaced from P_0 by the vector $\Lambda = \Lambda_r + j\Lambda_i$, Λ may be found by comparing the bracket term of (10) to the one of (11):

$$(a + j\Omega_1)(b + j\Omega_2) = (1 + j\Omega)^2 - \Lambda. \tag{12}$$

With

$$\Omega(\omega) = \frac{\Omega_1 + \Omega_2}{2}$$

and

$$\Omega_1 = \Omega - \Delta\Omega$$

$$\Omega_2 = \Omega + \Delta\Omega \tag{13}$$

the components of Λ are obtained from (12):

$$\Lambda_r = 1 - a \cdot b - \Delta\Omega^2 \tag{14}$$

$$\Lambda_i = \Delta\Omega(b - a) \tag{15}$$

$$\text{if} \quad a + b = 2 \quad \text{or} \quad Q = \frac{2Q_1Q_2}{Q_1 + Q_2} \text{ is chosen.} \tag{16}$$

In the general case of two circuits of different natural frequency and Q, (14) and (15) give the displacement of the origin along the parabola axis and perpendicular to it. Reference is made to Fig. 3.

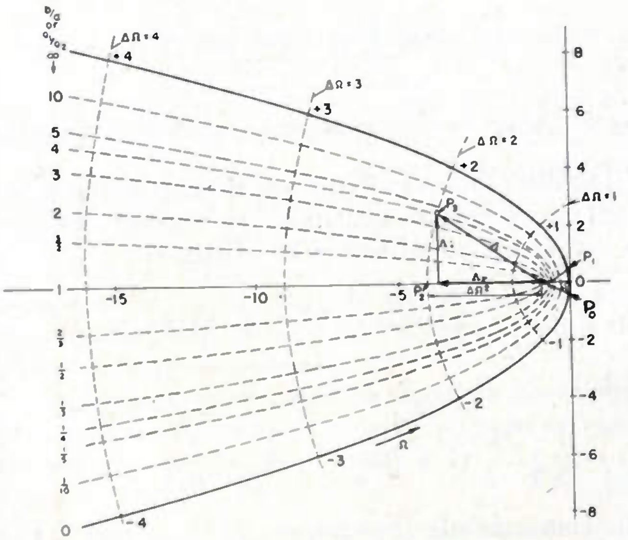


Fig. 3—Normalized parabolic locus for two circuits with net of co-ordinates for origin.

The origin of co-ordinates can quickly be found in any given case if a normalized-parabola chart such as shown in Fig. 3 has been prepared. The locus proper is the solid outer parabola with its frequency scale in Ω, where Ω according to (3) and (16) is:

$$\Omega(\omega) \cong \delta \cdot \frac{4Q_1Q_2}{Q_1 + Q_2}. \tag{17}$$

The dashed lines within the main parabola are the parabolic loci for the origin of co-ordinates.

For example:

Circuit 1: Resonant frequency $f_{10} = 1006.5$ kc, $Q_1 = 400$.

Circuit 2: Resonant frequency $f_{20} = 994$ kc, $Q_2 = 100$.

Equation (17) yields the frequency scale for this special case:

$$\Omega = 320\delta.$$

The following equation corresponds to (17) and yields

$$\Delta\Omega = \frac{2Q_1Q_2}{Q_1 + Q_2} \cdot \frac{f_{10} - f_{20}}{f_{\text{med}}}. \tag{18}$$

For the above example $\Delta\Omega = 2$ is obtained. Furthermore

$$\frac{Q_1}{Q_2} = \frac{b}{a} = 4.$$

As shown in Fig. 3, we now find the origin in the special case where the dashed parabola for $Q_1/Q_2 = 4$ crosses the perpendicular parabola with parameter $\Delta\Omega = 2$ (origin P_3). The predominant influence of circuit 1 with its higher Q results in an asymmetrical response function, the highest peak of amplification being close to $\Omega = +2$.

For practical use, a larger parabola with a greater density of co-ordinates would be preferable. The application of the same normalized parabola to coupled-circuit problems and to feedback circuits will be discussed later.

Special Cases with Two Single Circuits

If the tuning of the circuits is held constant but the ratio Q_1/Q_2 is changed, the origin will move along the proper curve of the family $\Delta\Omega = $ constant. If Q_1/Q_2 is held constant but the circuits are variably detuned, the origin will move along the proper parabola of the family $Q_1/Q_2 = $ constant.

With respect to P_0, the origin will move toward the vertex in the case of equal tuning but increasingly different Q. For the filter used in this example, but with both circuits tuned to the same frequency, P_1 would be obtained.

The origin will move in the opposite direction, away from P_0, if the Q's are kept equal and the circuits become more and more detuned. P_2 would be the origin in the case of the above circuits if these had equal Q's.

All cases $Q_1 \neq Q_2$ with $\Delta\Omega \neq 0$ result in asymmetrical response functions. Any position of the origin within the parabola is possible.

With Feedback

Whatever the origin within the normalized locus may be, for a given frequency the vector between it and the corresponding point of the locus represents the complex grid voltage which is required at the filter-preceding tube to produce unity output. Part of this voltage may be supplied from the output. Corresponding to (6) for the single-circuit stage, we obtain in case of two single circuits with feedback:

$$\frac{E_i}{E_o} = \frac{1}{A_0}\left[(1 + j\Omega)^2 - A - A_0 \cdot \beta\right], \qquad (19)$$

with $A_0 = a \cdot b_o \cdot A_{10} \cdot A_{20}$ = amplification for midfrequency. Equation (19) shows that the origin in the parabola depends on phase and magnitude of the feedback vector. By properly choosing A_0B, the origin may be shifted from one position to any other position desired.

Bandwidth Control

The normalized bandwidth B of a filter is defined by

$$B = 2 \cdot |\Omega_l| = 2\frac{Q}{f_0} \cdot b, \qquad (20)$$

with $\pm\Omega_l$ being the frequencies for which the response of the filter is 3 db lower than the maximum value and b being the bandwidth in cps. "Maximum bandwidth" is the bandwidth of a double-hump filter where the response at the peaks is 2 db higher than at the midfrequency. This is obtained with the origin of coordinates four units to the left of P_0 in the parabola.

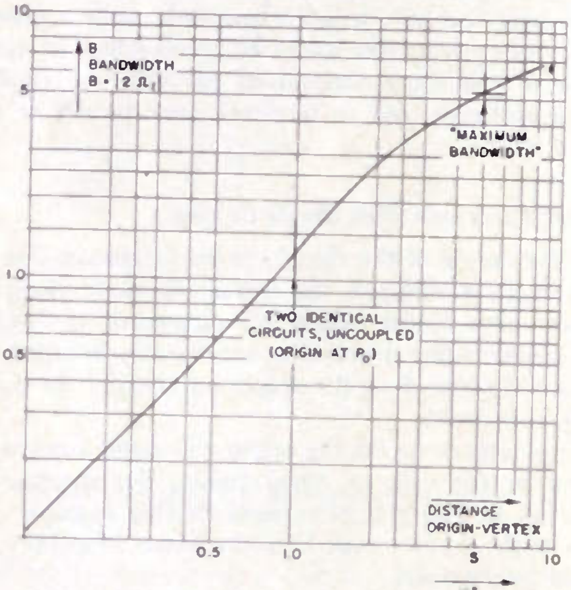


Fig. 4—Normalized bandwidth versus position of the origin within normalized parabola.

Two staggered-tuned stages may be adjusted to "maximum bandwidth" with no, or very small, feedback. The part of the output voltage that is fed back without phase shift may be variable. The position of the origin within the parabola which is left of P_0 in the wide-band position may thus be shifted towards the parabola vertex. The larger the amount of feedback, the more the circuit detuning is balanced. By shifting the origin beyond P_0 towards the vertex, very small bandwidths may be produced which would otherwise not be possible with the same circuits. Fig. 4 shows the normalized bandwidth B versus position of the origin on the parabola axis. The bandwidth B is given approximately by the number of units between origin and vertex. Obviously, any decrease in bandwidth due to feedback is connected with increase of amplitude at midfrequency.

IV. Two Coupled Circuits With and Without Feedback

A system of coupled circuits has natural frequencies which are not the resonance frequencies of its single circuits. However, it is possible to replace any coupled filter by successive uncoupled circuits if these are tuned to the natural frequencies of the coupled system and possess the same Q's as the coupled system for the respective natural resonance frequencies. The normalized parabolic locus, as shown in Fig. 3 is, therefore, also applicable in cases of two coupled circuits.

Practically identical response functions are obtained for any two coupled circuits regardless of the coupling mechanism chosen, if high-Q circuits are conventionally coupled.[5] The investigation will be restricted to such cases. With $Q \geq 100$, Ω may then be considered proportional to the frequency deviation from the midfrequency according to (3). With the coupling coefficient $k \leq 0.02$, k^2 can be neglected with respect to unity.

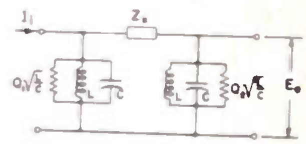


Fig. 5—Voltage coupled tuned circuits considered in the analysis.

The filter shown in Fig. 5 was chosen because it lends itself to convenient calculations. Its transfer admittance as that of a pi-structure can be written as:

$$\frac{I_1}{E_o} = Y_1 + Y_2 + Z_k \cdot Y_1 \cdot Y_2 \qquad (21)$$

with

$$Y_1 = \frac{1}{Q_1}\sqrt{\frac{C}{L}}\,(1 + j\Omega_1); \qquad Y_2 = \frac{1}{Q_2}\sqrt{\frac{C}{L}}\,(1 + j\Omega_2).$$

The characteristic impedances $\sqrt{L/C}$ of both circuits are equal. Furthermore, the coupling impedance, Z_k, may be a loss-free capacitance, C_k.

To normalize the coupling we define

$$K = k \cdot Q \qquad (22)$$

with k = coupling coefficient. Here $k = C_k/C$ and Q according to (16).

Because of the loading effect of the coupling capacity, the midfrequency of the filter response curve is lower than the medium frequency between the two natural circuit frequencies. This will be accounted for by the proper substitutions (23) because it is desirable to have $\Omega = 0$

[5] The influence of low circuit Q on the transfer admittance of two coupled circuits is illustrated in Fig. 3 of footnote reference 4. For $Q \geq 100$, the parabolic locus for two coupled circuits is accurate within 1 per cent for a frequency deviation, which corresponds to the 3-db points of the response curve.

refer to the band midfrequency, regardless of the type of coupling chosen.

$$\Omega(\omega) = \frac{\Omega_1 + \Omega_2}{2} + K$$
$$\Omega_1 = \Omega - K - \Delta\Omega \tag{23}$$
$$\Omega_2 = \Omega - K + \Delta\Omega.$$

An amplifier with transconductance g_m may precede the coupled-circuit filter and the following substitutions made:

$$\frac{Q}{Q_1} = a; \qquad \frac{Q}{Q_2} = b; \qquad \frac{1}{g_m Q}\sqrt{\frac{C}{L}} = \frac{1}{A_0},$$

where A_0 = amplification for a stage with one circuit of like characteristic impedance and Q at the resonance frequency.

With the above substitutions, (21) yields after transformations:

$$\frac{E_i}{E_o} = \frac{1}{jA_0K}\left[(a+j(\Omega-\Delta\Omega))(b+j(\Omega+\Delta\Omega))+K^2\right]. \tag{24}$$

By using the transformation of (12) with (13), we obtain:

$$\frac{E_i}{E_o} = \frac{1}{jA_0K}\left[(1+j\Omega)^2 - (\Lambda - K^2).\right] \tag{25}$$

The term in the bracket of (25) shows the expression for the normalized parabola $(1+j\Omega)^2$ (origin P_0), the displacement Λ determined by the ratio Q_1/Q_2 and $\Delta\Omega$, and, additionally, the effect of the circuit coupling: K^2.

The frequency-independent factor which determines the amplitude of the response function depends on the amount of coupling and is imaginary. Because the 90° phase difference between E_i and E_o is thus taken into consideration in the bracket-preceding factor, the position of the parabola in the complex plane remains the same as in the previous cases.

The total displacement D of the origin with respect to a filter with equally tuned, uncoupled circuits of equal Q results as:

$$D = \Lambda - K^2. \tag{26}$$

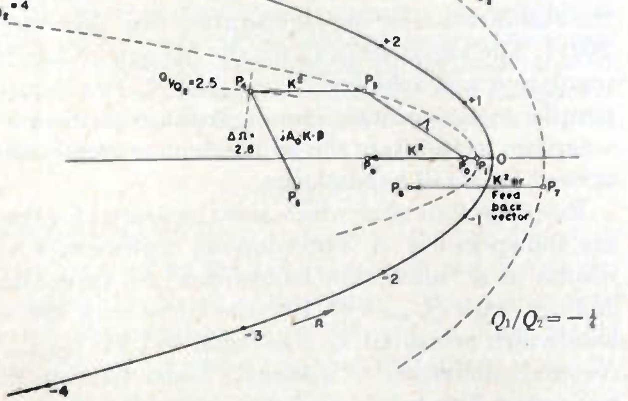


Fig. 6—Position of origin of co-ordinates in the most general case of detuned, coupled circuits of unequal Q, with feedback.

In (26) Λ may be complex. It depends on circuit tuning and Q ratio and is found with (14) and (15). (See Section III.) K^2 is always real if the coupling link is a reactance.[6]

Reference is made to Fig. 6. P_3 is the origin of co-ordinates for the uncoupled-circuit filter described as an example in Section III. ($Q_1/Q_2 = 4$, $\Delta\Omega = 2$.)

Let us now assume that the same circuits are additionally coupled without otherwise affecting the circuit data. Assume the coupling coefficient to be $k = C_k/C = 0.0125$. Equation (22) yields the normalized coupling coefficient $K = 2$. According to (26) the origin for the coupled filter is, therefore, at P_4, four units left of P_3. If the co-ordinates of P_4 are read it, becomes evident that the same response function would be obtained from an uncoupled-circuit filter with $Q_1/Q_2 \cong 2.5$ and $\Delta\Omega \cong 2.8$.

With the same circuits tuned to the same frequency, the origin would be at P_1 without coupling, and at P_6 for a normalized coupling factor $K = 2$. The influence of the ratio $Q_1/Q_2 = 4$ would be small in this case. No asymmetry would result and the system response function would be the same as for a filter with two equal circuits (Q given by (16)) and a slightly different amount of coupling. Generally, for circuits of unequal Q, equally tuned, we obtain:

$$D = 1 - a\cdot b - K^2. \tag{27}$$

The origin between P_0 and the parabola vertex without coupling shifts along the parabola axis away from the vertex if coupling is increased.

For circuits of equal Q, detuned, we obtain:

$$D = -\Delta\Omega^2 - K^2. \tag{28}$$

Both detuning and coupling shift the origin away from the vertex. A coupling of K corresponds to a detuning by $\Delta\Omega$.

With Feedback

If part of the output voltage across the secondary condensor is fed back and added to the grid voltage of the tube preceding the filter, the feedback vector which is independent of frequency appears in (25).

$$\frac{E_i}{E_o} = \frac{1}{jA_0\cdot K}\left[(1+j\Omega)^2 - (\Lambda - K^2) - jA_0K\cdot\beta\right]. \tag{29}$$

Referring to Fig. 6, the origin of co-ordinates for the coupled filter under discussion is P_4 without feedback. The feedback vector $jA_0K\cdot\beta$ shifts the origin to a different, arbitrary position, P_5, which depends on phase and amplitude as given by the complex ratio β and the amplification.

Origin P_7 is not stable because it is outside the parabolic locus proper. The locus for P_7 is determined by $Q_1/Q_2 = -\frac{1}{4}$. An amplifier with two single-circuit stages or with a very loosely coupled-circuit filter would oscillate if one of the circuits had a negative Q. However, stable conditions are possible by means of feedback or

[6] To find the order of magnitude for the asymmetry to be expected in the case of a coupling link with losses, shift the origin within the parabola by $K^2e^{(\pm j\delta)}$ with δ being the loss angle of the coupling link ($\tan\delta < 0.05$).

by a certain amount of coupling between the circuits. Thus, P_8 or any other stable point within the parabola may be obtained. This is important if a response function of extremely small bandwidth but of the desirable double-peak shape is called for. If, for example, the secondary circuit of a coupled filter is provided with a regenerative feedback, its individual Q may be raised to infinite or beyond to negative values. The effective Q of the whole filter may then become very high in accordance with (16) but still be positive. The origin can be within the parabola four units of P_0 if the coupling is properly adjusted.

Feedback circuits can be used to advantage in filter design if regeneration is stabilized by simultaneous negative feedback. (See circuit shown in Fig. 7.)

Bandwidth Control

Also, in case of coupled-circuit, filters an effective control can be obtained by feeding part of the output voltage back to the filter input? Using this principle,[7] a practical automatic bandwidth control may be designed as follows: By the proper amount of coupling with no or very small feedback, the desired "maximum bandwidth" is obtained. Part of the filter output voltage is varied proportionally to neighboring-channel interference, shifted by 90 degrees and added to the filter input. With such feedback properly dimensioned, the origin will automatically assume the optimum position between the point four units to the left of P_0 and the parabola vertex.

Another type of bandwidth control with coupled circuits seems worth mentioning because of its unusual features. Reference is made to Fig. 7.

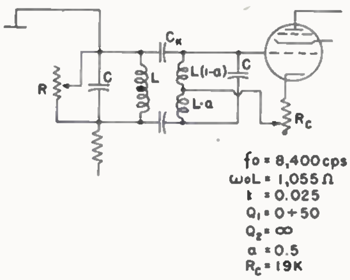


Fig. 7—Wide-range bandwidth regulation with constant amplitude at the band midfrequency.

By stabilized regeneration, the Q of the secondary circuit has been raised to infinity. The circuits are tuned to the same frequency. In this case the reciprocal of the amplification for the midfrequency is obtained from (25), as:

$$\left|\frac{E_i}{E_o}\right|_{\Omega=0} = \left|\frac{1}{jA_0K}[K^2]\right| = \frac{1}{g_m}\cdot k\cdot\sqrt{\frac{C}{L}}. \qquad (30)$$

It is remarkable that the amplification at the midfrequency does not depend on the Q of the primary circuit, Q_1. A change of Q_1, however, affects the amplitudes at all frequencies except the band midfrequency and hence results in a bandwidth regulation with constant amplitude for the midfrequency. For $Q_2=\infty$, (27) yields:

$$D = 1 - K^2 \quad \text{for} \quad Q_1 > 0, \qquad (31)$$

and from (16) and (22)

$$K = 2Q_1\cdot k \qquad (32)$$

is obtained.

According to (31), the distance between origin and parabola vertex is K^2. The origin's position varies with the square of Q_1. $K^2=5$ with $Q_{1,max}=(1/k)(\sqrt{5}/2)$ results in "maximum bandwidth."

Reference is made to Fig. 4, abscissa scale K^2.

From Fig. 4 the normalized bandwidth B is obtained. After determining B the bandwidth b (in cps) is obtained from (20) with $Q=2Q_1$. Note that the bandwidth B is measured in units which are proportional to the variable Q_1. Approximately and with good accuracy for small bandwidths

$$B \cong K^2.$$

Hence, with (20) and (32)

$$4Q_1\cdot\frac{b}{f_0} \cong (2Q_1k)^2$$

$$b \cong Q_1k^2f_0[\text{cps}] \quad \text{with} \quad Q_1 = \frac{R}{\omega_0L}. \qquad (33)$$

Response functions were measured for a filter at 8,400 cps with circuit Fig. 7. The secondary circuit is provided with a stabilized regeneration. In the practical case of considerable stabilization by high values R_c, the condition for infinite Q_2 is given by:

$$\frac{R_p}{R_c}a(1-a) = 1$$

with $$R_p = Q_{20}\sqrt{\frac{L}{C}}$$

$$Q_{20} = \text{secondary } Q \text{ without regeneration.} \qquad (34)$$

In the case of the measured filter with $a=0.5$ and $R_p=75{,}000$ ohms, (34) yields $R_c\cong 19{,}000$ ohms. The adjustment of the regeneration was done by varying R_c to the critical value of approximately 19,000 ohms for which, with the primary circuit short-circuited, oscillations of the secondary circuit started. Since the tube transconductance was in the order of 1/100[1/ohm], the stabilization of the regeneration was approximately 200:1. The variation of Q_1 for the purpose of bandwidth regulation was achieved by varying R. The rheostat R, parallel to the primary circuit, had a logarithmic characteristic to facilitate the adjustment to small values as needed for small bandwidths.

Response functions which were measured for the filter are shown in Fig. 8. The coupling coefficient $k=0.025$ results in a "maximum bandwidth" of approximately 233 cps, with $Q_{1\,max}=45$. For the curve with the widest bandwidth measured, Q_1 was equal to 1.04·"$Q_{1\,max}$" and resulted, therefore, in a slightly wider bandwidth and more than 2 db between peaks and saddle. The smallest bandwidth measured was obtained with $R=100$ ohms

[7] U.S.A. Patent number 2,033,330

and resulted in a measured bandwidth of 1.2 cps $\pm$ 1 cps. This curve still had good stability.

It will be noted that the amplitude for the band midfrequency is rather constant though no other adjustment than that of varying R had been performed during the measuring period. This fact, as given by (30) is due to the following: If the primary circuit of the filter

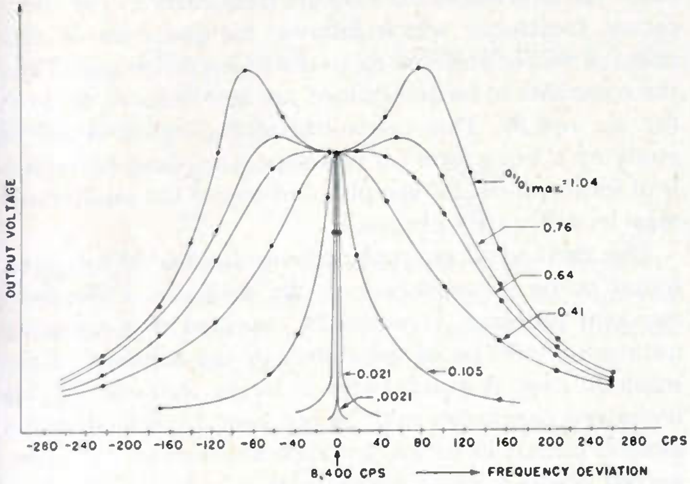


Fig. 8—Response functions measured with filter circuit Fig. 7. Bandwidth determined by the Q of the primary circuit, as indicated. For $Q_1 = Q_{1\,max}$ "maximum bandwidth" is obtained. $Q_{1\,max} = 45$.

$Q_1/Q_{1\,max}$	$K^2 = 5(Q_1/Q_{1\,max})^2$	With Fig. 4, Eq. 33, 20	From Fig. 8 meas.
		b c.p.s.	b c.p.s.
1.04	5.4	256	248
0.76	2.8	229	222
0.64	2.1	211	211
0.41	0.84	117	118
0.105	$5.6 \cdot 10^{-2}$	25	27
0.021	$2.2 \cdot 10^{-3}$	5	4 ± 1
0.0021	$2.2 \cdot 10^{-5}$	0.5	1.2 ± 1

is shunted by a resistor of decreasing value, the voltage across it and, therefore, the voltage induced in the secondary circuit decreases also.

However, the load on the secondary circuit, which is represented by the primary circuit, becomes simultaneously smaller and these two effects cancel. When the primary circuit is short-circuited, the voltage transmitted to the secondary circuit becomes zero, but the amplification of the unloaded secondary-circuit stage becomes infinite. No signal is transmitted, and the output amplitude is determined by self-oscillation.

Acknowledgments

The writer is indebted to Floyd Sabin, who built the circuit of Fig. 7 and measured the response functions shown in Fig. 8 of this paper.

Internal Resonance in Circuits Containing Nonlinear Resistance*

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Summary—The simultaneous presence of oscillations locked in synchronism is investigated for circuits containing a nonlinear element. Under these conditions the circuit is called internally resonant, the two frequencies of oscillation being related by an integral ratio r/s. The discussion is confined to self-excitation in a highly oscillatory circuit connected across a nonlinear element. Nonlinear oscillations are generally characterized by a certain Fourier spectrum of the component frequencies; as a first approximation, it is reasonable to assume that only the fundamental components need be considered. The nonlinear element is described in terms of its current voltage characteristic and an equivalent linearized parameter defined by considering only the two fundamental frequencies. The variation-of-parameters method is applied to the resultant "linearized" equations and the equilibrium conditions determined. The conditions for stable equilibrium are then investigated by assuming a small departure from the equilibrium and studying the motion of the system following this displacement.

I. Introduction

The behavior of vacuum-tube circuits is often determined by the nonlinearity of the circuit elements involved. Considering the self-excited oscillator, the existence of oscillations of finite amplitude in the steady state suggests that there is a nonlinear element which prevents the oscillations from building up indefinitely,[1] which from energy considerations would be impossible. The analysis of oscillator circuits is usually carried out by considering all circuit parameters to be linear in character; however, this method is often inadequate for their analytical treatment. As pointed out by Van der Pol,[2] the existence of self-excited oscillations is directly associated with the nonlinearity of the system and it would be expected that any questions involving these oscillations may be answered only by considering the nonlinearity of the differential equations involved. The investigation of the single-degree-of-freedom problem has been carried out in this manner, but, in general, the study of higher degrees of freedom is rather limited. The higher-degree-of-freedom problem immediately suggests the possibility of obtaining simultaneous oscillations at several frequencies. Major Skinner[3] has studied the two-degree-of-

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[1] Electrical Engineering Staff, Massachusetts Institute of Technology, "Applied Electronics," John Wiley and Sons, Inc., New York, N. Y., 1943.

[2] B. van der Pol, "The nonlinear theory of electric oscillations," Proc. I.R.E., vol. 22, pp. 1051–1086; September, 1934.

[3] Leo V. Skinner, "Criteria for Stability in Circuits Containing Non-linear Resistance," Thesis, University of Illinois; June, 1948.

freedom circuit from this standpoint and has determined the conditions under which simultaneous oscillations at two frequencies are possible. In performing an experimental verification of these conditions it was noticed that under certain conditions the two oscillations became locked in synchronism. The circuit employed was the two-degree-of-freedom coupled circuit (see Fig. 1) with the mutual inductance as the variable circuit parameter and in which Y_e represents the nonlinear element. An interesting feature is that the two oscillations remain locked in synchronism over a certain range of the mutual inductance with the phase angle between these oscillations varying throughout this range (see Fig. 3). Since the two oscillations are locked in synchronism they are related by an integral ratio; thus the condition of internal resonance may be expressed by the following relation:

$$\omega_2 = \frac{r}{s}\omega_1 \qquad \frac{r}{s} \neq 0. \tag{1}$$

The fact that the circuit remains internally resonant over a certain range of the mutual inductance indicates that the equality $\omega_{02} = (r/s)\omega_{01}$ need not be exactly satisfied for the circuit to remain internally resonant. It is the purpose of this paper to study the conditions under which a nonlinear circuit may become internally resonant. The discussion will be limited to self-excitation in the highly oscillatory circuit. The highly oscillatory restriction implies that all the conductances are small. The admittance Y_e describing the nonlinear element is also assumed small.

II. Glossary

ω_1, ω_2 = resonant angular frequencies
ω_{01}, ω_{02} = resonant angular frequencies without dissipation
ϕ_1, ϕ_2 = phase constants
a_1, a_2 = amplitudes of oscillation
ψ_1, ψ_2 = total phases $\psi_1 = \omega_1 t + \phi_1$, $\psi_2 = \omega_2 t + \phi_2$
θ = phase difference referred to fundamental period $= s\phi_2 - r\phi_1$
r/s = an integral ratio relating ω_1 and ω_2
Y_{e1}, Y_{e2} = equivalent linearized admittance
G_{e1}, G_{e2} = equivalent linearized conductance
B_{e1}, B_{e2} = equivalent linearized susceptance
μ = parameter $\ll 1$
δ = logarithmic decrement
$p = j\omega + \delta$.

III. Equivalent Linearized Admittance

Under the conditions discussed above with the two frequencies locked in synchronism, the voltage across the nonlinear element will be essentially of the form

$$v_1 = a_1 \cos(\omega_1 t + \phi_1) + a_2 \cos(\omega_2 t + \phi_2), \tag{2}$$

where a_1, a_2, ϕ_1, and ϕ_2 are the constants to be determined. If the conditions (1) for internal resonance are satisfied, the wave form represented by (2) will be periodic with period $2\pi s/\omega_1 = 2\pi r/\omega_2$. If this is considered as the fundamental period it is possible to speak of the phase difference between the two frequencies with respect to this fundamental period. This phase difference is defined by

$$x = \frac{s\phi_2 - r\phi_1}{sr} = \frac{\theta}{sr}. \tag{3}$$

Since the expression for θ occurs frequently in the theoretical treatment which follows, for purposes of this paper, θ will be referred to as the phase difference. Thus the constants to be determined are now reduced to three (a_1, a_2, and θ). This procedure seems logical since, in studying a wave form[4] of this type, the phase difference is of importance and the phase of one of the oscillations may be arbitrarily chosen.

The method of equivalent linearization[3,5,6] has been found to be of assistance in the solution of the nonresonant problem. It would be expected that a similar notation would be of assistance in the solution of the resonant case. A general idea as to the character of this linearized parameter may be obtained by considering a general circuit in which a nonlinear element Y_e is connected across a linear circuit with admittance Y which is a function of p. In order to obtain self-excitation of this circuit, the total admittance must be zero.

$$Y(p) + Y_e = 0. \tag{4}$$

Since all the conductances as well as the nonlinear element have been assumed small, they will be defined in terms of a small parameter $\mu \ll 1$. (i.e., $G_1 = \mu g_1$, $G_2 = \mu g_2$, $Y_e = \mu y_e$). Therefore, for the general circuit under consideration, the admittance of the linear circuit Y may be considered as being a function of both p and μ, and will be written as $Y(p, \mu)$. Employing this notation (4) may be written as

$$Y(p, \mu) + \mu y_e = 0. \tag{5}$$

In general, the admittance of the linear circuit will be of the form $Y(p, \mu) = A(p, \mu)/B(p, \mu)$, where $A(p, \mu)$ and $B(p, \mu)$ are polynomials in p prime to each other. Employing this notation

$$A(p, \mu) + \mu y_e B(p, \mu) = 0. \tag{6}$$

Assuming a solution for p of the form

$$p = p_0 + \mu p_1 + \mu^2 p_2 + \cdots$$

and expanding the functions $A(p, \mu)$ and $B(p, \mu)$ in Taylor expansions around the point $(p_0, 0)$

$$A(p_0, 0) + \mu p_1 \frac{\partial A}{\partial p}(p_0, 0) + \mu \frac{\partial A}{\partial \mu}(p_0, 0) + \mu y_e B(p_0, 0) = 0, \tag{7}$$

where only terms up to the order μ have been retained. In order to satisfy the condition expressed by (7)

[4] R. G. Manley, "Waveform Analysis," John Wiley and Sons, Inc., New York, N. Y.; 1945.
[5] N. Kryloff and N. Bogoliuboff, "Introduction to Non-Linear Mechanics, Princeton University Press, Princeton, N. J.; 1943.
[6] N. Minorsky, "Non-Linear Mechanics," J. W. Edwards, Ann Arbor, Mich.; 1947.

$$A(p_0, 0) = 0 \tag{8a}$$

$$p_1 \frac{\partial A}{\partial p}(p_0, 0) + \frac{\partial A}{\partial \mu}(p_0, 0) + y_e B(p_0, 0) = 0. \tag{8b}$$

(8a) provides information concerning the resonant angular frequencies without dissipation ω_{01}, ω_{02} (i.e., $p_0 = j\omega_{01}, j\omega_{02}$). (8b) may be solved for p_1 and a solution for p to this degree of approximation obtained

$$p_1 = \frac{-\partial A}{\partial A} \Big/ \frac{\partial A}{\partial p} - y_e B \Big/ \frac{\partial A}{\partial p} \text{ evaluated at } (p_0, 0). \tag{9}$$

It may be shown that both

$$\frac{\partial A}{\partial \mu} \Big/ \frac{\partial A}{\partial p} \quad \text{and} \quad B \Big/ \frac{\partial A}{\partial p}$$

as contained above are even functions of p_0. Therefore, if y_e is real, p_1 will be entirely real. This is true for the nonresonant condition. For the condition of internal resonance two cases must be considered

Case I $\qquad \omega_{02} = \frac{r}{s}\omega_{01} \qquad \omega_2 = \frac{r}{s}\omega_1.$

Case II $\qquad \omega_{02} \neq \frac{r}{s}\omega_{01} \qquad \omega_2 = \frac{r}{s}\omega_1.$

For Case I, since the frequencies ω_{01}, ω_{02} are already related by the ratio r/s and $p_0 = j\omega_{01}, j\omega_{02}$, no frequency "correction" of the order μ need be contributed by μp_1 and p_1 may be entirely real. For Case II, where the difference $\omega_{02} - (r/s)\omega_{01}$ is small (of the order μ), a frequency correction of the order μ must be contributed by μp_1 if the circuit is to be internally resonant. Thus Y_e must take on a complex form similar to the admittance operator in ordinary linear circuit theory.

Assuming that the nonlinear element is defined in terms of its current voltage characteristic of the form

$$i = F(v) = \mu f(v) = \sum_{j=1}^{j=n} c_j v^j \tag{10}$$

an expression for the linearized parameters may be obtained by substituting (2) into (10). It is apparent that in addition to the two basic frequencies this procedure will also introduce combination and harmonic terms. For the nonresonant case these additional terms are usually rejected since the energy at these frequencies is negligible. However, for the resonant case, these terms take on a new significance as some of them may be of frequency ω_1 or ω_2. Examining a general term, cos $[(n-m)\psi_1 - m\psi_2]$, it can be shown that if this is to be of frequency ω_1, ω_2 a minimum value of $n = r+s-1$ is necessary. For this minimum value the frequency will be ω_1 for $m = s$ and ω_2 for $m = s-1$. Thus for a minimum value of n the contributing terms will be

$$\begin{aligned}
&\cos[(r-1)\psi_1 - s\psi_2] \\
&\quad = \cos(\psi_1 + \theta) = \cos\psi_1\cos\theta - \sin\psi_1\sin\theta \\
&\cos[r\psi_1 - (s-1)\psi_2] \\
&\quad = \cos(\psi_2 - \theta) = \cos\psi_2\cos\theta + \sin\psi_2\sin\theta,
\end{aligned} \tag{11}$$

where θ is defined by (3). This expression indicates that, although some combination or harmonic terms may be of frequency ω_1, ω_2, their phase will differ from ϕ_1, ϕ_2. The other terms of frequency ω_1, ω_2 are those usually obtained[3] for the nonresonant case when the expression (10) is written in linearized form as

$$i = G_{e1}' a_1 \cos\psi_1 + G_{e2}' a_2 \cos\psi_2 \tag{12}$$

where G_{e1}', G_{e2}' are entirely real and represent the equivalent linearized conductances for the nonresonant case and all combination and harmonic terms have been dropped. For the resonant case, if n is sufficiently large for the particular ratio r/s under consideration, in addition to the terms included in G_{e1}', G_{e2}' there will also be terms contributed by combination and harmonic terms as indicated in (11). The complete expressions multiplying $a_1 \cos\psi_1$ and $a_2 \cos\psi_2$ will be designated as G_{e1} and G_{e2}. The remaining part of (11) multiplies $a_1 \sin\psi_1$ and $a_2 \sin\psi_2$ and new parameters B_{e1}, B_{e2} will be introduced to represent this contribution and defined so that for the resonant case (10) is written as

$$\begin{aligned}
i &= G_{e1}a_1\cos\psi_1 - B_{e1}a_1\sin\psi_1 \\
&\quad + G_{e2}a_2\cos\psi_2 - B_{e2}a_2\sin\psi_2.
\end{aligned} \tag{13}$$

If

$$\begin{aligned}
Y_{e1} &= G_{e1} + jB_{e1}, \\
Y_{e2} &= G_{e2} + jB_{e2}, \\
i &= Y_{e1}a_1\cos\psi_1 + Y_{e2}a_2\cos\psi_2.
\end{aligned} \tag{14}$$ [7,8]

Y_{e1} and Y_{e2} are complex operators in the same sense that the ordinary linear circuit admittance is complex and will be defined as the equivalent linearized admittances.

IV. Internal Resonance in the Two-Degree-of-Freedom Circuit

Considering the coupled circuit shown in Fig. 1, where the nonlinear element Y_e is defined by (10), the equivalent circuit is as shown in Fig. 2, where

$$\Gamma_1 = \frac{1}{L_1(1-g^2)},$$

$$\Gamma_2 = \frac{g^2}{M(1-g^2)},$$

$$\Gamma_3 = \frac{1}{L_2(1-g^2)},$$

$$g^2 = \frac{M^2}{L_1L_2}.$$

[7] Note.—The operator j as used in (14) is defined by the property

$$j\begin{Bmatrix}\cos\psi \\ \sin\psi\end{Bmatrix} = \begin{Bmatrix}\cos\left(\psi + \frac{\pi}{2}\right) \\ \sin\left(\psi + \frac{\pi}{2}\right)\end{Bmatrix}.$$

[8] E. J. Routh, "Advanced Rigid Dynamics," Macmillan and Co., London, chapt. 6; 1905.

The differential equations of the system are given by

$$(D^2C_1 + \Gamma_1)v_1 - \Gamma_2 v_2 = -\mu D[f(\ ) + g_1]v_1$$
$$-\Gamma_2 v_1 + (D^2C_2 + D\mu g_2 + \Gamma_3)v_2 = 0, \quad (15)$$

where D represents differentiation with respect to time.

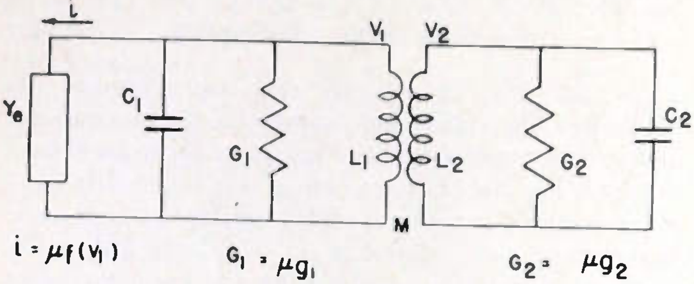


Fig. 1—Two degree of freedom circuit.

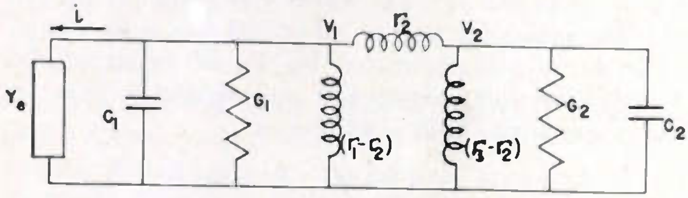


Fig. 2—Equivalent circuit indicated by differential equations.

The expression $-\mu D[f(\ )+g_1]v_1$ could be written as $-\mu[f'(v_1)+g_1]Dv_1$, however it simplifies the analysis considerably to employ the notation given in (15). Eliminating v_2 from these equations and retaining only terms to the order μ as a first approximation

$$(D^2C_1 + \Gamma_1)(D^2C_2 + \Gamma_3)v_1 - \Gamma_2^2 v_1$$
$$= -\mu[D(f(\ ) + g_1)(D^2C_2 + \Gamma_3)v_1$$
$$+(D^2C_1 + \Gamma_1)Dg_2 v_1]. \quad (16)$$

The solution of the homogeneous equation $(\mu=0)$ is of the form $v_1 = a_1 \cos(\omega_{01}t+\phi_1) + a_2\cos(\omega_{02}t+\phi_2)$, where ω_{01} and ω_{02} are known and a_1, a_2, ϕ_1, and ϕ_2 are unknown constants. Under these conditions the expression for v_2 will be of the form

$$v_2 = a_1k_1\cos(\omega_{01}t + \phi_1) + a_2k_2\cos(\omega_{02}t + \phi_2), \quad (17)$$

where

$$k_1 = \frac{\Gamma_1 - C_1\omega_{01}^2}{\Gamma_2} = \frac{\Gamma_2}{\Gamma_3 - C_2\omega_{01}^2},$$

$$k_2 = \frac{\Gamma_1 - C_1\omega_{02}^2}{\Gamma_2} = \frac{\Gamma_2}{\Gamma_3 - C_2\omega_{02}^2}.$$

Proceeding in the usual variation of parameters method, it is assumed that the constants are functions of time and the following expressions are obtained where $\dot{a}_1$ represents da_1/dt and so forth:

$$\dot{a}_1\cos(\omega_{01}t + \phi_1) - a_1\dot{\phi}_1\sin(\omega_{01}t + \phi_1)$$
$$+ \dot{a}_2\cos(\omega_{02}t + \phi_2) - a_2\dot{\phi}_2\sin(\omega_{02}t + \phi_2) = 0,$$

$$-\dot{a}_1\omega_{01}\sin(\omega_{01}t + \phi_1) - a_1\dot{\phi}_1\omega_{01}\cos(\omega_{01}t + \phi_1)$$
$$-\dot{a}_2\omega_{02}\sin(\omega_{02}t + \phi_2) - a_2\dot{\phi}_2\omega_{02}\cos(\omega_{02}t + \phi_2) = 0,$$

$$-\dot{a}_1\omega_{01}^2\cos(\omega_{01}t + \phi_1) + a_1\dot{\phi}_1\omega_{01}^2\sin(\omega_{01}t + \phi_1)$$
$$-\dot{a}_2\omega_{02}^2\cos(\omega_{02}t + \phi_2) + a_2\dot{\phi}_2\omega_{02}^2\sin(\omega_{02}t + \phi_2) = 0,$$

$$\dot{a}_1\omega_{01}^3\sin(\omega_{01}t + \phi_1) + a_1\dot{\phi}_1\omega_{01}^3\cos(\omega_{01}t + \phi_1)$$
$$+\dot{a}_2\omega_{02}^3\sin(\omega_{02}t + \phi_2) + a_2\dot{\phi}_2\omega_{02}^3\cos(\omega_{02}t + \phi_2) = T. \quad (18)$$

T represents the right side of (16) divided by C_1C_2. Solving these equations for $\dot{a}_1$, $\dot{a}_2$, $\dot{\phi}_1$ and $\dot{\phi}_2$

$$\dot{a}_1 = \frac{T}{\omega_{01}^3 - \omega_{01}\omega_{02}^2}\sin(\omega_{01}t + \phi_1),$$

$$\dot{a}_2 = \frac{T}{\omega_{02}^3 - \omega_{02}\omega_{01}^2}\sin(\omega_{02}t + \phi_2),$$

$$\dot{\phi}_1 = \frac{T}{a_1(\omega_{01}^3 - \omega_{01}\omega_{02}^2)}\cos(\omega_{01}t + \phi_1),$$

$$\phi^2 = \frac{T}{a_2(\omega_{02}^3 - \omega_{02}\omega_{01}^2)}\cos(\omega_{02}t + \phi_2). \quad (19)$$

Since the expression for T contains the parameter $\mu \ll 1$, it is reasonable to assume that a_1, a_2, and θ vary only slowly during one period. The equations of the first approximation are obtained by averaging these equations over a period. Actually this represents the first term of a Fourier expansion and a refined first approximation may be obtained by considering the other terms in this expansion. This will add harmonics of the frequency ω_1/s which is the frequency of the periodic wave form.

It will be noticed that the expression for T contains a term $\mu Df(v_1)$ if (2) is substituted into (10) and the differentiation indicated is carried out, the only term originating from $\mu Df(v_1)$ and contributing to the average of $\dot{a}_1$ over a cycle will be the same as was defined G_{e1} previously. Similarly, G_{e2} contributes to $\dot{a}_2$, B_{e1} contributes to $\dot{\phi}_1$, and B_{e2} contributes to $\dot{\phi}_2$. It follows that to this degree of approximation

$$\dot{a}_1 = -a_1\frac{\delta_{01}}{G_{01}}(G_{01} + G_{e1}),$$

$$\dot{a}_2 = -a_2\frac{\delta_{02}}{G_{02}}(G_{02} + G_{e2}),$$

$$\dot{\phi}_1 = -\frac{\delta_{01}}{G_{01}}B_{e1},$$

$$\dot{\phi}_2 = -\frac{\delta_{02}}{G_{02}}B_{e2},$$

$$\omega_1 = \omega_{01} + \dot{\phi}_1,$$

$$\omega_2 = \omega_{02} + \dot{\phi}_2. \quad (20)$$

In the transient state the change in phase difference θ will be

$$\dot{\theta} = s\dot{\psi}_2 - r\dot{\psi}_1$$

$$\dot{\theta} = s\omega_{02} - s\frac{\delta_{02}}{G_{02}}B_{e2} - r\omega_{01} + r\frac{\delta_{01}}{G_{01}}B_{e1}, \quad (21)$$

where

$$G_{01} = G_1 + k_1^2G_2,$$

$$G_{02} = G_1 + k_2^2 G_2,$$

$$\delta_{01} = \frac{(G_1 + k_1^2 G_2)(\omega_{01}^2 C_2 - \Gamma_3)}{2C_1C_2(\omega_{01}^2 - \omega_{02}^2)} = \frac{G_1 + k_1^2 G_2}{2(C_1 + k_1^2 C_2)},$$

$$\delta_{02} = \frac{(G_1 + k_2^2 G_2)(\omega_{02}^2 C_2 - \Gamma_3)}{2C_1C_2(\omega_{02}^2 - \omega_{01}^2)} = \frac{G_1 + k_2^2 G_2}{2(C_1 + k_2^2 C_2)}.$$

For a steady state a_1, a_2 and θ are constant and these expressions reduce to

$$G_{01} + G_{e1} = 0,$$
$$G_{02} + G_{e2} = 0,$$
$$s\omega_{02} - r\omega_{01} + r\frac{\delta_{01}}{G_{01}}B_{e1} - s\frac{\delta_{02}}{G_{02}}B_{e2} = 0. \qquad (22)$$

Since G_{e1}, G_{e2}, B_{e1}, and B_{e2} are all functions of the variables a_1, a_2, and θ, equations (22) may be solved to give the equilibrium values of these unknowns.

V. The Question of Stability

In the previous section the equilibrium conditions have been determined. It is now necessary to determine under what conditions this equilibrium is stable. It has been demonstrated by Liapounoff[8] that the equations of the first approximation are usually sufficient to give a correct answer to the question of stability in a nonlinear system. In order to simplify the notation let

$$\dot{a}_1 = X(a_1, a_2, \theta); \quad \dot{a}_2 = Y(a_1, a_2, \theta); \quad \dot{\theta} = Z(a_1, a_2, \theta). \qquad (23)$$

This notation indicates that X, Y, and Z are functions of (a_1, a_2, θ). Assuming a small departure (ξ, η, ρ) from the equilibrium position, the motion of the system following this displacement is studied. Expanding (23) in a Taylor series around the equilibrium position a_1, a_2, θ, and neglecting second degree and higher powered terms in ξ, η, ρ

$$X(a_1 + \xi, a_2 + \eta, \theta + \rho) = X(a_1, a_2, \theta) + \xi\frac{\partial X}{\partial a_1}(a_1, a_2, \theta) + \eta\frac{\partial X}{\partial a_2}(a_1, a_2, \theta) + \rho\frac{\partial X}{\partial \theta}(a_1, a_2, \theta). \qquad (24)$$

By the notation (23)

$$X(a_1, + \xi, a_2 + \eta, \theta + \rho) = \frac{d(a_1 + \xi)}{dt} = X(a_1, a_2, \theta) + \frac{d\xi}{dt}. \qquad (25)$$

It follows that

$$\frac{d\xi}{dt} = \xi\frac{\partial X}{\partial a_1} + \eta\frac{\partial X}{\partial a_2} + \rho\frac{\partial X}{\partial \theta}$$
$$\frac{d\eta}{dt} = \xi\frac{\partial Y}{\partial a_1} + \eta\frac{\partial Y}{\partial a_2} + \rho\frac{\partial Y}{\partial \theta}$$
$$\frac{d\rho}{dt} = \xi\frac{\partial Z}{\partial a_1} + \eta\frac{\partial Z}{\partial a_2} + \rho\frac{\partial Z}{\partial \theta} \qquad (26)$$

all evaluated at (a_1, a_2, θ). The substitutions $\xi = D_1e^{pt}$, $\eta = D_2e^{pt}$, and $\rho = D_3e^{pt}$ reduce these expressions to a system of algebraic equations

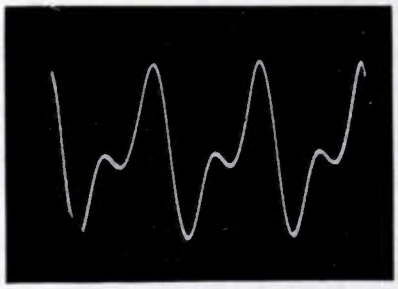
$$\left(p - \frac{\partial Y}{\partial a_1}\right)D_1 - \frac{\partial X}{\partial a_2}D_2 - \frac{\partial X}{\partial \theta}D_3 = 0$$
$$-\frac{\partial X}{\partial a_1}D_1 + \left(p - \frac{\partial Y}{\partial a_2}\right)D_2 - \frac{\partial Y}{\partial \theta}D_3 = 0$$
$$-\frac{\partial Z}{\partial a_1}D_1 - \frac{\partial Z}{\partial a_2}D_2 + \left(p - \frac{\partial Z}{\partial \theta}\right)D_3 = 0. \qquad (27)$$

In order to obtain a nontrivial solution for D_1, D_2, and D_3, their determinant must vanish. Expansion of this determinant provides a cubic equation in p called the characteristic equation of the system.

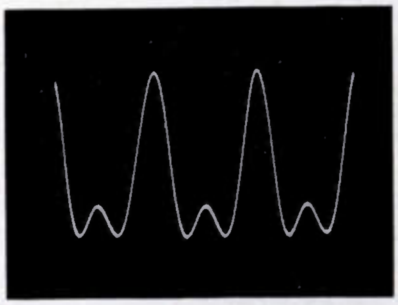
$$p^3 + q_1p^2 + q_2p + q_1 = 0. \qquad (28)$$

A motion of this type is stable[9,10] if the exponent p has no real positive part. Thus the question of stability reduces to an investigation of the roots of the characteristic equation (28). The necessary and sufficient conditions that none of the roots be positive if real or have

(a)



(b)



(c)

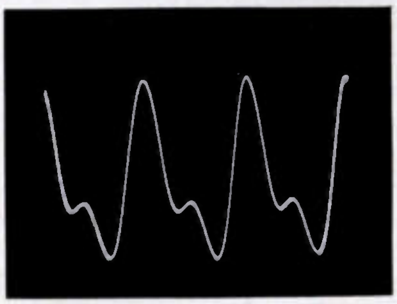


Fig. 3—(a) Internal resonance of order $r=1$, $s=2$, $\theta=90°$. (b) Internal resonance of order $r=1$, $s=2$, $\theta=0°$. (c) Internal resonance of order $r=1$, $s=2$, θ negative.

[9] T. von Karman and M. A. Biot, "Mathematical Methods in Engineering," McGraw-Hill Book Co., New York, N. Y., p. 244; 1940.

[10] E. W. Herold, "Negative resistance and devices for obtaining it," Proc. I.R.E., vol. 23, pp. 1201–1223; October, 1935.

positive real parts if complex are given by the following relations involving the coefficients in (28):

$$q_1 > 0, \quad q_3 > 0, \quad q_1q_2 - q_3 > 0. \tag{29}$$

These conditions also exclude the possibility of pure imaginary roots which would represent oscillation around the equilibrium position without damping.

VI. Application

The preceding theory will now be applied to the case of internal resonance $r=1$, $s=2$, assuming that the nonlinear element is defined by a third-degree polynomial

$$i = \alpha v + \beta v^2 + \gamma v^3, \tag{30}$$

Substituting (2) into (30) the linearized parameters as defined previously may be obtained

$$G_{e1} = \alpha + \frac{3\gamma}{4}(a_1^2 + 2a_2^2) + \frac{a_2^2\beta}{2a_1}\cos\theta,$$

$$G_{e2} = \alpha + \frac{3\gamma}{4}(a_2^2 + a_1^2) + a_1\beta\cos\theta,$$

$$B_{e1} = \frac{a_2^2\beta}{2a_1}\sin\theta,$$

$$B_{e2} = -a_1\beta\sin\theta. \tag{31}$$

The equations of the system may now be written directly as

$$X = -a_1K_1\left[G_{01} + \alpha + \frac{3\gamma}{4}(a_1^2 + 2a_2^2) + \frac{a_2^2\beta}{2a_1}\cos\theta\right],$$

$$Y = -a_2K_2\left[G_{02} + \alpha + \frac{3\gamma}{4}(a_2^2 + 2a_1^2) + a_1\beta\cos\theta\right],$$

$$Z = 2\omega_{02} - \omega_{01} + K_1\frac{\beta a_2^2}{2a_1}\sin\theta + K_2\,2\beta a_1\sin\theta, \tag{32}$$

where $K_1=\delta_{01}/G_{01}>0$, $K_2=\delta_{02}/G_{02}>0$, and the equilibrium conditions are given by setting these expressions equal to zero. The special case $2\omega_{02}=\omega_{01}$ will be investigated since the analysis is simplified appreciably by the fact that under this condition $\sin\theta=0$, and several terms in the stability determinant will be zero. Making use of the equilibrium conditions and expanding the determinant of (27), the following expressions for the coefficients of the characteristic equation are obtained:

$$q_1 = K_1\left[-\frac{\beta a_2^2}{2a_1}\cos\theta + \frac{3}{2}\gamma a_1^2\right] + K_2\left[\frac{3}{2}\gamma a_2^2\right] - \left[2\beta a_1K_2 + \frac{\beta a_2^2}{2a_1}K_1\right]\cos\theta,$$

$$q_2 = K_1K_2\left[-\frac{\beta a_2^2}{2a_1}\cos\theta + \frac{3}{2}\gamma a_1^2\right]\left[\frac{3}{2}\gamma a_2^2\right] - \left[K_1\left(-\frac{\beta a_2^2}{2a_1}\cos\theta + \frac{3}{2}\gamma a_1^2\right) + K_2\frac{3}{2}\gamma a_2^2\right] \cdot\left[2\beta a_1K_2 + \frac{\beta a_2^2}{2a_1}K_1\right]\cos\theta - K_1K_2[\beta a_2\cos\theta + 3\gamma a_1a_2]^2,$$

$$q_3 = -K_1K_2\left[-\frac{\beta a_2^2}{2a_1}\cos\theta + \frac{3}{2}\gamma a_1^2\right]\left[\frac{3}{2}\gamma a_2^2\right] \cdot\left[2\beta a_1K_2 + \frac{\beta a_2^2}{2a_1}K_1\right]\cos\theta + \left[K_1K_2[\beta a_2\cos\theta + 3\gamma a_1a_2\right]^2 \cdot\left[2\beta a_1K_2 + \frac{\beta a_2^2}{2a_1}K_1\right]\cos\theta. \tag{33}$$

Four cases must be investigated

Case I $\gamma<0$; $\beta\cos\theta>0$
Case II $\gamma<0$; $\beta\cos\theta<0$
Case III $\gamma>0$; $\beta\cos\theta>0$
Case IV $\gamma>0$; $\beta\cos\theta<0$.

It can be shown that all the conditions (29) are satisfied only by case IV, subject to the restriction that

$$-\frac{3}{4}\beta\gamma\frac{a_2^4}{a_1}\cos\theta + \frac{9}{4}\gamma^2a_1^2a_2^2 > (\beta a_2\cos\theta + 3\gamma a_1a_2)^2. \tag{34}$$

The conditions for stable equilibrium may be represented graphically as families of curves, Fig. 4, by converting the equations into a normalized form

$$\frac{(a_1-3R)^2}{9R^2(1-S_2)} + \frac{a_2^2}{18R^2(1-S_2)} = 1,$$

$$a_2^2 = -\frac{a_1\left[\frac{3}{2}a_1^2 + R^2S_1\right]}{3a_1 - 9R},$$

$$Ra_2^2 > a_1^3 - 8Ra_1^2 + 12R^2a_1,$$

$$\gamma > 0,$$

$$\beta\cos\theta < 0,$$

$$\cos\theta = \pm 1 \tag{35}$$

where

$$R = -\frac{1}{9}\frac{\beta\cos\theta}{\gamma},$$

$$S_1 = 162\frac{\gamma}{\beta^2}(G_{01} + \alpha),$$

$$S_2 = 6\frac{\gamma}{\beta^2}(G_{02} + \alpha).$$

The accuracy of calculated results will to a great extent be dependent upon the exactness with which the current-voltage characteristic of the nonlinear element can be approximated by a polynomial expansion in v of a limited number of terms. In the original thesis an experimental verification was carried out with a close correspondence between calculated and observed results. In this case the nonlinear element Y_e was a vacuum tube connected as a transitron oscillator.[10,11]

[11] C. Brunetti, "The transitron oscillator," Proc. I.R.E., vol. 27, pp. 88–94; February, 1939.

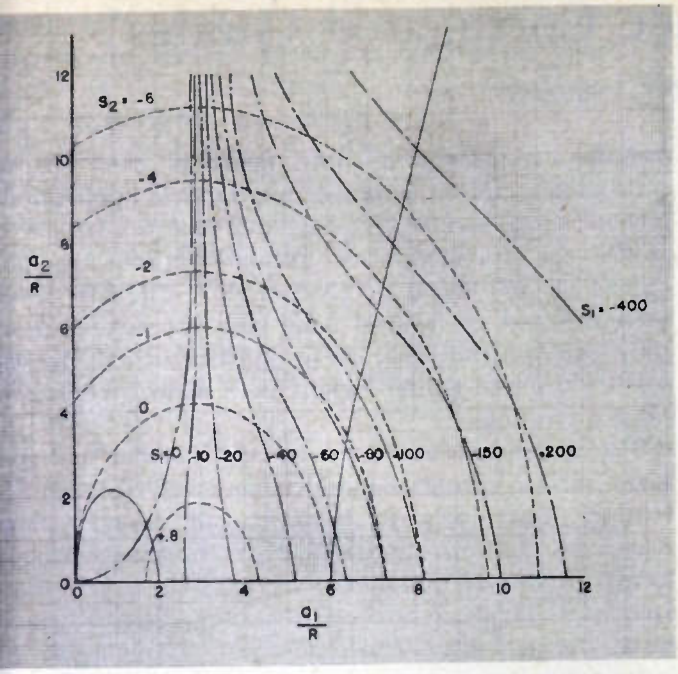


Fig. 4—Conditions for stable equilibrium $\omega_{02} = \frac{1}{2}\omega_{01}$.

VII. Conclusion

The simultaneous presence of oscillations locked in synchronism or otherwise is directly associated with the character of a nonlinear element in an oscillator. The highly oscillatory two-degree-of-freedom coupled circuit has been considered in the analysis; however, the same procedure may be followed for any two-degree-of-freedom circuit which is highly oscillatory and the analysis may readily be extended to higher degrees of freedom. The theory may also be extended to include synchronization with small applied voltages.

In the nonresonant problem[3] the study of equilibrium conditions and stability is restricted by the fact that a separate analysis is required for each value of n as defined in (10). The resonant problem is also complicated by the fact that for a given n the study is dependent upon the particular ratio r/s under consideration. For the two-degree-of-freedom circuit and any order of internal resonance r/s a value of n of at least $r+s-1$ is required. This value of n does not necessarily mean that the oscillations will be stable with the two frequencies locked in synchronism but represents a minimum value which will provide the necessary combination and harmonic terms of the same frequency as either of the two basic frequencies. For simplicity the minimum value of n was employed in the application illustrated in this paper; however the methods indicated are readily applicable to larger values of n.

The resonant frequencies of the circuit without dissipation need not be exactly related by a ratio r/s in order for the circuit to become internally resonant. There exists a zone about the point $\omega_{02} = (r/s)\omega_{01}$ which will provide stable oscillation at two frequencies ω_1, ω_2 which are related by the ratio r/s. This suggests that the nonlinear element may be represented by a parameter which possesses properties similar to an admittance with real and imaginary parts. For the internally resonant case the linearized parameter has been shown to be of the same form as the ordinary linear circuit admittance. The imaginary part of this parameter is provided completely by combination and harmonic terms and is a trigonometric function of the phase difference, $\theta = s\phi_2 - r\phi_1$, between the two oscillations. In the equilibrium state the value of θ reaches an equilibrium just sufficient to make the imaginary parts of the linearized parameters of the correct magnitude and sign to compensate for any discrepancy from the equality $\omega_{02} = (r/s)\omega_{01}$. It is apparent from the form of the linearized parameters that there exists a maximum discrepancy which may be compensated for by a variation in θ, since θ appears in these expressions as a sine function with maximum magnitude 1. Thus there exists a maximum zone over which the two frequencies may remain locked in synchronism. This does not mean that the equilibrium will be stable for all values of θ; the case $\sin\theta = \pm 1$ represents the maximum possible zone. Analysis of the stability of the system by the method indicated in Section V will indicate over what portion of this zone the equilibrium is stable.

In the original thesis two methods of solution were considered in conjunction with this problem; the perturbation method and the variation of parameters method. As a first approximation both methods provide identical results. The variation of parameters method as presented in this paper suggests a relatively more concise method of approach. Both methods may be readily extended to circuits of higher degrees of freedom.

An interesting feature of the results is that the simultaneous presence of oscillations at two frequencies locked in synchronism is possible when the nonlinear element is represented by a polynomial of less than the fifth degree. The number of ratios r/s for which this is possible is limited by the minimum value of $n = r+s-1$. For the nonresonant case it has been shown by Skinner[3] that a polynomial of at least the fifth degree is required for stable operation at the two frequencies simultaneously. However, in this analysis, he excluded the case when any of the combination or harmonic terms arising from the polynomial expansion (10) were of the same frequency as either of the two basic frequencies. This assumption has the effect of excluding the internally resonant condition. For this reason Skinner qualified his statement by defining his frequencies as being incommensurate. Stability in the internally resonant case appears to be a direct result of the additional terms contributed by the combination and harmonic terms which are of the same frequency as either of the two basic frequencies and have been retained in the analysis.

VIII. Acknowledgment

The author wishes to acknowledge the valuable suggestions given by Lloyd T. DeVore in connection with the preparation of this paper.

A Study of Single-Surface Corrugated Guides*

WALTER ROTMAN†, ASSOCIATE, IRE

Summary—A series of experiments were conducted to determine the properties of single-surface corrugated structures as transmission lines and as radiators. Two types of surfaces were tested: the first is a flat grooved plate fed by a waveguide; the second, a circular, corrugated cylinder fed by a coaxial line. A modification of the second type results in a spirally grooved rod with similar properties.

The measured field parameters are found to be predictable from existing theory. For properly designed structures the energy is essentially bound to the corrugated surface; little radiation occurs, and the attenuation of the traveling wave is due chiefly to losses in the metal. The effect of filling the corrugations with solid dielectric is also analyzed.

INTRODUCTION

CONVENTIONAL transmission systems for the guidance of electromagnetic energy at microwave frequencies are generally divided into two classes: coaxial lines and hollow waveguides. Both of these systems effectively confine the electromagnetic energy to a closed region of space by means of conducting walls. A third type of transmission, in which the energy is not confined to a given region but rather is "bound" to a surface,[1,2] is possible under special conditions. Although the energy distribution for such transmission theoretically must extend throughout all space, the decrease in amplitude with distance from the guiding medium may be made so great that for all practical purposes the waves are essentially confined to a region within a few wavelengths of the guide. In general, the surfaces for such transmission must be capable of supporting a tangential component of electric field in the direction of energy propagation. At least four such surfaces are known at the present time: surfaces of media with complex magnetic permeabilities; sheets of dielectric, either with or without a metal backing;[3] sheets formed of imperfect conductors;[4] and corrugated metallic sheets. The waves on such surfaces differ from those in conventional transmission systems with smooth conducting walls in that their impedance and phase velocity are dependent upon the nature of the surface, the velocity always being less than that of free space.

Corrugated structures have long been employed in communications and in allied fields for their special transmission and filter properties. Typical applications include single-surface, coaxial, and hollow corrugated guides for traveling-wave tubes,[5,6] microwave antennas,[7-11] and linear accelerators.[12-14] The low wave velocities and the special band-pass characteristics of these guides permit their use as phase-control devices of smaller size and lighter weight than dielectric-filled or ridge waveguide.

The theoretical treatment of both circular and flat single-surface corrugated guides has received much attention. The analyses, however, all assume some reasonable form of local field distribution without formal proof. Furthermore, the expressions for phase velocity and magnitude of space harmonics are obtained in the form of transcendental equations involving an infinite number of terms. Usually, only the first one or two terms of the series are evaluated because of the complexity of numerical calculations. For these reasons an experimental verification of the important properties of corrugated surfaces seemed advisable.

The experiments, in addition, permit a study of methods of generating the surface-bound waves by means of coaxial or waveguide "launchers" and a study of the effect of modifying the surfaces in some manner not amenable to simple analysis. Likewise, the basic hypothesis that these waves are truly surface-bound and will not radiate may be checked. Finally methods of causing and controlling radiation from slightly altered forms of these surfaces for antenna applications have been investigated, but are described in detail here.

For these measurements, two general types of corrugated surfaces were considered. The first is a flat, grooved plate fed by a waveguide (Fig. 1(center)) and the second is a circular corrugated cylinder (Figs. 1(left) and 2(right)). A modification of the second type

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This material is condensed from the author's "A Study of Single-Surface Corrugated Guides," Air Force Cambridge Research Laboratories, Technical Report #5055, Cambridge, Mass. The CRL Report also includes a discussion of the radiation characteristics of modified corrugated surfaces.

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[1] For a general discussion of the nature of surface waves, see J. A. Stratton, "Electromagnetic Theory," McGraw-Hill Book Co., New York, N. Y., and London, pp. 583–587; 1941.

[2] G. Goubau, "Surface wave transmission line," *Radio and Television News*, vol. 43, p. 10; Radio-Electronic Engineering Section, May, 1950.

[3] D. W. Fry, "Some recent developments in the design of centrimetric aerial systems," *Proc. IEE*, vol. 93, Part IIIA, Sec. 10, p. 1947; 1946.

[4] See pp. 524–534 of footnote reference 1.

[5] J. R. Pierce, "Traveling-wave tubes," *Bell. Sys. Tech. Jour.*, vol. 29, p. 189, Chapters IV, V and VI; April, 1950.

[6] "Traveling Wave Tubes Project," 3d Quarterly Progress Report, (Contract No. N6-ONR 251), Stanford Univ., Period Covering July 1, 1947 to September 30, 1947.

[7] H. Goldstein, "Theoretical Analysis of Corrugated Surfaces," Ph.D. Thesis, MIT; 1943.

[8] H. Goldstein, R. L., Report 494, 1943. "The Theory of Corrugated Transmission Lines and Waveguides," April, 1944.

[9] For the most recent advances in the application of corrugated surfaces to antenna problems, see "Ridge and Corrugated Antenna Studies," Stanford Research Institute, Quarterly Progress Reports 1–5, Covering Period July, 1949, through October, 1950.

[10] W. E. Kock, "Metallic delay lenses," *Bell Sys. Tech. Jour.*, vol. 27, pp. 58–82; January, 1948.

[11] W. E. Kock, "Refracting sound waves," *Acous. Soc. Amer. Jour.* vol. 21, pp. 471–481; September, 1949.

[12] R. B. R. Shersby-Harvie, "Travelling wave linear accelerators," *Proc. Phys. Soc.*, vol. 61, Part 3, p. 255; September 1948.

[13] L. Brillouin, "Waveguides for slow waves," *Jour. Appl. Phys.*, vol. 19, p. 1023; November, 1948.

[14] E. L. Chu and W. W. Hansen, "The theory of disk-loaded waveguides," *Jour. Appl. Phys.*, vol. 18, p. 996; November, 1947.

esults in a spirally grooved rod (Figs. 1(right) and (left)) with similar electromagnetic properties. The esults of the measurements and comparison with heory are given in the remainder of this report.

The following properties were measured for each of he corrugated surfaces considered: phase velocity and ttenuation for the dominant mode as a function of requency, the external field distribution in three orthogonal directions, and the efficiency of the "launchers." If no surface radiation occurs the attenuation of the traveling wave after it has left the launcher will be just that due to metal and dielectric losses. An additional check for radiation losses consists of measuring field intensity far from the corrugated surface with an exploring dipole.

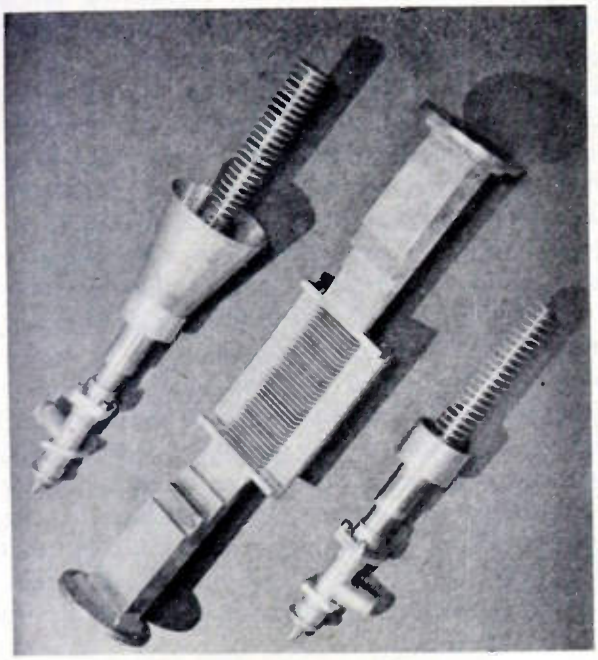

Fig. 1—Single-surface corrugated guides.

Properties of Corrugated Surfaces

The electromagnetic characteristics of both the flat and the circular corrugated surfaces may be described in the same qualitative terms. In both cases the corrugations are of rectangular cross section and uniform depth. The dominant mode propagates in a direction normal to the length of the corrugations, which is assumed infinite in extent for the flat surface.

The general theory indicates that the surface will truly act as guides for electromagnetic waves, provided the loading due to the slots cut in the surface is series-inductive. This inductive loading occurs when the depth of the slot (the slot being considered as a radial transmission line for the cylindrical case) is less than a a quarter wavelength plus an integral number of half wavelengths. The guide thus exhibits band-pass characteristics, with transmission possible over a number of discrete bands of frequencies. The phase velocity along the surface is a function of the slot depth, varying from free-space velocity at zero depth to zero velocity at one-quarter wavelength depth. The ratio of electric to magnetic fields in the transverse plane varies from the free-space ratio for no slot to infinity for a quarter-wave slot depth. The variation of field intensity as a function of distance from the surface changes from a small exponential decrease for a shallow slot to a large exponential decrease as the slot depth approaches a quarter wavelength. (The exponential is replaced by a Hankel Function for the cylindrical case, but the effect is the same.) Since the magnetic fields lie totally within the transverse plane, the mode is classified as a transverse magnetic or *TM* type.

In addition to the fundamental mode described, an infinite number of space harmonics is associated with the traveling wave. The relative amplitudes of these harmonics are functions of the slot width, depth, and spacing. These harmonics, caused by diffraction at the individual corrugations, may have either positive or negative phase velocities corresponding to propagation towards or away from the source, respectively. In particular, if the slots are spaced one-half guide wavelength apart, a space harmonic of amplitude and phase velocity equal to that of the fundamental, but of negative sign, will be established, producing a resonant condition in the line. Although this phenomenon imposes further restraint upon permissible regions of energy transmission, effect of the space harmonics may be reduced greatly by spacing the corrugations closely.

Measurements on Flat Corrugated Surfaces

Although it is difficult to build an extremely wide, flat, corrugated structure to approximate the theoretical

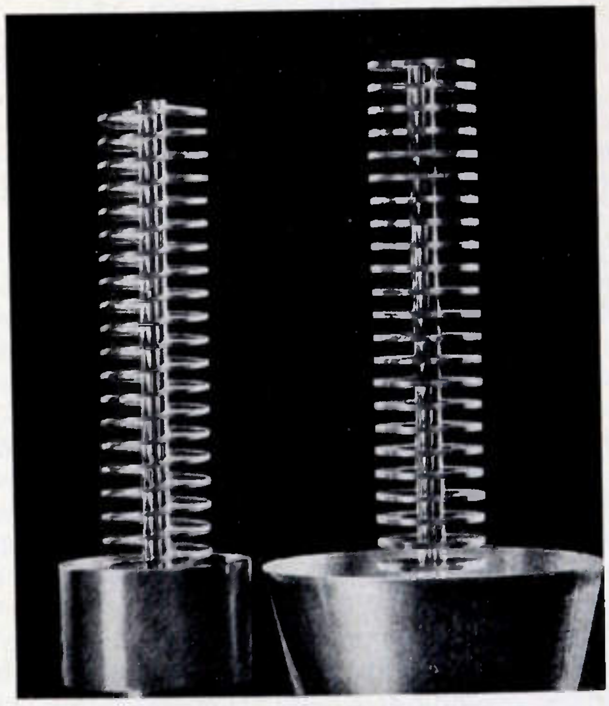

Fig. 2—Close-up of single-surface guides.

infinitely wide model, it seemed probable that a corrugated strip of width comparable to a wavelength might have similar properties. No theoretical treatment of this type of structure is known to exist; also, it was thought that substantial radiation with a corresponding excessive attenuation of the traveling wave might possibly take place at the two edges of the strip.

Accordingly, the first tests investigated whether power could be transmitted effectively over a reasonable length of such a guide. A strip corrugated guide 0.85λ wide at the design wavelength of 6 centimeters, fed by a rectangular waveguide transformer, was constructed according to the dimensions shown in Figs. 1 and 3. This unit was designed (on the basis of Equation (1) of the Appendix) to have a phase velocity one-half that of light at the design wavelength. The strip was initially terminated by a second waveguide transformer, identical to the source feed, (Fig. 1 in the hope that

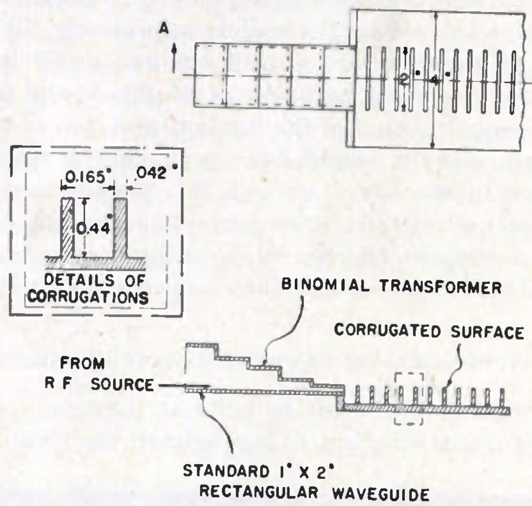

Fig. 3—Single, flat, corrugated surface with waveguide feed.

the energy would be transmitted effectively across the strip and be received in the termination. A comparison of the power transmitted by the source and that received in the load would then give a figure of merit for the entire transmission system.

The losses in power from the system are of two types: radiation losses at the transmitting and receiving junctions, and losses from the corrugated surface due to attenuation in the metal and possible radiation along the line. Since the corrugated surface losses are roughly proportional to the length of the strip while the junction losses are essentially independent of this length, successive measurements of power transmission for different lengths of strip should permit the separation of the two types of losses.

Tests soon demonstrated the practical difficulties of this measurement. Although a power transmission ratio of 70 per cent (ratio of output to input) was obtained for a corrugated guide length of 10 inches, this ratio changed by less than 1 per cent when the path length was doubled. The losses from the junctions were evidently great enough to obscure completely (within the precision of measurement) the losses from the corrugated strip.

Since it was now known that the radiation losses from the corrugated strip itself were not excessive, an exploration of the surface fields by means of rf probe techniques was attempted. The guide wavelength was determined by setting up standing waves on the corrugated line and measuring the distance between minima of field intensity, which are a half guide wavelength apart. For this measurement the terminating waveguide was replaced by a solid metallic reflecting plate which acted approximately as a short circuit at the far end of the strip. The attenuation along the corrugated line could also be obtained by measuring the standing-wave ratio as a function of distance from the source.

The rf probe, a rigid coaxial cable of small diameter with the center conductor protruding from one end, was mounted on a small lathe bed with the probe tip normal to the corrugated surface. Three-dimensional motion of the probe tip was obtained, the rf output of the coaxial line being proportional to the normal component of the electric field intensity. The distortion of the field by the presence of measuring devices was investigated by observing the magnitude of the wave reflected by the probe as its distance from the corrugated surface was varied from 1/16 to 1.5 inches. The observed distortion at any position was not sufficient to warrant any special precaution. In addition, the input impedance to the corrugated guide launcher, which would be altered by the presence of a highly reactive probe, was measured with and without the probe in place. Fortunately, no difference in impedance value was observed.

The variation in field intensity over the corrugated surface was measured in three orthogonal directions: along the guide axis, along a straight line in the transverse plane parallel to the surface, and normal to the surface. Comparison with the theoretical values obtained for an infinitely wide plate was made whenever applicable. Since the positions of minima of electric field were not spaced at exactly equal intervals along the line because of the disturbing effect of the higher-order space harmonics, the guide wavelength was obtained by averaging the distance between a large number of consecutive minima. For small magnitudes of space harmonics, this method should very nearly balance out their perturbation of the fundamental field since the phase velocities of the space harmonics are usually quite different than that of the dominant mode.

The measured guide wavelength as a function of free-space wavelength is compared (Fig. 4) to the theoretical values as obtained from two separate analayses by Cutler and Walkinshaw which result in equations (1) and (4), respectively, of the Appendix. On the basis of the mathematical and physical techniques employed, (1) may be regarded as a zero-order approximation and (4) as a first-order approximation to the exact solution. It is seen that, while both equations show fair agreement with experiment, (4) gives a closer approximation. Presumably, if the mathematical solution could be made exact by including all the terms

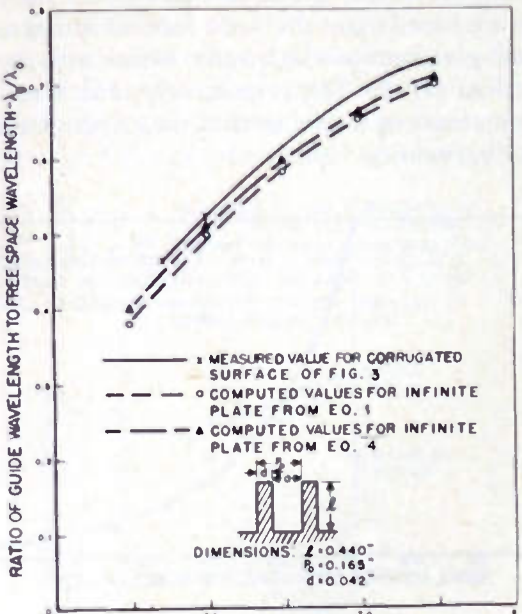

Fig. 4—Guide wavelength of dominant mode on a flat, corrugated surface.

representing space harmonics, an even better correlation could be obtained. Perfect agreement between theory and measurement for this structure is not to be expected, since the theory was derived on the basis of an infinitely wide corrugated plate with no variation in field intensity across its width, while the corrugated strip tested was less than a wavelength wide and showed some variation of field intensity with width.

This measured variation of field intensity is shown in Fig. 5. It is seen that the intensity remains within 60 per cent of the maximum value over the width of the strip, but beyond the edge it falls off at a rapid rate, being only 5 per cent of the maximum value at ½ inch beyond one edge. This experiment, which indicates that very little radiation occurs from the edges, is substantiated by the attenuation measurements. The variation in field intensity in the direction normal to the surface was measured for two different wavelengths. The expected exponential decrease in intensity with distance was obtained with good agreement between observed and theoretical values (as obtained from (1), (7) and (9) of the Appendix).

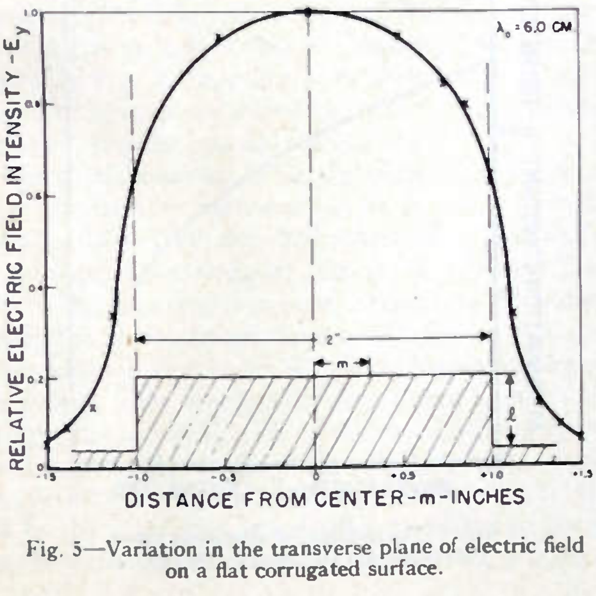

Fig. 5—Variation in the transverse plane of electric field on a flat corrugated surface.

The attenuation along the corrugated strip was measured by a comparison of the standing wave ratios at various positions along the strip. The attenuation constant measured at a wavelength of 6 cm was 0.4 db per meter, and at 7 cm was 0.3 db per meter. These values are approximately what would be expected from calculations based upon metal losses alone and are low enough for the corrugated strip to be used as an effective transmission line over short distances (providing that a nonradiating "launcher" can be constructed).

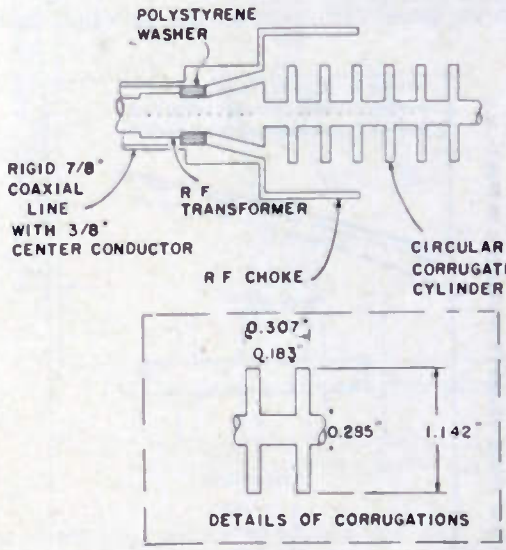

Fig. 6—Circular corrugated cylinder with coaxial feed.

A detailed description of the rectangular waveguide feed and binomial transformer is not presented, since it is believed that more effective methods of coupling the corrugated strip to the waveguide may be designed. Measurements indicated that about 15 per cent of the power is radiated into space at each junction. A more efficient junction is produced if the corrugations are built up gradually within the waveguide so that the electromagnetic energy is already bound to the guide surface before the opening is reached, any abrupt discontinuity in field conditions thereby being avoided.

Measurements on Circular Corrugated Cylinders

A circular corrugated cylinder fed by a rigid coaxial line was constructed to the dimensions shown in Figs.

2 and 6. The outer circumference of the cylinder was restricted to less than one wavelength at all operating frequencies to prevent the propagation of higher order modes exhibiting field variation in azimuth. This guide was designed (equation (2) of the Appendix) to have a phase velocity one half that of free space at a wavelength of 10 cm. Coaxial transformers match the impedance of the corrugated line to that of the coaxial line. Since a gradual tapering of the coaxial line to the corrugations appears to offer greater promise for a wide-band junction, the analysis of the binomial transformer will not be discussed here.

The techniques used to measure the properties of the circular corrugated cylinder were very similar to those employed for the flat corrugated surface. Two types of tests were made: the first was to measure the power transmission over the surface of the cylinder into a resistive load; the second, to probe the external rf fields on the guide surface.

In the first type of test an over-all efficiency of power transfer of 70 to 75 per cent was realized, with the efficiency being only slightly dependent upon the length of line used. Although the addition of rf chokes and shields to suppress radiation at the junction increased

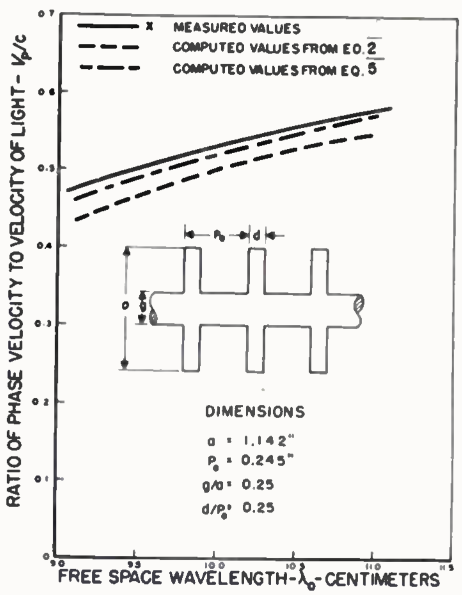

Fig. 7—Phase velocity of dominant mode on a circular corrugated guide.

the efficiency to over 90 per cent, the inaccuracies in power measurements made estimation of attenuation losses on the corrugated surface undependable.

The second method—that of probing the external rf fields on the guide surface—again offered more promise. The same probe used for the flat, corrugated plate measurements was retained, with its output connected to a 10-cm spectrum analyzer. The radial electric field was measured in three orthogonal directions: parallel to the axis of the guide, along a radius, and around the circumference of the cylinder. Measurement of guide wavelength was made in the same manner as for the flat plate. The dependence of guide wavelength upon free-space wavelength is shown in Fig. 7 and compared to the theoretical results obtained from (2) and (5) of the Appendix. The similarity of these results to those for the flat corrugated plate (Fig. 4) is worthy of note. In particular, (4) for the flat surface and (5) for the circular surface are based upon the same form of initial assumptions and give results which agree better with measured values than (1) and (2), respectively, which are cruder approximations in the sense that they neglect the effect of space harmonics.

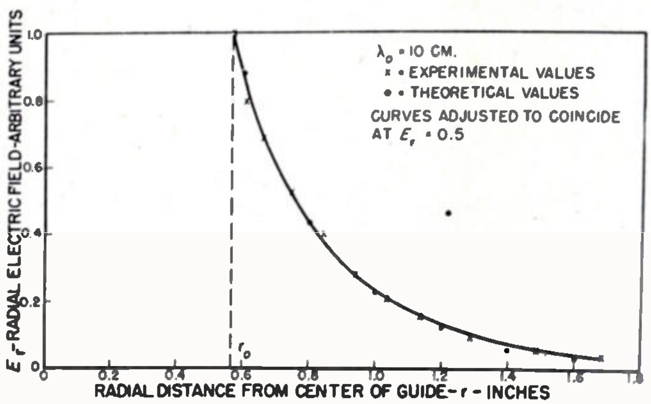

Fig. 8—Radial variation of dominant mode on a corrugated circular cylinder.

As symmetry arguments would indicate, no variation in field intensity with azimuth angle ϕ was found to exist. The radial dependence of the electric field agrees well (Fig. 8) with the theoretical expectation (as computed from (8) and the measured guide wavelength), provided that computed and measured values are adjusted for coincidence at one particular value of the radius (since only relative values of field intensity were measured).

The attenuation along the corrugated cylinder, which

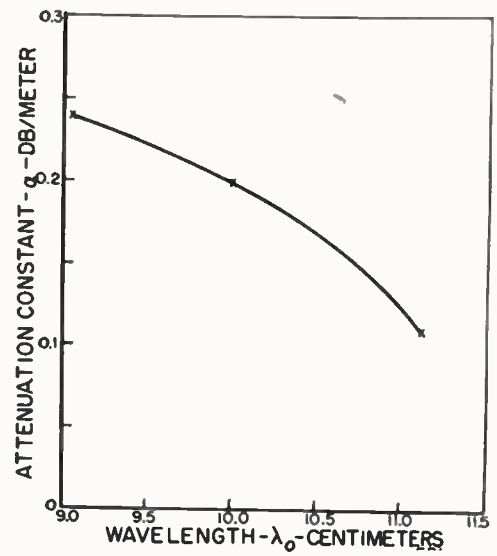

Fig. 9—Attenuation of dominant mode on a circular corrugated cylinder.

was measured by the standing-wave-ratio method previously described, is shown in Fig. 9 as a function of free-space wavelength. It is estimated that these attenuation constants are approximately equal to the theoretical values computed on the basis of metal losses on the surface of the cylinder. The prediction that the surface would not radiate if inductively loaded is thus supported.

For size reduction and mechanical purposes, the effect of filling the slots of the corrugated cylinder was studied. The dielectric-loaded, cylindrical surface guide (see sketch in Fig. 10) was investigated by turning down the

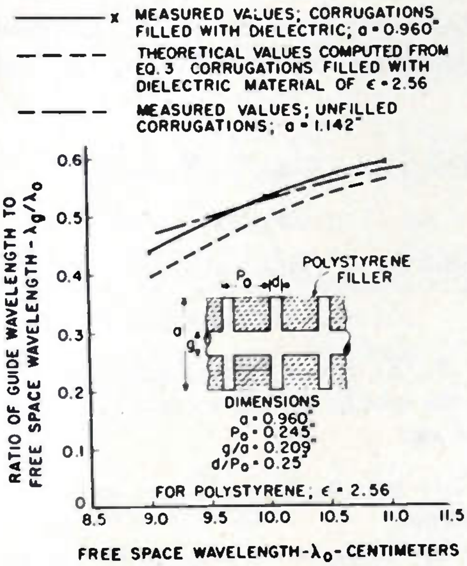

Fig. 10—Guide wavelength of dominant mode on a circular corrugated cylinder filled with dielectric.

outer diameter of the original cylinder from 1.142 to 0.960 inch and filling all of the corrugations with polystyrene ($\epsilon = 2.56$). The guide was fed by the same coaxial transformer used for the cylinder before modification. According to (3), the dominant mode in the resultant smooth, corrugated cylinder should have the same phase velocity ($\frac{1}{2}$ that of free space) at 10-cm wavelength as before modification. The measured guide wavelength as a function of free-space wavelength is shown in Fig. 10 together with theoretical values predicted by (3) and measured values for the air-filled, unmodified, circular corrugated cylinder.

The guide wavelength at the design frequency is almost identical for both models; however, in both structures it differs from the theoretical value (one half that of free-space wavelength), based on (2) and (3), respectively, by about 5 per cent. Equation (3) is derived under the same assumptions as (1) and (2). If a more exact numerical analysis which evaluates all the space harmonics were employed, very likely a much closer theoretical approximation to the measured values could be obtained. The measured rate of change of guide wavelength with free-space wavelength is practically identical to the computed value. The measured attenuation on the dielectric-filled corrugated cylinder is more than double that for the unfilled corrugations, which seems reasonable since the losses in the dielectric medium are comparable in magnitude to those due to the currents in the metal.

The relative mechanical simplicity of constructing metallic cylinders with continuously spiraling grooves around their peripheries created an interest as to whether the modes previously considered would propagate on a spiral as well as on transverse corrugations. Although intuition indicated that the dominant mode would still propagate, theoretical analysis appeared to be prohibitively difficult because of the peculiar geometry. Accordingly, an experimental check was attempted. A spiral structure (Figs. 1 and 2) whose dimensions (outer and inner diameters, wall thickness, and pitch) were equal to the corresponding parameters of the air-filled circular corrugated cylinder, was constructed. The measured guide wavelengths (Fig. 11)

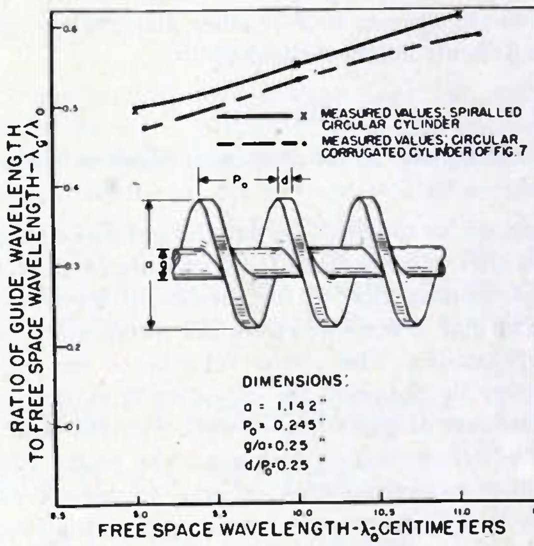

Fig. 11—Guide wavelength of dominant mode on a spiralled circular cylinder.

agree with the values obtained for the transversely corrugated guide to within 3 per cent over a considerable range of frequencies. The radial dependence of electric fields also agrees closely with the values obtained for the transverse corrugations. The chief difference in the two models appears to be a slight azimuthal dependence of fields in the spiral case.

Conclusion

The experiments described in this report demonstrate conclusively that surface-bound waves may exist and propagate on snigle-surface corrugated structures of various geometric shapes whose propagation characteristics are adequately described by existing theoretical treatments. The attenuation on such guides is due chiefly to metal losses, rather than radiation. Other types of periodically varying surfaces, such as circular cylinders cut with spiraling grooves, may transmit the same types of waves as those permissible on the transversely corrugated structures. Further investigation, not described here, has shown that controlled radiation may be produced either by matching the impedance of the wave on the corrugations to free space by means of tapered corrugated sections, or by introducing abrupt, discrete discontinuities in the periodicity of the surface. The former method gives radiation patterns which, at the present stage of the art, have poor radiation pattern characteristics, while the latter technique offers practical difficulties in impedance matching to the source of power.

A greater control over phase velocity as a function of frequency has recently been made possible by a novel technique of periodically varying the depth of the corrugations.[15] In particular, the phase velocity in a rectangular hollow waveguide with one wall corrugated and periodically detuned has been kept practically constant over a considerable band of wavelengths. The application of this analysis to single-surface guided waves would appear to be rather straightforward, although difficult mathematically.

Appendix

The Phase Velocity of the Dominant Mode on Corrugated Surfaces

Equations for the fields on flat and circular corrugated surfaces are available from the literature in the form of an infinite summation of surface-bound waves outside the guide and a somewhat similar summation within the corrugations. The phase velocity of the external waves may be obtained by imposing boundary conditions. Because of the complex nature of the fields, the mathematical process of evaluating the phase velocity, although a straightforward process, is usually carried out only to a first- or second-order approximation leading to slightly different values, depending upon how many terms are retained.

Simple approximate expressions for the phase velocities on the flat and the circular corrugated surfaces have been derived by Cutler[16] on the basis that only the dominant or fundamental mode propagates on the guide surface. His method of analysis matches the average impedance at the common boundary between this fundamental mode outside the guide and a transmission-line mode inside the corrugations or slots. His relation between the guide dimensions and the guide wavelength may be written as follows:

a. For the infinitely wide flat corrugated surface: (see Fig. 4 for parameters).

$$\frac{1}{\tan \beta_0 l} = \left(\frac{P_0 - d}{P_0}\right)\left(\frac{\beta_0}{\gamma_a}\right). \tag{1}$$

[15] A. W. Lines, G. R. Nicoll, and A. M. Woodward, "Some Properties of Corrugated Waveguide," British TRE Report No. T2114, Telecommunication Research Establishment, August, 1948.

[16] C. C. Cutler, "Electromagnetic Waves Guided by Corrugated Conducting Surfaces," Bell Telephone Labs, Inc., Whippany, N. J., October, 1944.

b. For the circular corrugated cylinder: (see Fig. 7 for parameters)

$$\frac{F_1(\beta_0 a)}{F_0(\beta_0 a)} = \left(\frac{P_0 - d}{P_0}\right)\left(\frac{\beta_0}{\gamma_a}\right)\frac{K_1(\gamma_a a)}{K_0(\gamma_a a)} \tag{2}$$

where

$$F_0(\beta_0 a) = J_0(\beta_0 a)N_0(\beta_0 g) - N_0(\beta_0 a)J_0(\beta_0 g)$$

$$F_1(\beta_0 a) = J_1(\beta_0 a)N_0(\beta_0 g) - N_1(\beta_0 a)J_0(\beta_0 g).$$

c. For the circular corrugated cylinder with the slots filled with dielectric: (see Fig. 10 for parameters)

$$\frac{F_1(\beta_0\sqrt{\epsilon}\, a)}{F_0(\beta_0\sqrt{\epsilon}\, a)} = \left(\frac{P_0 - d}{P_0}\right)\left(\frac{\beta_0}{\sqrt{\epsilon}\,\gamma_a}\right)\frac{K_1(\gamma_a a)}{K_0(\gamma_a a)} \tag{3}$$

where

$$F_0(\beta_0\sqrt{\epsilon}\, a) = J_0(\beta_0\sqrt{\epsilon}\, a)N_0(\beta_0\sqrt{\epsilon}\, g) - N_0(\beta_0\sqrt{\epsilon}\, a)J_0(\beta_0\sqrt{\epsilon}\, g)$$

$$F_1(\beta_0\sqrt{\epsilon}\, a) = J_1(\beta_0\sqrt{\epsilon}\, a)N_0(\beta_0\sqrt{\epsilon}\, g) - N_{10}(\beta_0\sqrt{\epsilon}\, a)J_0(\beta_0\sqrt{\epsilon}\, g).$$

J_0, J_1, N_0, K_0, and K_1 are all standard Bessel functions. ϵ is the specific inductive capacity of the dielectric in the slot and

$\beta_0 = \sqrt{k_a^2 - \gamma_a^2} = 2\pi/\lambda_0$ is the free-space propagation constant

$k_a = 2\pi/\lambda_g = 2\pi f/v_p$ is the propagation constant along the surface of the guide

$v_p =$ the phase velocity

$l =$ the slot depth

$P_0 =$ the distance between corresponding points in adjacent slots

$d =$ the thickness of the slot walls

$g =$ the inner radius of cylinder

$a =$ the outer radius of cylinder.

An alternate expression which acknowledges the presence of the space harmonics has been derived by Walkinshaw[17] in his analysis of the circular hollow waveguide with internal corrugations. The method assumes that both the fundamental and the space harmonics are present on the surface and that only the transmission-line mode exists within the corrugations. The external and internal fields are then matched at the common boundary by means of a Fourier analysis, assuming an appropriate form of electric field distribution over the mouth of each corrugation from quasi-static approximations. The result is obtained in the form of an infinite series, each term of the series being associated with an individual space harmonic. To simplify computation, only the first (or dominant) term of the series is retained, leaving the result:

[17] W. Walkinshaw, "Theoretical design of linear accelerator for electrons," *Proc. Phys. Soc.*, vol. 61, part 3, p. 246; September, 1948.

a. For the infinitely wide flat corrugated surface

$$\frac{1}{\tan \beta_0 l} = \left(\frac{P_0 - d}{P_0}\right)\left(\frac{\beta_0}{\gamma_a}\right)\left[J_0\left(k_a \frac{P_0 - d}{2}\right)\right] \times \left\{\frac{\sin [k_a(P_0 - d)/2]}{k_a(P_0 - d)/2}\right\}. \tag{4}$$

b. For the circular corrugated cylinder

$$\frac{F_1(\beta_0 a)}{F_0(\beta_0 a)} = \frac{P_0 - d}{P_0} \frac{K_1(\gamma_a a)}{K_0(\gamma_a a)}\left[J_0\left(k_a \frac{P_0 - d}{2}\right)\right] \times \left\{\frac{\sin [k_a(P_0 - d)/2]}{k_a(P_0 - d)/2}\right\} \tag{5}$$

where all symbols are defined as for (1), (2), and (3).

Equations (4) and (5) are similar to (1) and (2) respectively, except for the factor

$$\left[J_0\left(k_a \frac{P_0 - d}{2}\right)\right]\left\{\frac{\sin [k_a(P_0 - d)/2]}{k_a(P_0 - d)/2}\right\}$$

which is of the form $(J_0(X))$ $(\sin X/X)$ and is a slowly varying function for small values of X, being approximately equal to unity. Both equations give results which agree within a few per cent of one another.

The relation between guide wavelength, free-space wavelength, and phase velocity is

$$\frac{\lambda_g}{\lambda_0} = \frac{v_p}{c} \tag{6}$$

where c is the velocity of light.

The variation of field intensity (see the theoretical curve of Fig. 8) in the direction normal to the corrugated surfaces is given by the following:

a. for the flat corrugated surface of infinite width,

$$E = E_0 \epsilon^{-\gamma_a y}; \tag{7}$$

b. for the circular corrugated cylinder,

$$E = E_0 H_1^{(2)}(j\gamma_a r) \tag{8}$$

where $H_1^{(2)}(X)$, the Hankel function of the first order and second kind, is a real quantity for imaginary values of X and decreases monotomically with increasing X, and γ_a is given by

$$\gamma_a = 2\pi \sqrt{\frac{1}{\lambda_g^2} - \frac{1}{\lambda_0^2}}. \tag{9}$$

Acknowledgment

The author wishes to acknowledge the substantial contributions of F. J. Zucker, who first suggested the corrugated guide experiments. His comments on the theoretical aspects of the subject assisted materially in determining the course of research.

The author is also grateful to R. M. Barrett, who supervised and directed the research program, and to John E. Walsh for his valuable criticism of the text.

Approximation Methods in Radial Transmission Line Theory with Application to Horns*

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Summary—**The radial transmission line theory provides a rigorous description of the propagation of electromagnetic energy in certain cylindrical regions. The complexity of direct calculation may be avoided by a simpler, although approximate, description based on the asymptotic identity of radial and uniform transmission line formulas at large radii. Approximate expressions are developed for input admittance, frequency sensitivity, and higher mode interaction effects on radial lines. Estimates of the order of accuracy are included, and applications are made to sectoral horns.**

I. Introduction

Many microwave structures can be regarded as composite structures whose constituent elements are waveguide regions and discontinuity regions. The applicability of transmission line methods to the analysis of the propagation of electromagnetic waves in the waveguide regions is well known.[1] Such analyses are based on a representation of an arbitrary electromagnetic wave in terms of elementary modes of propagation.

If for any mode of propagation a voltage V and a current I are employed as measures of the electric field and magnetic field, transverse to the propagation direction r, then V and I obey transmission line equations. In uniform waveguide regions, i.e., regions in which cross-sectional surfaces transverse to the propagation direction are identical, one finds that Z_0 and κ, the characteristic impedance and the propagation wave number, respectively, indicative of both the waveguide cross section and the mode form, are constants independent of the coordinate r. In many nonuniform waveguide regions, i.e., regions in which the cross-sectional surfaces transverse to the transmission direction are merely

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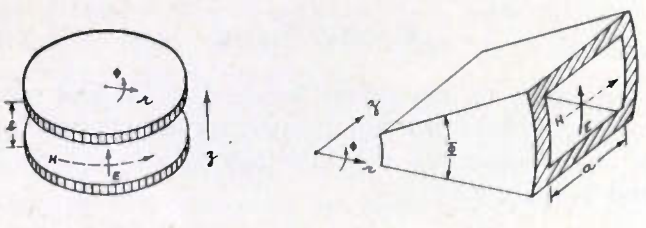
[1] S. A. Schelkunoff, "Electromagnetic Waves," D. Van Nostrand Co. Inc., New York, N. Y., in particular, Secs. 8.6 and 10.13; 1943.

similar to one another, one finds that Z_0 and κ are no longer constants, but are in general dependent on the transmission coordinate r.

In the following we shall consider a class of nonuniform waveguide regions, the so called radial (or cylindrical) waveguide regions illustrated in Fig. 1, wherein the cross-sectional surfaces are of cylindrical form or sectors



(a) (b)

Fig. 1—Typical radial waveguides. E-type and H-type modes can exist in the geometry of either (a) or (b), but a Φ independent E-type mode can exist only in the geometry of (a).

thereof. The cross-sectional surfaces are described by cylindrical coordinates $z\,\phi$ and the transmission direction is characterized by the coordinate r. Such waveguide regions are divided into two types: the E-type characterized by a field pattern with an E_z but no H_z component of field, and the H-type characterized by an H_z but no E_z component of field. As in a uniform transmission line, the radial transmission line description[2,3] is conveniently expressed in terms of either normalized admittances or reflection coefficients.

For convenience we summarize[4] the transmission line description for the case of an axially symmetric H-type radial line. The fundamental admittance relation is

$$Y'(r) = \frac{j\zeta(x, y) + Y'(r_0)Ct(x, y)}{\zeta(x, y)ct(x, y) + jY'(r_0)} \tag{1}$$

where the notation follows that given in footnote reference 3, equation (38) and where $Y'(r)$ is the normalized admittance at any radius r looking in the direction of increasing r. Alternatively, the fundamental reflection coefficient relation for an H-type radial line is

$$\Gamma_v(r) = \Gamma_2(r_0)e^{j2[\eta_1(x)-\eta_1(y)]}, \tag{2}$$

where $\Gamma_v(r)$, the voltage reflection coefficient at r, is defined as the ratio of the amplitudes of the outgoing and the ingoing (Hankel function) waves at a point r. Equation (2) is analogous to equation (47a) of footnote reference 3.

The relation between admittance and reflection coefficient at any point on an H-type radial line is given by

$$Y'(r) = Y_\infty'(r)\frac{1 - \Gamma_v(r)e^{j2[\eta_0(x)-\eta_1(x)]}}{1 + \Gamma_v(r)}, \tag{3}$$

[2] S. Ramo and J. R. Whinnery, "Fields and Waves in Modern Radio," John Wiley and Sons, Inc., New York, N. Y., Sec. 9.08; 1944.

[3] C. G. Montgomery, *et al.*, "Principles of Microwave Circuits," Radiation Laboratory Series, vol. 8, chap. 8; McGraw-Hill Book Co., Inc., New York, N. Y.; 1948.

[4] These relations are obtained by duality replacement from the E-type equations given in reference 3.

where

$$Y_\infty'(r) = j\frac{H_0^{(2)}(kr)}{H_1^{(2)}(kr)} = \frac{h_0(x)}{h_1(x)}e^{j[\eta_1(x)-\eta_0(x)]}, \tag{3a}$$

and is the normalized input admittance of an infinite radial line. The differential forms of (1) and (2), which are useful for frequency sensitivity calculations, are obtainable from equations (59) and (58) of footnote reference 3. These expressions enable one to calculate the change in relative input admittance or voltage reflection coefficient due either to a change in relative output admittance or output voltage reflection coefficient, respectively, or to a change in wavelength, or to a simultaneous change in both factors.

The formulas summarized above provide a rigorous description of a symmetric H-type radial transmission line (and by the duality replacement of Y' by Z', etc., also of an E-type radial line). The complexity of calculations with these formulas in certain engineering applications suggests the desirability of employing a simpler, albeit approximate form of radial line description. Such a description is based on the asymptotic identity of the radial and uniform transmission line formulas at large radii. The details of this approximate description, including the estimate of the order of accuracy, are discussed below.

In Section II the input admittance of a typical long radial line (an electromagnetic horn) is computed both rigorously and approximately. In Section III the calculation of frequency sensitivity in long and short radial lines is discussed and an approximate method is developed for long radial lines. By typical calculations the accuracy of this approximation is shown to be quite good. In Section IV the problem of higher mode interaction between discontinuities on radial lines is discussed.

II. The Approximate Admittance Relation

Equation (1) relates the relative input admittance at some point x of a radial line to the relative output admittance at some other point y. This useful expression involves a number of products of Bessel functions and is therefore somewhat complicated. However, as x and y become large, certain of these Bessel function combinations asymptotically approach the trigonometric functions encountered in uniform lines. This behavior in the limit of large argument is to be expected from the plane parallel character at large radii of portions of the cylindrical cross-sectional surfaces of a radial guide.

The fact that for large argument the radial line expression (1) reduces to the uniform line expression enables one to obtain approximately correct values for the relative input admittances of long radial lines. Such approximate admittance calculations are much simpler than the rigorous radial line ones. To enlarge the usefulness of the approximations it is necessary to estimate the error introduced. This estimate of error is obtained from the asymptotic expansions for the Bessel functions.

The rigorous admittance relation (1) is first applied to a radiating sectoral E-plane horn to obtain the dependence of the relative input admittance at the horn throat on the length of the horn for two different horn angles. An approximate admittance relation is then developed, and the approximate results for certain cases are compared with the corresponding rigorous results.

A. Input Admittance of a Waveguide Horn

The horn considered here has an infinite width and is fed by a parallel plate waveguide. This type of horn was chosen because expressions are available for the junction capacity at the horn throat and for the radiative output admittance at the horn mouth. If these two quantities are known for any other type of sectoral horn, radial line theory can also be applied to it.[5] The horn, shown in Fig. 2, is sometimes called an E-plane horn. The dominant mode is the lowest H-type mode; as shown, the electric field lies in the plane of the figure, and the magnetic field perpendicular thereto. The equivalent circuit is shown in Fig. 3.

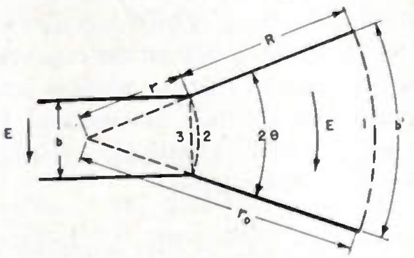


Fig. 2—Geometry of the infinite width sectoral E-plane horn.

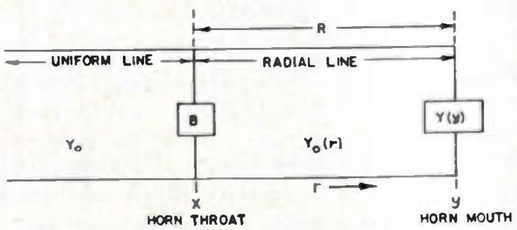


Fig. 3—Equivalent circuit for the horn of Fig. 2.

It is desired to obtain information about the input admittance of the horn as seen at reference plane 3 in the uniform line, i.e., at the end of the parallel plate guide feeding the horn (see Fig. 2). The admittance relation (1) gives the relative input admittance at reference plane 2 in the radial line in terms of the relative output admittance at reference plane 1, the open end of the horn. The relative input admittance at reference plane 3 is related to the relative input admittance at 2 by

$$Y_{in}'(x) = \frac{\sin\theta}{\theta} Y'(x) + j\frac{B}{Y_0} \qquad (4)$$

[5] If the horn is finite instead of infinite, all the transmission line relations used here still apply if κ is substituted for k whenever it appears where

$$\kappa^2 = k^2 - \frac{\pi^2}{a^2}, \qquad k = \frac{2\pi}{\lambda}$$

and a is the horn width.

where

$Y'(x)$ = the relative input admittance at 2 in the radial line

$Y'_{in}(x)$ = the relative input admittance at 3 in the uniform line

Y_0 = the characteristic admittance of the uniform line

B = the junction susceptance at the throat

θ = half the horn angle,

and where $\sin\theta/\theta$ is the ratio of the characteristic admittance of the radial line to that of the uniform line.

Thus, $Y'_{in}(x)$ can be determined from a knowledge of the horn geometry and two additional quantities: the relative output admittance $Y'(y)$, and B/Y_0. The values of $Y'(y)$ for various values of b'/λ (b' is defined in Fig. 2) are obtained from the literature.[6] The calculations made in the handbook are for a parallel plate waveguide but the results apply with sufficient accuracy to the horn mouth. They are given for the range $0<(b'/\lambda)<1$ (in the "Waveguide Handbook," b' is called b). The expression for the junction susceptance at the horn throat has been derived by conformal mapping, and is valid for b/λ small, where b is the height of the parallel-plate guide (see Fig. 2). The expression is

$$\frac{B}{Y_0} = \frac{kb}{\theta}\left\{\ln\frac{\theta}{\sin\theta} + \frac{\theta}{\pi}\left[\Psi\left(\frac{\theta}{\pi}\right) + 0.577\right]\right\}, \qquad (5)$$

where

$$k = \frac{2\pi}{\lambda}, \quad \text{and} \quad \Psi(z) = \frac{d}{dz}\ln(z!)$$

and the other symbols have been defined above.

From the above relations the relative input admittance $Y'_{in}(x)$ in the uniform line has been calculated as a function of the horn length R keeping the horn angle, uniform guide height, and wavelength constant. Results have been obtained for two different horn angles: $22\frac{1}{2}°$ and 45°. The height b of the uniform guide feeding the horn has been chosen to be the same for both sets of calculations, i.e., $\lambda/4$. The results are given graphically in Fig. 4. From the graphs it is seen that as the horn length increases, the relative input admittance, $Y'_{in}(x)$, spirals in towards a certain limiting value; a tighter spiral is obtained for the larger horn angle.

As the horn length increases $Y'_{in}(x)$ approaches the limiting value $Y_\infty'_{in}(x)$, which is the relative input admittance at x in the uniform line for a horn of infinite length. At large horn lengths, higher modes are capable of propagating, although they are rapidly damped out for shorter horn lengths. Thus the effect of any higher modes generated at the horn mouth cannot be felt at the horn throat x (since higher modes cannot be propagated in the immediate neighborhood of x). $Y_\infty'_{in}(x)$ is

[6] N. Marcuvitz, "Waveguide Handbook," Radiation Laboratory Series, vol. 10, McGraw-Hill Book Co., Inc., New York, N. Y., p. 322; 1951.

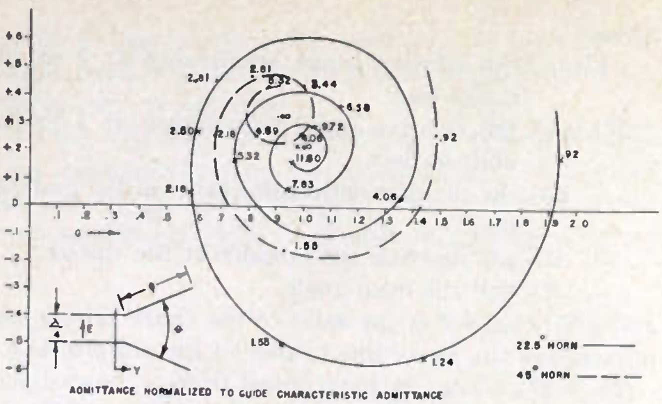


Fig. 4—The relative input admittance in the uniform line feeding the horn as a function of the horn length.

calculated from the relative input admittance, $Y_\infty'(x)$ (see (3)), of an infinite H-type radial line, together with (4) above, which relates $Y_\infty'(x)$ for the radial line to the $Y_\infty'{}_{in}(x)$ for the uniform line. As x increases, the value of $Y_\infty'(x)$ approaches unity. This behavior of an infinite radial line is different from that of a uniform line; in the latter the input admittance of an infinite line is equal to the characteristic admittance of the line, so that the relative input admittance is always unity. The values of $Y_\infty'{}_{in}(x)$ are different for the two horn angles primarily because the values of x (horn throat) are different.

B. The Approximate Admittance Relation

Asymptotic large argument expansions for the Bessel functions are well known.[7] If the first-order terms only in such expansions are employed in the admittance relation (1), the latter reduces to the admittance relation for the uniform line. A better approximation to the relative admittance is obtained on inclusion of the second-order terms in the Bessel function expansions. Let us define the approximate relative admittance thus obtained as $Y_r'(x)$, and the relative admittance obtained by means of the uniform line expression as $Y_u'(x)$. The difference, $\Delta Y'(x)$, defined as

$$\Delta Y'(x) = Y_r'(x) - Y_u'(x) \tag{6}$$

can be used as a fair estimate of the error introduced by using the uniform line instead of the radial line relation.

The insertion of the second-order, large argument expansions of the Bessel functions into (3) leads after some rearrangement to

$$Y_r'(x) = \frac{[j + Y'(y)\cot v] + \left[-Y'(y)\frac{1}{8}\left(\frac{1}{x}+\frac{3}{y}\right) + j\frac{1}{8}\left(\frac{1}{x}-\frac{1}{y}\right)\cot v\right]}{[\cot v + jY'(y)] + \left[\frac{1}{8}\left(\frac{3}{x}+\frac{1}{y}\right) - j\frac{3}{8}\left(\frac{1}{x}-\frac{1}{y}\right)Y'(y)\cot v\right]}, \tag{7}$$

where

$$v = (y - x).$$

Since (57) is in the form

$$Y_r' = \frac{A + \delta}{B + \epsilon},$$

where δ and ϵ are small, Y_r' can be written as:

$$Y_r' = \frac{A}{B} + \frac{1}{B}\left(\delta - \epsilon\frac{A}{B}\right).$$

But A/B is exactly $Y_u'(x)$, hence (7) may be used to express $\Delta Y'(x)$. After algebraic manipulation, one obtains:

$$\Delta Y'(x) = j\frac{1}{8}\left[\frac{1}{x} - \frac{1}{y}\right][1 + 3Y_u'^2(x)] - \frac{1}{2y}\left[\frac{Y_u'(x) + Y'(y)}{\cot(y-x) + jY'(y)}\right]. \tag{8}$$

Equation (8) is valid for arbitrary values of the relative output admittance $Y'(y)$. In certain radial line applications, for example, in horns, the relative output admittance is often close to unity, while in cavity resonators, short circuits are not uncommon. Substituting these special values of $Y'(y)$ into (8), one obtains:
for $Y'(y) = 1$ (matched output)

$$\Delta Y'(x) = j\left[\frac{1}{2x} - \frac{1}{2y}e^{-2j(y-x)}\right]. \tag{9}$$

For $Y'(y) = \infty$ (short circuit)

$$\Delta Y'(x) = \frac{j}{2y} + j\frac{1}{8}\left[\frac{1}{x} - \frac{1}{y}\right][1 - 3\cot^2(y-x)]. \tag{10}$$

In order to test the accuracy of the approximation (8), calculations of the relative input admittance of a waveguide horn were made as a function of horn length using both the radial line and uniform line expressions, and then their difference was computed. In Table I these exact calculations are compared with the approximate values for the difference as computed from from (8). The horn chosen has $b = \lambda/4$, and the horn angle $= 22\frac{1}{2}°$. It may be seen from Table I that the computed $\Delta Y'$ never differs from the exact difference by more than $\pm 0.02 \pm j0.02$.

An investigation of further corrections, i.e., the inclusion of further terms in the asymptotic expressions, is rather complicated for arbitrary output admittances

[7] G. N. Watson, "Theory of Bessel Functions," Cambridge University Press; 1922.

TABLE I

Horn angle $=22\frac{1}{2}°$

$$\frac{b}{\lambda}=\frac{1}{4}$$

$x=kr$	$y=kr_0$	Correct Difference $Y'(x)-Y_u'(x)$	Approximate Difference Using Equation (8)
4.10	4.39	$-0.061-j.041$	$-0.081-j.043$
4.10	5.96	$0.061+j.104$	$0.080+j.098$
4.10	6.91	$-0.008+j.047$	$0.011+j.051$
4.10	10.68	$-0.069+j.082$	$-0.059+j.103$
4.10	15.70	$0.022+j.122$	$0.037+j.120$
8.16	15.70	$-0.023+j.091$	$-0.015+j.096$

but quite simple for output admittances near unity. This latter special case is most common in long line applications. For this case, the difference between $Y'(x)$ and the sum $Y_u'(x)+\Delta Y'(x)$ depends primarily on the value of x, and is larger if x is less. For example, for $x>6.3$ (no matter what the value of y is), this difference is no greater than $\pm 0.01 \pm j0.01$.

The quantity $\Delta Y'(x)$ in (8) may be employed in two ways: first, as an estimate of the error introduced when the uniform line admittance relation is used instead of the radial line expression; second, as an alternative method of calculating $Y'(x) \cong Y_u'(x)+\Delta Y'(x)$. This second use for $\Delta Y'(x)$ does not yield the exact value of $Y'(x)$, but the investigation of the second correction has shown that the error introduced is no greater than $\pm 0.01 \pm j0.01$ when x is greater than 6.3. It should be pointed out that it is unnecessary to compute accurately the correction $\Delta Y'(x)$ in order to obtain high over-all accuracy in $Y'(x)=Y_u'(x)+\Delta Y'(x)$.

III. Frequency Sensitivity

The frequency sensitivity of a microwave structure may be inferred from a knowledge of the frequency derivative of the admittance at some point in the structure. Calculations of this sort are of use in estimating the broad band properties of long radial lines (horns) or in calculating the Q's of short radial lines (cavities). Frequency sensitivity calculations may be classified according to the length of the radial line. For short lines, the applications are such that the output admittance is generally far from a match, so that an admittance formulation, which more easily handles arbitrary output admittances, is preferable. However, for long radial lines, the admittance formulation yields an accurate indication of the variation of admittance with frequency only for infinitesimal frequency shifts. Since, in addition, long line applications generally involve output admittances close to a match, a reflection factor approach is more desirable.

A. Long Lines—Reflection Factor Formulation

It will be convenient to speak in terms of the reflection factor Γ referred to a uniform guide, both because it is physically easier to comprehend, and because in many applications, such as horns or tapered waveguide sections, we are interested in the standing-wave ratio referred to the uniform guide. It is desired to relate changes in the voltage reflection factor $\Gamma(x)$, referred to a uniform guide feeding a radial guide, to changes in the wavelength, the output voltage reflection factor $\Gamma_v(y)$ in the radial guide, or both.

An expression is first obtained relating the voltage reflection factor in the uniform line $\Gamma(x)$ to that in the radial line $\Gamma_v(x)$. The well-known expression (18) of footnote reference 3 for the uniform guide yields the relation between $\Gamma(x)$ and the relative admittance in the uniform guide. Application of (4) then relates $\Gamma(x)$ to the relative admittance $Y'(x)$ in the radial guide. Equation (3) gives $Y'(x)$ in terms of the voltage reflection factor $\Gamma_v(x)$ in the radial guide. Substituting (3) into the expression for $\Gamma(x)$, one obtains the following relation between $\Gamma(x)$ and $\Gamma_v(x)$:

$$\Gamma(x) = \frac{[1-b-aY_\infty'(x)] + \Gamma_v(x)[1-b+aY_\infty'(x)e^{-2j(\eta_1(x)-\eta_0(x))}]}{[1+b+aY_\infty'(x)] + \Gamma_v(x)[1+b-aY_\infty'(x)e^{-2j(\eta_1(x)-\eta_0(x))}]}, \tag{11}$$

where $a=\sin\theta/\theta$, and is usually close to unity, $b=j(B/Y_0)$, and is small, and Y_∞' is given by (3a).

Equation (11) is true in general, regardless of the reflection introduced at the output. However, for lines which are fairly well matched, the above expression can be considerably simplified since then Γ_v is small. Taking this into account, and introducing the symbols

$$\Gamma_\infty(x) = \frac{1-b-aY_\infty'(x)}{1+b+aY_\infty'(x)} \tag{12}$$

$$\Gamma'(x) = \Gamma(x) - \Gamma_\infty(x) \tag{13}$$

applying them to (11), then expressing $\Gamma_v(x)$ in terms of $\Gamma_v(y)$ using (2), and applying simplications permissible for long lines, one obtains finally

$$\Gamma'(x) = \Gamma_v(y)e^{-2j(y-x)}. \tag{14}$$

For frequency sensitivity considerations, it is desirable to form the differential of the logarithm of (14), wherein noting that $\Delta(y-x)=-(y-x)\Delta\lambda/\lambda$, one obtains

$$\frac{\Delta\Gamma'(x)}{\Gamma'(x)} = \frac{\Delta\Gamma_v(y)}{\Gamma_v(y)} + 2j(y-x)\frac{\Delta\lambda}{\lambda}. \tag{15}$$

Equation (15) yields the quantity $\Gamma'(x)$, and the corresponding desired $\Gamma(x)$ is obtained from (13). The procedure followed here is essentially that of transforming to a primed system, where the behavior is similar to that of a uniform line so that the calculations are greatly simplified, and then transforming back to the unprimed system.

The relation between the relative output admittance and the voltage reflection factor is

$$\Gamma_v(y) = \frac{Y_\infty'(y) - Y'(y)}{Y_\infty'^*(y) + Y'(y)}, \tag{16}$$

where $Y_\infty'^*(y)$ is the complex conjugate of $Y_\infty'(y)$. However, for long lines, and where $Y'(y)$ may be written as $Y'(y) = 1+\epsilon$, where ϵ is small, (16) reduces to

$$\Gamma_v(y) \cong -\frac{\epsilon}{2}. \tag{17}$$

In order to determine the range of wavelength shifts for which the above formulation is valid, calculations were made using (15) and (13) and are compared with exact calculations, using (11). The radial guide chosen for the calculations has the following characteristics: The relative output admittance $Y'(y) = 1.050$, so that $\epsilon = 0.050$. For simplicity, it is assumed that $Y'(y)$ remains constant over the wavelength changes applied. At the mean wavelength λ_0 the locations of the horn throat and the horn mouth are given by $x=4$, and $y=40$, respectively. The comparison between the values of $\Gamma(x)$ calculated exactly and calculated according to the above formulation are given in Table II.

TABLE II
$Y'(y) = 1.050$
At $\lambda = \lambda_0$, $y=40$
$x=4$

$\frac{\Delta\lambda}{\lambda}$	$x=kr$	$y=kr_0$	$\Gamma(x)$, exact calculation	$\Gamma(x)$ from frequency sensitivity formula
0.04	3.85	38.46	$0.0650\, e^{j74.8°}$	$0.0640\, e^{j74.2°}$
0	4.00	40.00	$0.0616\, e^{j120.8°}$	
−0.02	4.08	40.81	$0.0846\, e^{j102.3°}$	$0.0850\, e^{j102.0°}$
−0.111	4.44	44.44	$0.0810\, e^{j97.2°}$	$0.0860\, e^{j98.0°}$

Comparison of the last two columns in Table II indicates that even up to an 11 per cent wavelength change, the agreement is very good. It may be seen that for the 11 per cent wavelength change, there is a slight disagreement between the amplitudes of $\Gamma(x)$. The principal reason for this is that $\Gamma_\infty(x)$ really changes slightly, while it was assumed above that it remains constant. If one therefore wishes to extend the range of validity of the above method, he need simply calculate, in using (13), the value of $\Gamma_\infty(x)$ corresponding to the new x, not the x for λ_0.

IV. Higher Mode Interaction Effects Between Discontinuities

The usual admittance analysis in a waveguide region involves only a single mode; however, near a discontinuity additional higher modes exist which usually decay rapidly as they travel away from the discontinuity. It is therefore possible to account for the effect of the discontinuity everywhere except near it by considering only the lowest mode and by replacing the discontinuity by an equivalent lumped circuit. The problem to be treated here is this: How close can two discontinuities be placed while still using their equivalent circuit parameters calculated on the assumption that the obstacles are isolated, i.e., considering only dominant mode interaction?

A discontinuity confined to a single surface transverse to the direction of propagation (in this case a cylinder $r=$const.) e.g., an iris, is perhaps the simplest to analyze. Its equivalent circuit is simply a shunt admittance whose value is given by[8]

$$Y = \sum_n [Y_n^{(1)} + Y_n^{(2)}] V_n^2, \tag{18}$$

where the summation is over all higher modes excited, and where $Y_n^{(1)}$ and $Y_n^{(2)}$ are the admittances associated with the nth mode looking inward and outward, respectively. V_n is the ratio of the voltage amplitude of the nth mode to that of the lowest mode.

It is assumed in (18) that none of the higher modes are propagating at the radius under consideration. Since under most conditions (excluding higher mode resonances) the second discontinuity does not change V_n to the first order, (18) states that the change in Y due to the presence of another discontinuity can be calculated from the knowledge of how the $Y_n^{(1)}$ or the $Y_n^{(2)}$ change, depending on whether the second discontinuity is closer to or further from the origin, respectively. In analogy with uniform lines, it is to be expected that the effect on the input admittance caused by a second discontinuity will grow less as the order of the mode increases. Thus the estimate of the interaction effect on the first signficant higher mode will be an upper bound for the interaction effect on the discontinuity admittance. The problem is thus simplified to: How much does the higher mode input admittance of a given length of radial line depend on its termination?

As an approximation, the Bessel functions are replaced by their small argument values.[9] This condition is analogous to the mode being cut off in a uniform guide. In a radial guide of infinite width any higher mode will behave as a propagating mode far from the origin and as a cutoff mode close to the origin.

The higher mode radial line functions analogous to equations (38) of footnote reference 3 are

$$\zeta_p(x, y) = \frac{J_p(x)N_p(y) - N_p(x)J_p(y)}{J_p'(x)N_p'(y) - N_p'(x)J_p'(y)}, \tag{19a}$$

$$ct(x, y) = \frac{N_p'(x)J_p(y) - J_p'(x)N_p(y)}{J_p(x)N_p(y) - N_p(x)J_p(y)}, \tag{19b}$$

$$Ct(x, y) = \frac{J_p(x)N_p'(y) - N_p(x)J_p'(y)}{J_p'(x)N_p'(y) - N_p'(x)J_p'(y)}. \tag{19c}$$

[8] S. A. Schelkunoff, "Electromagnetic Waves," D. Van Nostrand Co., Inc., New York, N. Y., p. 491; 1943.

[9] E. Jahnke and F. Emde, "Tables of Functions," Dover, New York, N. Y., pp. 128, 132; 1943.

Using the small argument approximations for the Bessel functions, the radial functions (19) become

$$\zeta_p(x, y) = -\frac{xy}{p^2}, \tag{20a}$$

$$ct_p(x, y) = \frac{p}{x}\,\frac{1+\left(\dfrac{x}{y}\right)^{2p}}{1-\left(\dfrac{x}{y}\right)^{2p}}, \tag{20b}$$

$$Ct_p(x, y) = -\frac{x}{p}\,\frac{1+\left(\dfrac{x}{y}\right)^{2p}}{1-\left(\dfrac{x}{y}\right)^{2p}}. \tag{20c}$$

Substituting these into the admittance relation for the higher modes, one obtains

$$B_n(x) = \frac{1 + B_n(y)\left[\dfrac{1+(x/y)^{2p}}{1-(x/y)^{2p}}\right]}{\left[\dfrac{1+(x/y)^{2p}}{1-(x/y)^{2p}}\right] + B_n(y)}, \tag{21}$$

where

$$p = \frac{n\pi}{\phi_0}, \quad \text{and} \quad B_n(z) = -\frac{jp}{z}\,Y_n'(z).$$

In order that $B_n(x)$ be approximately independent of $B_n(y)$ it is necessary that the quantity

$$\left[\frac{1+(x/y)^{2p}}{1-(x/y)^{2p}}\right]$$

be nearly unity, which in turn means that $2(x/y)^{2p}$ must be small compared to unity. Then, assuming $2(x/y)^{2p}$ small, (21) reduces to

$$B_n(x) \cong 1 + 2\left(\frac{x}{y}\right)^{2p} \frac{B_n(y) - 1}{B_n(y) + 1}.$$

Since no definite knowledge of the magnitude of $B_n(y)$ is possible without analyzing the discontinuities in detail, some assumption must be made which will be typical of ordinary discontinuities. Under the restriction that $[B_n(y)-1]/[B_n(y)+1]$ is of the order of one, $2(x/y)^{2p}$ is simply the per unit change in the input susceptance at a radius x of a higher mode radial transmission line due to the presence of a discontinuity at radius y.

The discontinuity admittance is given by (18). The presence of a single neighboring discontinuity will change either $Y_n^{(1)}$ or $Y_n^{(2)}$. The higher the order of the mode the less is the change expected in the Y_n, so that $2(x/y)^{2p}$ is an upper bound on half of the per unit change in Y. Therefore

$$\Delta = \left(\frac{x}{y}\right)^{2\pi n/\phi_0}, \tag{22}$$

where Δ is an approximate upper bound on the per unit change in the equivalent susceptance of a capacitance iris at radius x due to another discontinuity at (larger) radius y, ϕ_0 is the angular opening of the radial line, n is the order of the first higher mode excited by the iris. In most cases $n=1$, except when discontinuity is symmetrical about plane $\phi=\phi_0/2$, in which case $n=2$.

Although the result (22) was obtained for irises, it is probably typical for all discontinuities which excite only the H-type modes. Such discontinuities would include the throat of an E-plane horn, a capacitive iris, or an axial wire or rod. The result also applies to the change in the value of the outer discontinuity parameter, but y must always be the larger radius.

It is convenient to consider Δ as a function of $d=y-x$ and $b'=y\phi_0$ (the arc length at the outer discontinuity in electrical radians). Then (22) becomes:

$$\Delta = \left(1 - \phi_0\frac{d}{b'}\right)^{2\pi n/\phi_0}. \tag{23}$$

Equation (23) is plotted in Fig. 5 for $n=1$ and $\Delta=0.1$, 0.01. As explained above, this holds for asymmetrical discontinuities. For symmetrical discontinuities, one must use the half angle for ϕ_0 instead of the full angle.

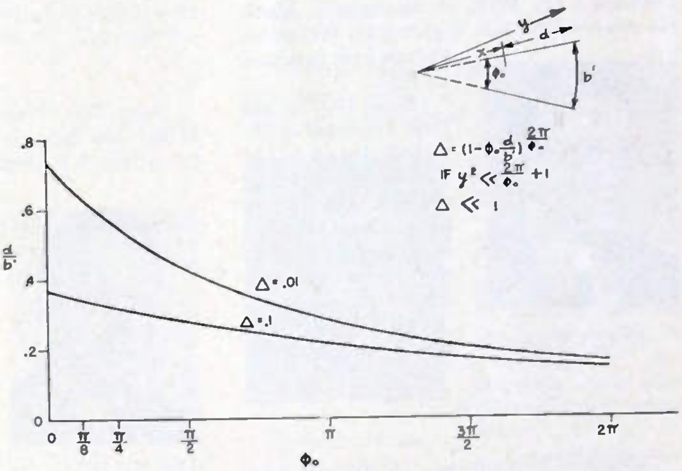


Fig. 5—Graphs of equation (23) for $n=1$.

As an example, the interaction between the discontinuities at the horn mouth and throat will be considered, for the case of the horn shown in Fig. 2. The worst interaction will occur for the largest angle and the shortest horn length, i.e., the 45° horn, of length 0.92 radian, corresponding to the cases considered earlier in this paper. Since the discontinuities are both symmetrical, $\phi_0=\pi/8$, or half of the angular opening. The per unit change in the throat discontinuity susceptance as given by (23) is then $\Delta=0.064$. The discontinuity susceptance, of value 0.24, of the junction (horn throat) between the horn and the uniform waveguide, may therefore be incorrect by as much as 6 per cent for this case. However, the horn input admittance is in error at most by $(0.06)(j.24)=j.015$, compared to the actual admittance value of $1.94+j.18$.

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He served as radar officer with the Signal Corps in England and with the Air Force in India, China, and the Marianas. Upon his return to the United States in 1946, he was assigned to the University of Illinois for postgraduate training in electrical engineering, where he received the M.S. degree in 1947 and the Ph.D. degree in 1949. He is now a major in the USAF.

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J. V. N. Granger

In 1945 he joined the staff of the Central Communications Research Laboratories of Division 13, OSRD, at Harvard, leaving in 1946 to resume his studies. In 1947 he became group leader of the Antenna Group at the Cruft Laboratory. In May, 1949, he was named supervisor of the Aircraft Radiation Systems Laboratory, Stanford Research Institute, Stanford, Calif. He has recently received a part-time appointment as acting associate professor in the electrical engineering department of Stanford University for the Summer Quarter, 1951.

Dr. Granger is a member of the Panel on Antennas and Propagation of the Research and Development Board, the American Physical Society, Sigma Xi, and also of Commission 6 of the U.S.A. National Committee of the URSI.

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For the biography and photograph of Donald B. Harris, please see page 949 of the August, 1950, issue of the Proceedings of the I.R.E.

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Harwick Johnson

From 1938 to 1942 he was a research fellow and teaching assistant in the department of electrical engineering at the University of Wisconsin.

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Rudolf Kompfner was born in Vienna, Austria, on May 16, 1909. He attended the Realschule and Technische Hochschule in Vienna, and was graduated from the faculty of architecture in 1933. In 1934 he went to England to continue his studies in architecture privately, and became the director of a building firm in 1937. He has devoted much of his spare time to the study of television, radio, and physics.

Rudolf Kompfner

Mr. Kompfner entered the Admiralty service in 1941 as temporary experimental

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Arthur E. Laemmel (S'43–A'46) was born in Brooklyn, N. Y., on December 14, 1923. He received the degree of B.E.E. from the Polytechnic Institute of Brooklyn in 1944, and has been associated with the Microwave Research Institute of the Polytechnic Institute since that time.

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From 1935 to 1936 Dr. Marcuvitz was a research fellow at the Polytechnic Institute. From 1936 to 1940, he was employed by the RCA Manufacturing Company, and did work on high-g_m vacuum tubes, iconoscopes, and orthicons. From 1940 to 1941 he held a research fellowship at Polytechnic Institute, and during 1941 to 1946 he did research on waveguides at the MIT Radiation Laboratory.

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Joachim W. Muehlner (M'50) was born on January 10, 1913, in Dresden, Germany. He received the Dipl. Ing. (M.Sc.) degree in 1937 and the Dr. Ing. (Ph.D.) degree in 1939, both from the University of Dresden.

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As research assistant and instructor in electronics at the same institution from 1937 to 1941, he also participated in the development of the guided-missile electronics instrumentation equipment. From 1942 to 1945 he was in charge of the Doppler-type velocity and position measuring instrumentation at Rocket Center Peenemuende.

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For photograph and biography of Tetsu Morita, see page 950 of August, 1950, issue of the Proceedings of the I.R.E.

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O. W. Muckenhirn

As a co-operative student, he worked with the General Electric Company in Schenectady, N. Y., from February to June, 1937. From March to June, 1938, he held an assistantship in electrical engineering at MIT.

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Newton Monk (SM'48) was born in Stoughton, Mass., on December 5, 1897. After an interruption in his college work during World War I, for service in the U. S. Army Signal Corps, he was graduated from Harvard College in 1920 with the degree of A.B. In 1922 he received the degree of B.S. in communication engineering from the Harvard Engineering School.

Newton Monk

Immediately following graduation, Mr. Monk joined the American Telephone and Telegraph Company, where he worked on interference suppression and carrier transmission. In 1934 he transferred to the Bell Telephone Laboratories, where he continued his work on carrier transmission and pioneered in the application of carrier systems to railroad communication lines.

During World War II he was active in the development of voice-frequency and carrier equipment for the Signal Corps. Subsequently, he has been in charge of a group developing mobile radio systems for railroads and other special applications.

Mr. Monk is a member of the American Institute of Electrical Engineers and has been active on several of its committees. He is also affiliated with the Communications Section of the Association of American Railroads, and is a member of the Harvard Engineering Society.

Contributors to the Proceedings of the I.R.E.

Arthur A. Oliner (M'47) was born in Shanghai, China, on March 5, 1921. He received the degrees of B.A. from Brooklyn College in 1941, and Ph.D. in physics from Cornell University in 1946. From 1941 to 1944, he was a graduate-teaching assistant in the physics department at Cornell University, and participated in an Office of Scientific and Research Development project at this institution during 1944–1945, as a research assistant.

A. A. Oliner

Since 1946, Dr. Oliner has been a research associate at the Microwave Research Institute of the Polytechnic Institute of Brooklyn, where he has been engaged in research on waveguides, especially the investigation of slot coupling and radiating elements, and precision measurement methods for waveguide discontinuities. He has also taught graduate courses in physics and electrical engineering.

Dr. Oliner is a member of the American Physical Society and Sigma Xi.

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Carlyle V. Parker (S'36–A'38–SM'47) was born in West Carthage, N. Y., on September 4, 1912. In 1936 he received the degrees of B.S.E., in physics, and B.S.E., in electrical engineering, both from the University of Michigan, where he was elected a member of Phi Eta Sigma, Phi Kappa Phi, and Sigma Xi. Immediately after graduation he joined the technical staff of the Bell Telephone Laboratories, Inc., in the vacuum-tube research department, later transferring to the circuit research department.

Carlyle V. Parker

In March, 1941, Mr. Parker became a contract engineer for the United States Naval Research Laboratory, in Washington, D. C., receiving a Civil Service appointment in October of that year. While at the Laboratory he has been primarily concerned with the development of radar and radar-beacon systems and techniques. He began graduate work in physics at Columbia University in 1936, and has continued his studies at the University of Maryland since 1945.

Mr. Parker is a member of the American Physical Society.

❖

J. L. Potter (A'31–SM'44) was born in Carthage, Mo., on December 4, 1905. He received the B.S., M.S. and E.E. degrees from Kansas State College in 1928, 1930, and 1939, respectively.

J. L. Potter

In 1928 Mr. Potter joined the Western Electric Company in Chicago, working in the Electrical Inspection and Electrical Measurements Laboratory, and returned in 1929 to Kansas State as a graduate assistant. From 1930 to 1939 he was a member of the electrical engineering department at the University of Iowa. In 1939 he joined the electrical engineering department at Rutgers University, and at present he is professor and chairman of the department. Various industrial and consulting work was performed during the summer months. He is the author of several technical papers, and has been granted patents on a sweep circuit and amplifying systems.

Mr. Potter is a member of the American Institute of Electrical Engineers, National Society of Professional Engineers, American Society of Engineering Education, Sigma Xi, Tau Beta Pi, Sigma Tau, Eta Kappa Nu, and Phi Kappa Phi.

❖

F. N. H. Robinson was born in April, 1925, in England. He received his education at the Leys School and later at Christ's College, in Cambridge, where he was graduated in 1945 with the degree in physics.

F. N. H. Robinson

After leaving Cambridge, Mr. Robinson joined the Royal Naval Scientific Service and was associated with the S.E.R.L. Baldock company, devoting his time to the field of microwave electronics.

Mr. Robinson is at present engaged in low-temperature research at the Clarendon Laboratory, in Oxford, England.

Louis A. Rosenthal (S'43–A'45) was born in New York, N. Y., on August 16, 1922. He received the B.E.E. degree from the College of the City of of New York in 1943, and the M.E.E. degree from Brooklyn Polytechnic Institute in 1947. Upon graduation he worked for the Star Electric Motor Company until January, 1944, at which time he went to Lehigh University to teach in the A.S.T.P. program.

L. A. Rosenthal

In April, 1944, he moved to Rutgers University, where he is at present assistant professor of electrical engineering.

Mr. Rosenthal has been a part-time consultant for the Bakelite Corporation, Naval Ordnance Laboratories, and the Markite Co. For the past four years he has been engaged in half-time government sponsored research and half-time teaching.

Mr. Rosenthal is a member of Eta Kappa Nu, Sigma Xi and the American Institute of Electrical Engineers.

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Walter Rotman (A'49) was born in St. Louis, Mo., on August 24, 1922. He received the B.S. and M.S. degrees in electrical engineering from Massachusetts Institute of Technology in 1948. From 1942 to 1945 he was associated with the Army Air Forces, working on radar. During his student career he served as a research assistant in the Magnetron Group of the Research Laboratory of Electronics at MIT, from 1947 to 1948.

Walter Rotman

In 1948, Mr. Rotman joined the Airborne Antenna Group of the Air Force Cambridge Research Laboratories, where he is now engaged in antenna research. His chief field of interest has been the investigation of progressive wave antennas.

He is an associate member of Sigma Xi.

Correspondence

More About Telepathic Communication*

R. J. Bibbero's[1] entertainingly speculative comments on the telepathy sense lead to the same bewilderment most investigators feel when they attempt to account for observed extrasensory-perception phenomena. Perhaps some clues may be found in the incredible sensitivity of the obscure senses possessed by some animals (and humans) and in the discovery in insects of infrared senses which strangely correspond with Bibbero's belief in the existence of hyperfrequency effects.

Doctor Tromp[2] found that dowsers actually locate underground water and bodies of metal by virtue of their feeble diamagnetic and ferromagnetic fields. Some "healers" can also detect sick areas in a body by sensing the tiny differences in body electrical potentials over such areas. This was verified by the electrocardiogram and by body potential checks.

Professor Yaegley of the Pennsylvania State College and other investigators confused homing birds by attaching magnets to their wings and by releasing them near radio and radar transmitters.

Donald Fink mentioned the fact that Professor Miles of Yale University had discovered that insects smell by means of infrared senses and that 7 male moths released from a moving train over a 1-mile stretch found, within hours, a female moth sealed in an hermetic container. This was confirmed by J. P. Duane and J. E. Tyler of Interchemical when they found that a female moth's temperature was some 11° higher than that of her surroundings and that she emitted infrared radiations about 8 microns in wavelength. The male moth's antenna hairs were 40 to 80 microns long, and they varied in 4 micron steps, or half the female moth's main infrared emission.[3]

Doctor Rajewsky, a German biologist, has the rather unusual theory that the human brain not only employs electrical impulses to control the body, but also radio waves, with individual cells acting as miniature oscillators.[4]

Recently the author came across the case of a parapsychic who is able to see, with his eyes closed in a dark room, faint, pulsating, Doppler-type shadows whenever trains, automobiles, or planes pass by.

Referring again to Bibbero's quantitative speculations,[1] what is the signal emf induced in the magnetic sensing centers behind the temples of a dowser by the very weak diamagnetic field of a stream of water 50-feet underground? Dowsers have successfully located infinitesimal magnetic fields in the laboratory while medical instruments verified their "pointing." What is a female moth's radiated infrared field strength at a radius of 1 mile? All 7 of the male moths located the female moth in a few hours. What is the signal emf induced in a migrating bird's magnetic sensing center by the earth's feeble magnetic field? According to recent observations birds successfully travel thousands of miles over unfamiliar territory partly by way of the magnetic field and induced microvolt nerve potentials.

If we turn to the more prosaic animal senses, we find ourselves similarly impressed with their sensitivity potentials. In hearing, for instance, incredibly minute amount of acoustic energy will activate animal eardrums. If one sits quietly in an anechoic sound chamber, a rushing sound—caused either by the bloodstream or by air-molecule bombardment—can be heard. How many micro-micro-microwatts does an air molecule develop in bouncing off an eardrum surface? A keen-eyed individual can look up into a cloudless sky and see the Brownian movement of molecules and bacteria on the liquid surfaces of his eyeball, or look up into a clear night sky and see stars which are millions of light years away. What is the light energy in micro-microwatts of such optical images measured at the retina? Certain gases, in concentrations of the order of one part per hundreds of millions, can be detected by the relatively insensitive human nose and certain chemicals, in similar dilutions, can be tasted in water solutions. When one considers the far greater sensitivity of an eagle's eyes, a bloodhound's nose, a deer's ears, or a cat's taste buds, parapsychic phenomena appear somewhat less miraculous. Unusual radiowave reception via duct and multiple-reflection effects is another interesting possibility.[5-7] The writer has just completed some experiments on teleradiasthesia (dowsing over photos, letters, objects, etc.,) with some incredible results which indicate some sort of infrared smell-sense and electromagnetic-wave effects which respond to routine antenna-array techniques applied to hyperfrequency radar equipment. The results have been submitted to Dr. Tromp for laboratory checks and, if verified, pose some significant possibilities. Recent researches by Dr. Tromp also point to some sort of a transponder effect in some dowsers.

Perhaps telepathy tests conducted with the aid of paraboloid mirrors, waveguides, infrared shields, coil loops, antenna, and the like will uncover some interesting data. For all his scientific progress man still has a great deal to learn about man.

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* Received by the Institute, April 2, 1951.

[1] R. J. Bibbero, "Telepathic Communication," PROC. I.R.E., vol. 39, pp. 290–291; March, 1951.

[2] S. W. Tromp, "Physical Physics," Elsevier Publishing Co., New York, N. Y., and Holland.

[3] D. G. Fink, "Cross Talk," *Electronics*, vol. 21, p. 71; February, 1948.

[4] R. T. Lustig, "Human Radio Waves?," *Science Digest*; July, 1948.

[5] "Science," *Time*, September 25, 1950.

[6] F. W. Nordsiek, "Man's five senses now fifteen plus," *Science Digest*, February, 1951.

[7] U. S. Army Signal Corps, "Radio Wave Confuse Carrier Pigeons, *Science Digest*; November, 1945.

The Value of Abstracting Services*

Many radio engineers and scientists must have read with interest Mr. Coile's informative article, "Periodical Literature for Electronic Engineers,"[1] in the December, 1950, issue of the PROCEEDINGS OF THE I.R.E. The statistical results that he gives should be considered very seriously by all workers in the radio field. But we must be careful not to be dazzled by mere numbers in scientific and technical literature; quality as well as quantity is an important factor in the growth of the world's scientific knowledge.

New knowledge does not necessarily increase in direct proportion to the number of papers published any more than is the total quantity of news issued at any time directly related to the number of daily newspapers published. Furthermore, individuals differ in the manner in which they desire to acquire their knowledge. Just as some read a daily and others a weekly newspaper so do scientists and engineers have their individual preferences for periodicals of different types, many of which convey to them substantially the same information on the progress being made in the field in which they are interested. Since we are all overwhelmed by the enormous output of published literature in scientific subjects, it behooves us all the more urgently to sift this material in a manner which is helpful to the reader who cannot possibly manage to look through it all.

In his paper Mr. Coile refers to the apparent limitations of the abstracting service which is provided by the Department of Scientific and Industrial Research and which results in "Abstracts and References," published in the PROCEEDINGS OF THE I.R.E. in the United States and in the *Wireless Engineer* in England.[1]

Since Mr. Coile likes numbers, I may say that our abstracting staff look through considerably more journals than those listed in his paper. For example, during 1948 the total number of journals perused was approximately 360 although the number from which the abstracts were made was 195, as stated by Mr. Coile. This latter figure is in itself some 50 per cent greater than that mentioned (134) by the IRE committees in their annual reviews of progress.

The head of an abstracting service has to consider from time to time how best to use whatever space is allotted to him for abstracts, especially when these are printed in journals whose success depends, to a large extent, on the publication of original scientific and technical papers. The policy adopted in recent years has been generally to give precedence to abstracts of important papers containing original material, to reduce the average length of abstracts, to avoid the duplication of references to articles

* Received by the Institute April 2, 1951.

[1] R. C. Coile, "Periodical literature for electronic engineers," PROC. I.R.E., vol. 38, pp. 1380–1382; December, 1950.

Correspondence

containing essentially the same material and particularly of book reviews in different periodicals, and to abstract serial articles as a whole rather than in their individual parts. An abstract service should effectively provide the user with a key to the information contained in the subject. The staff available for abstracting services are bound to be handicapped by inadequate training and lack of experience in the fields of modern scientific research and applications. This limitation, as well as a shortage of paper, makes a too comprehensive abstracting service impracticable. Instead, it should be a critically selective service, drawing attention to the most important publications as soon as these become available.

It is seriously suggested that the present scope of the "Abstracts and References," as published in the PROCEEDINGS, represents a reasonable compromise between a mere catalogue of references to every paper published in every journal and a series of reviews of progress in special fields of the type now being published from time to time in various countries.

I am indebted to Mr. J. W. Head, who was formerly in charge of the abstracting section of the Radio Research Organization, and to Mrs. D. Loman, who has succeeded him, for some of the information and opinions expressed in this letter.

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Polarization Errors of Radio Direction Finders, A Proposed Classification*

INTRODUCTION

Radio direction finders and rotating-pattern navigation systems are generally designed to receive and transmit waves of a single linear polarization. For example, all high-frequency direction finders in use today are designed to receive vertically polarized waves and vhf omniranges are designed to transmit horizontally polarized waves. Variable errors of indication are obtained with both receiving direction finders and navigation systems whenever the polarization of the received wave changes. The terms "night effect," "airplane effect," "polarization effect," and more commonly, simply "polarization error" have been used in discussing the bearing changes and errors which accompany changes in the polarization of a received wave.

A direction-finding system designed for vertically polarized waves will, in general, produce either no bearing at all or an incorrect bearing, when the received wave is purely horizontally polarized. The fault, however, resides in the direction finder, and not in the horizontally polarized wave component. The vertically polarized wave component is "wanted" and the horizontally polarized component is "unwanted," simply because the direction finder is calibrated or oriented for vertical polarization, and does not respond with equal amplitude and with a consistent directivity to all polarizations.

A received wave component is always "wanted" if the direction-finder antenna responds strongly to it with a directivity pattern oriented in agreement with the initial scale calibration of the indicator. Any wave component which excites either a weak response or which excites a differently oriented response is likely to produce errors and is therefore "unwanted."

PROPOSED CLASSIFICATION

The classification of polarization errors proposed here is based on the thesis that all polarization errors result from instrumental faults or deficiencies, and that a wave component is "unwanted" only when the instrument either exhibits no response at all, or when it exhibits an incorrectly oriented pattern of directivity.

Errors of the first proposed class result from an inherent strong response to an "unwanted" wave component.

Errors of the second proposed class result because the instrument fails to respond to a primary wave component and takes a bearing on a parasitic re-radiator instead of on the true source.

A. Primary Instrumental Polarization Errors

A direction finder with either a loop antenna or an elevated U-Adcock antenna will exhibit a 90-degree bearing error if the received wave is purely horizontally polarized. (Ideally, the loop will produce no bearing at all unless the horizontally polarized wave has a down-coming component.) If the wave has mixed polarization, the bearings may be definite or indefinite and the error may vary between zero and 90 degrees.

The polarization error in this case is due to an inherent response of the antenna to an "unwanted" primary component of the received wave, and exists regardless of the excellence of the instrument and of the site on which it is installed.

It is proposed that such an error be called a "primary instrumental polarization error."

B. Secondary Polarization Errors

A vertically polarized spaced-collector direction finder having a low inherent primary polarization error will sometimes exhibit large errors when the received wave is strong and almost purely horizontally polarized. This situation arises from the presence at the direction finder of vertically polarized secondary fields.

A secondary field might, for example, be received from an oblique conductor in the vicinity which has abstracted energy from the strong horizontally polarized primary component and re-radiated a vertically polarized secondary component. The bearing errors indicated when such a secondary field is received might be termed "secondary polarization errors," or "polarization-sensitive site errors."

The conductors forming part of a direction finder may also serve as parasitic sources of secondary fields which produce varying errors as the polarization of the primary field changes. The latter errors are due to instrumental faults which could be corrected and might be termed "secondary instrumental polarization errors."

The importance of secondary polarization errors is very great in the case of both receiving direction finders and navigation systems. This importance comes about not because the secondary fields in themselves are of large magnitude, but rather because the response of the receiving antenna to the primary field is either zero or very small. This suggests that the proper approach to a solution of the secondary polarization error problem is to insure that the receiving antenna is capable of responding strongly and with correctly oriented directivity to the primary field, regardless of its polarization. When the response to the primary wave is large, "unwanted pickup" of the secondary field is usually too small by comparison to affect accuracy of direction finding.

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* Received by the Institute, April 10, 1951.

A Short Proof that an Isotropic Antenna Is Impossible*

Brouwer[1] has proved that for the usual spherical coordinate system it is impossible to find a vector function $\boldsymbol{A}(\theta, \phi)$ which is a continuous function of (θ, ϕ), and has the properties that $\boldsymbol{A} \cdot \boldsymbol{r} = 0$ and $\boldsymbol{A} \neq 0$ for all values of (θ, ϕ).

Consider a radiating antenna located at the origin of a spherical co-ordinate system. At a great distance from the antenna the electric field intensity due to the antenna can be expressed in the form

$$\boldsymbol{E} = \frac{1}{r}\left[E_\theta(\theta, \phi)\sin(\omega t - \beta r + \gamma)\boldsymbol{u}_\theta + E_\phi(\theta, \phi)\sin(\omega t - \beta r + \delta)\boldsymbol{u}_\phi\right],$$

where $\boldsymbol{u}_\theta$ and $\boldsymbol{u}_\phi$ are unit vectors in the θ and ϕ directions, respectively, and $\beta = 2\pi/\lambda$. According to the above theorem $E_\theta(\theta, \phi)$ must be zero for some value (θ_0, ϕ_0) of (θ, ϕ) since $\sin(\omega t - \beta r + \gamma) \not\equiv 0$. Consequently there is no radiation in the (θ_0, ϕ_0) direction with θ-polarization. Thus an antenna that radiates uniformly in every direction and polarization is impossible. Since a radiating isotropic antenna is impossible, a receiving isotropic antenna is also impossible according to the reciprocity theorem for antennas.

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* Received by the Institute, April 30, 1951.

[1] L. E. J. Brouwer, "Over continue vectordistributies op oppervlakken," *Proc., Koninklijke Akademie van Wetenschappen* (Amsterdam), vol. 11, pp. 850–858; 1909.

Institute News and Radio Notes

Technical Committee Notes

The **Standards Committee** convened on May 11, under the Chairmanship of Axel G. Jensen. This was the first meeting for the period May 1, 1951–April 30, 1952, called by the new Chairman, Mr. Jensen. For the benefit of those who were not familiar with the functions of the Standards Committee, the Chairman explained that it is the task of this Committee to prepared standards on methods of measurement and testing, as well as to write definitions of terms. The scopes of all technical committees were reviewed to ascertain if these were still adequate, or if changes were necessary. Co-ordination in the preparation of IRE definitions was discussed. Allen F. Pomeroy was assigned the task of drawing up suitable instructions for the preparation of definitions for inclusion in the revised Standards Committee Manual.

The Executive Committee of the Institute approved the **Standard on Transducer Terms, 1951,** prepared by a task group of the Standards Committee, and appearing in this issue of the PROCEEDINGS. Reprints may be purchased shortly from Headquarters for a nominal charge.

The **AIEE/IRE Committee on High-Frequency Measurements** will sponsor a joint session on High-Frequency Measurements as part of the 1951 National Electronics Conference in Chicago in October of this year. This Committee has been enlarged with more members on both the IRE and AIEE sides, and each society having a chairman of its respective representation on the committee. E. P. Felch is Chairman of the Joint Committee, E. I. Green is Chairman of IRE representation.

Peter S. Christaldi will serve as Vice-Chairman of the **Measurements and Instrumentation Committee** for the term ending April 30, 1952.

A meeting of the **Piezoelectric Crystals Committee** was held on April 30, under the Chairmanship of R. A. Sykes.

The **Circuits Committee** under the Chairmanship of C. H. Page held a meeting on May 16. This Committee is reviewing definitions related to linear active circuits.

A meeting of the **Wave Propagation Committee** was held on May 22, H. G. Booker, Chairman, presiding. Much work has been accomplished in the various subcommittees and will be submitted to the Standards Committee for approval. Work on the preparation of Definitions of Radio Astronomy has been completed and will be submitted to the Standards Committee for approval. This Committee has also prepared a bibliography entitled "Tropospheric Propagation: A Selective Guide to Literature," which will also be submitted for the approval of the Standards Committee.

The **Committee on Electron Tubes and Solid-State Devices** held a meeting on April 20, under the Chairmanship of L. S. Nergaard. The possibility of forming a Professional Group on Electron Tubes is under consideration by this Committee. The Committee has been engaged in the preparation of Magnetron Definitions, Radiation Counter Tube Definitions and Methods of Test of Gas-Filled Radiation Counter Tubes. A **Joint Conference on Electron Devices** sponsored by this Committee and the AIEE Committee on Electronics, was held on June 20 and 21 at the University of New Hampshire. This is the first time the AIEE Committee has participated in this annual event which proved to be a most successful affair. Sessions were well attended by an enthusiastic audience. H. J. Reich, Chairman of the Planning Committee, and his efficient staff of committee members, are to be commended for putting on such an excellent Conference.

The **Committee on Antennas and Waveguides** convened on June 5, A. G. Fox, Chairman, presiding. Its Subcommittee on Waveguide Definitions has been reactivated under the Chairmanship of P. H. Smith for the purpose of determining what Technical Committees were interested in the waveguide terms adopted thus far.

The **Video Techniques Committee** held a meeting on June 7, under the Chairmanship of W. J. Poch. In the past year this Committee prepared three standards and two tutorial papers on video records for publication in PROCEEDINGS. Subcommittees were organized and their scopes reviewed.

Emporium Section of IRE to Hold Summer Seminar

The Emporium Section of the IRE will hold its 12th Annual Summer Seminar on August 24 and 25, 1951. This popular affair will include a well-rounded series of papers on Military Electronics, with particular stress on Reliability.

The technical session begins on Friday evening, August 24, with two papers on subminiaturization, starting at 7:30 P.M. The Saturday morning program starts at 9:30 A.M. with two papers scheduled, and closes just before noon.

The social session will be held Saturday at 2:30 P.M. with a picnic at Erskine's Grove. Busses will leave the Sylvania Club at 2:15 P.M. The picnic committee will arrange an all-around program of trapshooting, volley ball, softball, cards, refreshments, and a picnic dinner. Picnic tickets will be $1.50.

Members and guests from out of town should get in touch with Art J. Schultz, of Sylvania's commercial engineering section at Emporium, for picnic tickets and room reservations.

Physicists From Europe to Conduct Nuclear Symposium

Six of the distinguished European physicists who are coming to this country in September will conduct a two-day symposium at Oak Ridge, Tenn., on September 13 and 14 on the general subject, "Nuclear Physics in Europe."

Symposium leaders and their subjects are as follows: Prof. E. Amaldi, University of Rome—"Contribution to the Investigation of Extensive Showers"; Prof. J. Rotblatt, University of London—"Studies of Nuclear Processes Involving 8 Mev Deuterons by the Photographic Method"; Prof. R. E. Peierls, University of Birmingham—"Problems of Nuclear Forces and Nuclear Structure"; Prof. S. Devons, Imperial College of Science and Technology, London—"Projected Experiments on Gamma Emission"; Prof. P. Huber, University of Basel—"Elastic Scattering of Fast Neutrons from Various Light Elements"; Prof. J. Mattauch, University of Bern—"Problems of Modern Mass Spectrometry."

The Oak Ridge symposium is being given under the joint sponsorship of the Oak Ridge National Laboratory and the Oak Ridge Institute of Nuclear Studies. It is unrestricted in nature, and all interested physicists and others are invited to attend.

The symposium will be offered without charge, although those attending will be expected to pay their own living and traveling expenses.

Additional information may be obtained from the University Relations Division, Oak Ridge Institute of Nuclear Studies, P.O. Box 117, Oak Ridge, Tenn.

Professional Group Notes

The petition to form an **IRE Professional Group on Engineering Management** has been approved by the Institute. The following members will serve on the Group's Administrative Committee: Ralph I. Cole, Acting Chairman; W. J. Barkley, C. J. Breitwieser, D. A. Deisinger, Albert D. Emurian, Loren F. Jones, Allan A. Kunze, F. W. Schor, and Tom C. Rives. The Group's purpose is to disseminate information on, as well as promote discussion of those problems confronting engineering managers that especially require the combination of engineering and management know-how in the direction of engineering projects and programs. Anyone interested in joining the Group may obtain application forms by writing to IRE Headquarters.

The **IRE Professional Group on Airborne Electronics**, recently formed, joined with the Dayton Section of IRE in sponsoring the National Conference on Airborne Electronics on May 23, 24, and 25, in Dayton, Ohio. On the first evening of the Conference, the Group held a symposium with J. F. Byrne of Motorola, Inc., serving as moderator, at which representatives of the aviation and electronics industries, together with those of the Armed Services, met to discuss mutual problems. On the following day, the Group Administrative Committee held a business meeting at which IRE President I. S. Coggeshall spoke. On the final day, the first formal meeting of Group members was held at a special luncheon. The meeting was addressed by W. R. G. Baker, Chairman of the Committee on Professional Groups, and R. M. Ashby, chief of the electronics section, North American Aviation Company. During the luncheon the officers of the Group were announced as follows: John E. Keto, Chairman; George Rappaport, Vice-Chairman; and Joseph General, Secretary-Treasurer. It has been decided by the Administrative Committee that the scope of the Group will embrace airborne as well as associated ground electronics equipment for aircraft and guided missiles. A Constitution and Bylaws for the Group have been drawn up and submitted to the Committee on Professional Groups for approval.

First Meeting of the Airborne Electronics Group

At head table (left to right): Gilbert Arenstein, Group Administrative Committee; John Reid, Director, Region 5; Charles Marshall, Chairman, Dayton Section; I. S. Coggeshall, President, IRE; John E. Keto, Group Chairman; R. M. Ashby, North American Aviation Co.; W. R. G. Baker, Chairman, Professional Group Committee; George Rappaport, Group Chairman; Peter Sandretto, Federal Telecommunications Labs.; L. G. Cumming, Technical Secretary, IRE; J. F. Heyd, Group Administrative Committee; and Joseph General, Group Secretary-Treasurer.

As reported in these Notes last month, the **IRE Professional Group on Antennas and Propagation** will sponsor three sessions at the IRE Western Convention. In addition, it is planning to sponsor a technical meeting jointly with the USA National Committee of the International Scientific Radio Union (URSI). This meeting will be held on October 8, 9 and 10, 1951, on the campus of Cornell University. Subjects to be covered at the October meeting will be: "Radio Standards and Methods of Measurement," "Tropospheric Radio Propagation," "Ionospheric Radio Propagation," "Extraterrestrial Radio Noise," and "Electronics."

The **IRE Professional Group on Audio** is currently making plans to sponsor a panel on acoustics at the RTMA Radio Fall Meeting to be held in Toronto on October 29, 30 and 31, 1951. The Group, having already published six newsletters and three separate technical papers, is making definite plans for future publications. A suggested format for the bimonthly newsletters has been drawn up, and a program for expenditure of money collected by the Group and matched by IRE has been estimated. Members will shortly be asked to pay $2.00 per year to cover these services. The local Group in Boston has recently held two technical meetings, on May 3 and on June 7, at the Massachusetts Institute of Technology. On May 3, a paper on "Trends in the Field of Magnetic Recording" was presented by S. J. Begun of the Brush Development Company and on June 7 a paper entitled, "Subharmonic Generation in Loud-Speakers" was read by W. J. Cunningham of Yale University. William G. Burt, Jr. presided at both meetings which were attended by approximately 100 people.

The Administrative Committee of the **IRE Professional Group on Broadcast Transmission Systems** has drawn up a Constitution which has been circulated to the Committee on Professional Groups for approval. Bylaws to the Constitution have also been drawn up. Lewis Winner, Chairman of the Group for the past year, has been elected to continue as Chairman for the year 1951–1952. Ballots for election of new members to the Administrative Committee have been mailed to all members of the Group.

Medical Electronics Group To Be Formed

A petition for the formation of a Professional Group on Medical Electronics is being prepared. The purpose of such a Group would be to make up a forum for those persons who are interested in the application of electronic techniques to the problems of the medical profession, including research. It is requested that all IRE members interested in this field send their names to L. H. Montgomery, Dept. of Anatomy, School of Medicine, Vanderbilt University, Nashville, Tenn., for inclusion on the list of prospective Group members. Those so listed will be notified by mail as to the progress of the Group.

A Constitution for the **IRE Professional Group on Circuit Theory** has been approved by the Committee on Professional Groups and the Institute's Executive Committee. A West Coast Committee has been formed under the Chairmanship of Joseph M. Pettit and the Committee has given assistance to the IRE Western Convention Committee in the organization of a session on Circuit Theory.

The Administrative Committee of the **IRE Professional Group on Industrial Electronics** held a business meeting in Cleveland on May 22 in connection with the Industrial Electronics Conference in which the Group participated. Eugene Mittelmann has been elected Chairman of the Group and George P. Bosomworth has been elected Vice-Chairman. Carl E. Smith will serve as Secretary-Treasurer. A Constitution has been drafted and mailed to members of the Administrative Committee for approval.

A local Group of the **IRE Professional Group on Information Theory** is being organized in the Los Angeles Section under the guidance of DeForest L. Trautman. All those specifically interested in Circuit Theory have been invited to take part in this local Group activity until demand might indicate an additional and separate local Group on Circuit Theory. An organizational meeting was held on June 20 at the University of California.

In addition to participating in the Sixth National Instrument Conference in Houston next fall and in the 1951 National Electronics Conference in October as reported in these Notes previously, the **IRE Professional Group on Instrumentation** is considering the co-sponsorship of a second Conference on Improved Quality Components to be held in Washington, D. C., in the spring of 1952. It has been suggested that the Conference be sponsored by the IRE, the AIEE and the RTMA as was the case last year. The Group has procured a paper by D. A. Alsberg entitled "A Precise Frequency Method of Vector Impedance" which is scheduled for eventual publication in the PROCEEDINGS.

Ballots to elect new members of the Administrative Committee of the **IRE Professional Group on Nuclear Science** have been mailed to the entire membership. The Group is planning a technical meeting to be held jointly with the Atomic Energy Commission during the coming winter. The Conference will likely take place at one of the AEC Laboratories. L. R. Hafstad, a member of the Group's Administrative Committee and Director of the Reactor Development Division of the AEC, has been designated the AEC Representative, Professional Group.

A Constitution for the **IRE Professional Group on Quality Control** has been approved by the Committee on Professional Groups and the Institute's Executive Committee. J. R. Steen, a member of the Group's Administrative Committee, is making arrangements for the Group to sponsor a session at the RTMA Radio Fall Meeting in Toronto next October, and R. M. Krueger of the Administrative Committee is making arrangements for the Group to sponsor a session at the National Electronics Conference in the fall. Chairman R. F. Rollman and L. Bass are making plans for a joint meeting to be held at the National AIEE and ASQC Meeting in Syracuse in 1952.

A paper by Harry H. Goode entitled, "Simulation—Its Place in System Design," which was presented as part of the program of the **IRE Professional Group on Radio Telemetry and Remote Control** during the 1951 National Convention, has been submitted for publication by the Group and is scheduled to appear in a forthcoming issue of the PROCEEDINGS.

R. V. Dondanville is making plans to organize a local Group of the **IRE Professional Group on Vehicular Communications** in Chicago. If this local Group is established in sufficient time, the national technical meeting of the Group will be held in Chicago this fall. A paper by James A. Craig, entitled, "Notes on an Automatic Radio Frequency Repeater" has been procured by the Group and is scheduled for publication in the PROCEEDINGS.

DAYTON AIRBORNE ELECTRONICS CONFERENCE DRAWS OVER 1,400

Over 1,400 engineers and scientists visited the many exhibits and attended lectures at the highly successful IRE National Conference on Airborne Electronics held in Dayton, Ohio, on May 23, 24, and 25, 1951.

Co-sponsored by the Dayton Section and the newly formed Professional Group on Airborne Electronics, the conference was truly national in scope with participants and visitors from almost every state attending. Every major electronics and aviation firm was represented.

One of the many highlights of the conference was a symposium where representatives of both the aviation and electronics industries, together with those of the Armed Services, met in open forum to discuss their mutual problems. The results fulfilled all expectations in that facts were brought out in the symposium which helped both industries to understand each other's problems, as well as the needs of the Armed Services.

Leading engineers in the radio-electronic field presented over 70 technical papers on electronic equipment, electronic instrumentation, vacuum tubes, shock and vibration, microwaves, measurements, circuits, and systems analysis. These papers covered the very latest developments as applied to airborne electronics.

The high point of the Conference was the annual banquet where an award was made to Maj. Gen. H. M. McClelland, Director, Communications-Electronics, Joint Chiefs of Staff, as a "Pioneer in Airborne Electronics." H. B. Richmond, Chairman of the Board, General Radio Company, delivered the principal speech. His topic was "Diplomacy and the Changing Radio Relationships between the United States and Europe." G. L. Haller, Dean, School of Chemistry-Physics, Pennsylvania State College, was toastmaster. These talks were well received by the highly appreciative capacity audience.

AIRBORNE ELECTRONICS AWARD

Maj. Gen. H. M. McClelland, "Pioneer in Airborne Electronics," presented with a scroll by Dr. George L. Haller (right) during Airborne Electronics Conference as IRE President I. S. Coggeshall looks on.

The first formal meeting of the Professional Group on Airborne Electronics was held at the Friday luncheon where the group was addressed by W. R. G. Baker, Chairman, Professional Groups Committee, and R. M. Ashby, Chief, Electronics Section, North American Aviation Company. The officers of the new group were announced during the meeting. These were John E. Keto, Chairman; George Rappaport, Vice-Chairman; and Joseph General, Secretary-Treasurer. The tremendous interest of the conference attendees in the new Professional Group was evidenced by their gratifying turnout for the initial luncheon meeting.

A full ladies program was carried out, including trips to the Dayton Art Institute, National Cash Register Company, a get-together tea, and a luncheon-fashion show.

The Conference Committee wishes to thank the exhibitors, advertisers, speakers and attendees for making the 1951 Conference on Airborne Electronics such a resounding success!

Calendar of COMING EVENTS

AIEE Pacific General Meeting, Portland, Ore. August 20–23

1951 IRE Western Convention, San Francisco, Calif., August 22–24

Instrument Society of America Meeting (IRE Instrumentation group participating), Houston, Texas, September 10–14

4th Conference on Gaseous Electronics, General Electric Research Laboratory, Schenectady, N. Y., October 4–6

URSI-IRE (Antennas and Propagation Group) meeting, Cornell University, Ithaca, N. Y., October 8–10

1951 National Electronics Conference, Edgewater Beach Hotel, Chicago, Ill., October 22–24

AIEE Fall General Meeting, Cleveland, Ohio, October 22–26

Optical Society of America 36th Annual Meeting, Hotel Sherman, Chicago, Ill., October 23–24

Radio Fall Meeting, King Edward Hotel, Toronto, Ont., Canada, October 29–31

Joint IRE/AIEE Computer Conference, Benjamin Franklin Hotel, Philadelphia, Pa., December 10–12

1952 IRE National Convention Waldorf-Astoria Hotel and Grand Central Palace, New York, N. Y., March 3–6

NBS ANNOUNCES NEW CALIBRATION SERVICE

To aid in determining the properties of dielectrics and their dependence on frequency, temperature, and humidity, the National Bureau of Standards has established radio-frequency standards for dielectric measurements. For solid dielectric specimens, dielectric constant and power factor calibration services are now available in the frequency range from 10 kc to approximately 600 mc. Somewhat more limited calibration services are also offered for gases and liquids.

NEW TULSA SECTION AND ROME SUBSECTION FORMED

At the May meetings of the IRE Executive Committee and Board of Directors, approval was granted for the establishment of the new Tulsa Section and the new Rome Subsection of the Syracuse Section.

With these additions, the Institute now has 59 Sections and 14 Subsections.

NBS Announces Air-Launched Automatic Weather Station

A self-contained automatic weather station, which transmits weather data by radio, has been developed by Percival D. Lowell and William Hakkarinen of the National Bureau of Standards for the Navy Bureau of Ships. Named the "Grasshopper," the device can be parachuted from aircraft onto inaccessible territory. Developed during World War II, it will automatically set itself up, and periodically make and transmit weather observations. It may also be used as a radio marker beacon.

After the station has parachuted to earth, controlled explosive charges are used to disengage the parachute, raise the station to an upright operating position, and erect a telescoping antenna. Weather-responsive devices then cause resistance changes which switch a radio transmitter on and off at a rate susceptible of translation by a receiving station into temperature, pressure, and humidity readings.

The developmental model of the weather station had an output on the order of 5 watts. Operating on a frequency in the neighborhood of 5 megacycles, it performed reliably over land at ranges of more than 100 miles. The dry batteries used provided power for transmission of weather reports at 3-hour intervals for more than 15 days.

Supreme Court Rules on Standards for Patents

The Supreme Court has raised its standards for the granting of patent rights permitting the holder of a patent to collect royalties and other fees.

The court gave an opinion declaring: "The function of a patent is to add to the sum of useful knowledge."

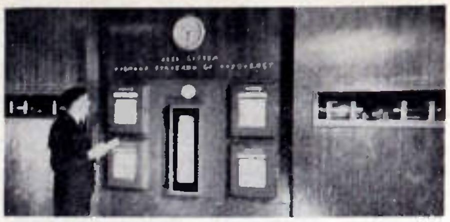
In two opinions setting forth the court's unanimous decision, the justices refused patent infringement claims on a grocery check-out device made by the Supermarket Equavent Corp. The company had sued the Atlantic and Pacific Tea Company.

"The Constitution," according to Associate Justice William O. Douglas, "never sanctioned the patenting of gadgets. Patents serve a higher end—the advancement of science. An invention need not be as startling as an atomic bomb to be patentable. But it has to be of such quality and distinction that masters of the scientific field in which it falls will recognize it as an advance."

Justice Douglas explained that for years the Supreme Court has steered clear of upsetting findings dealing with inventions when those findings have been approved by two lower courts. But the Supreme Court now holds that that rule "never had a place in patent law."

In the future, Justice Douglas said, "the controlling point will be the 'standard of inventions'—a patent must serve the ends of science—push back the frontiers of chemistry, physics, and the like and make a distinctive contribution to scientific knowledge."

Bell System Frequency Standard



Four extremely stable 100-kc crystal oscillators (at extreme left and right) with a precision of one part in a billion control a master timekeeper recently installed at the Bell Telephone Laboratories at Murray Hill. N. J. Serving as the primary standard of frequency and time for the Bell System, the apparatus is used to monitor or regulate equipment for radio and television networks and for radio-telephone services.

Gaseous Electronics Conference Slated

The fourth Conference on Gaseous Electronics, sponsored by the Division of Electron Physics of the American Physical Society, will meet October 4–6 in Schenectady, N. Y., it has been announced. Sessions will be held at the General Electric Research Laboratory, in the new buildings at the Knolls which were formally dedicated last autumn.

The program, now being arranged, will include papers pertaining to the fundamental physics of gas discharges. Full information may be had from secretary of the conference, Dr. James D. Cobine, GE Research Laboratory, P.O. Box 1088, Schenectady, N. Y.

New ECPD Manual Available

A guidance manual intended for engineers who are aiding young men interested in the engineering profession was recently published by the Engineers' Council for Professional Development. The 15-page pamphlet, prepared by the ECPD Guidance Committee, urges members of local engineering societies and section and chapters of national engineering societies to establish guidance committees to aid high-school pupils to determine whether they are qualified for careers in engineering. The present critical engineering manpower shortage emphasizes the need for guidance of the type indicated in this manual.

The pamphlet explains briefly how to organize advisory committees and how to select committee members. It contains suggestions for working with high-school and secondary-school students, and lists aids especially useful in the counseling of high-school boys.

The guidance manual is supplemented by an appendix, "Shall I Study Engineering?," a questionnaire to be filled out by the student for the use of the counseling engineer.

Copies of the manual with the questionnaire may be obtained from ECPD, 29 W. 39 St., New York 18, N. Y., for 20 cents. The cost of the manual when purchased separately is 15 cents, and the price of the questionnaire 10 cents. A deduction of 25 per cent for purchases of twenty-five or more is allowed.

Two Firms Enter Agreement

The International Telephone and Telegraph Corporation and the Western Electric Company, the latter acting for itself and for the American Telephone and Telegraph Company, have concluded a world-wide nonexclusive crosslicensing patent agreement, the companies announced recently.

The agreement is effective for a minimum of six years to December 31, 1956. It can be terminated on that date by one year's prior notice, but otherwise will continue indefinitely, subject always to one year's notice.

Under the agreement the two companies license each other to use inventions made before and during the life of the agreement as long as the patents are effective.

Educational Institutions are Elected to Research Council

Election of four educational institutions to active membership in the Engineering College Research Council of the American Society for Engineering Education has been announced by Gerald A. Rosselot, director, Georgia Technical Engineering Experiment Station and chairman of the Council.

The four are: the California Institute of Technology, Dartmouth College (Thayer School of Engineering), Montana State College, and the University of Toledo.

Extensive activity in engineering research supplementing an effective program of undergraduate engineering education is required for election to the Research Council. These elections bring to 88 the number of institutions active in the organization, which represents the leading educational and research centers in the United States.

FIRST CALL!

AUTHORS FOR IRE NATIONAL CONVENTION!

W. H. Doherty, Chairman of the Technical Program Committee for the 1952 IRE National Convention, to be held March 3–6, requests that prospective authors submit the following information:

1. Name and address of author
2. Title of paper
3. A 100-word abstract and additional information up to 500 words (both in triplicate) to permit an accurate evaluation of the paper for inclusion in the Technical Program.

Please address all material to W. H. Doherty, Bell Telephone Laboratories, Inc., Murray Hill, N. J. The deadline for acceptance is November 5, 1951. Your prompt submissions will be appreciated.

Industrial Engineering Notes[1]

RTMA Elections

Robert C. Sprague, president of the Sprague Electric Co., was re-elected **Chairman of the RTMA Board of Directors,** as the Association concluded its 27th annual convention on June 7, at the Stevens Hotel, Chicago. . . . The Board of Directors, at its organization meeting also re-elected **Leslie F. Muter as RTMA Treasurer** for his 16th term and renamed **W. R. G. Baker as Director of the Engineering Department, James D. Secrest as Secretary and General Manager,** and **John W. Van Allen as General Counsel.** . . . In addition, two new Directors and 12 former Directors were elected during the convention. The two new Directors are Robert S. Alexander, President of Wells-Gardner and Co., who succeeded George M. Gardner on the latter's resignation from the Board of Directors, and Harlan B. Foulke, vice-president and director of sales of Arvin Industries following his resignation as a Director. . . . Directors who were re-elected for three-year terms are: Benjamin Abrams, Emerson Radio and Phonograph Corp.; Max F. Balcom, Sylvania Electric Products Inc.; W. J. Barkley, Collins Radio Co.; H. C. Bonfig, Zenith Radio Corp.; Herbert W. Clough, Belden Manufacturing Co.; John W. Craig, Crosley Division, Avco Mfg. Corp.; E. G. Fossum, Stewart-Warner Electric Division; G. Richard Fryling, Erie Resistor Corp.; J. J. Kahn, Standard Transformer Corp.; F. R. Lack, Western Electric Co., Inc.; W. A. MacDonald, Hazeltine Electronics Corp.; and A. D. Plamondon, Jr., The Indiana Steel Products Co.

Television News

The **Supreme Court** upheld the Federal Communications Commissions' decision **authorizing color television standards on the field sequential method of the Columbia Broadcasting System.** The FCC decision had been appealed by the Radio Corporation of America and the Emerson Radio and Phonograph Corporation. . . . Following the decision, RCA announced it would continue its experimental color TV broadcasts and CBS said it expected to add a substantial schedule of color television programs within a few months. . . . Senator Edwin C. Johnson (D. Colo.), Chairman of the Senate Interstate and Foreign Commerce Committee, last week urged the **FCC to consider the legal points raised by the Communications Bar Association with respect to the proposed TV allocation plan.** The Bar contends that the Communications Act does not authorize the Commission to adopt a system of block allocations on a geographical basis. . . . A report by the **Ad Hoc Committee of the National Television System Committee,** outlining the "broad framework of a new composite system of color television," achieved by combining the best elements of the furthest advances of existing systems, was released to the television industry recently. This Committee was established by W. R. G. Baker on November 20, 1950, to make an "up-to-date appraisal of the state of the art" of color television. The committee comprised David B. Smith of Philco Corporation; Robert M. Bowie of Sylvania Electric Products Inc.; Elmer W. Engstrom of Radio Corporation of America; Thomas T. Goldsmith, Jr., of Allen B. DuMont Laboratories; Ira J. Kaar, of the General Electric Company; and Arthur V. Loughren of Hazeltine Electronics Corporation. The committee held meetings between November, 1950, and February of this year to witness various demonstrations and to hear discussions of developments and improvements in color television. The report distributed to the industry said that the special subcommittee recommended the use of the present transmission standards for black-and-white television for the transmission of compatible color television. In the report, the members of the special committee stated that the report represents the technical views of the committee serving in an "individual engineering capacity," and did not commit the companies by which they are employed. W. R. G. Baker, chairman of the National Television System Committee, an independent technical group sponsored but not directed by the Board of Directors of the Radio Television Manufacturers Association, said in a letter accompanying the report that it "demonstrates the considerable progress that has been made toward achieving a compatible television system which gives a high-quality picture and is economically practicable." Dr. Baker recommended in his letter that all television companies be urged to take part in the development of a new color-television system and that this work be co-ordinated by panels of technical experts working under the National Television System Committee.

FCC News

FCC Commissioner Frieda Hennock was appointed to a judgeship in the U. S. District Court of New York by President Truman. The appointment is subject to confirmation by the Senate. Miss Hennock has been with the FCC since July, 1948 . . . The President nominated FCC Chairman **Wayne Coy** for a new seven-year term on the Commission . . . Despite the recent NPA construction order requiring special authorization for radio and TV broadcasting station construction, the FCC indicated this week that it will continue its policy of granting permits and letting applicants for new stations take their chances of obtaining authorization from the NPA.

Controls

The Electronic Parts and Components Distributors Industry Advisory Committee asked the National Production Authority to **extend** DO-97 (Maintenance, Repair and Operating) ratings to licensed amateur radio operators. The Committee cited that "ham" operators are vital to civilian defense and constitute an important auxiliary system of communication during other emergencies. . . . **E. T. Morris, Jr.**, who is on leave from the Westinghouse Electric Corp., on Tuesday, May 22, **took over the directorship of the Electronics Division of the National Production Authority.** He will continue to serve as Chairman of the Electronics Production Board of the Defense Production Administration.

DPA Publishes Guidebook for Small Manufacturers

The Defense Production Administration this week published a 31-page guidebook designed to help small manufacturers during the defense mobilization period. . . . J. C. Pritchard, Deputy Administrator for Small Business, described the guidebook, "Mobilization Guide for Small Business," as a quick source of help to businessmen in locating services the Government is providing so that they can share, so far as requirements exist and their capabilities permit, in defense orders and in obtaining material available to supply consumer requirements. . . . The guidebook is obtainable at all field offices of the Department of Commerce.

Industry Statistics

According to preliminary reports by the Bureau of the Census, almost every home had a radio in April, 1950, and about 5.1 million dwelling units (one out of eight) had a television set. . . . Reflecting the lull in television set sales, TV picture tube sales to manufacturers dropped 54 per cent in April below the March figure, according to RTMA's monthly cathode-ray tube report. . . . RTMA's TV tube sales report showed that 89 per cent of all tubes sold to set manufacturers were rectangular in form and 95 per cent were 16 inches and larger in size. . . . Production of television receivers in April was cut back 36 per cent below the monthly rate established in the first quarter of this year, according to RTMA's estimate of the industry's output. Radio set output, however, dropped only five per cent below the first quarter rate. . . . RTMA's estimates, which include production by members of the Association and nonmembers, showed a total of 1,337,042 radios and 469,157 TV sets manufactured in April. This compares with the quarter's monthly average of 1,411,866 radios and 733,223 television receivers.

[1] The data on which these NOTES are based were selected by permission from *Industry Reports*, issues of May 18, May 25, June 1, and June 11, published by the Radio-Television Manufacturers Association, whose helpfulness is gratefully acknowledged.

IRE People

Harry F. Olson (A'37–VA'39–SM'48–F'49), director of the acoustical laboratory of the RCA Laboratories, Princeton, N. J., has been elected president of the Acoustical Society of America for the year 1952, it was announced at the society's spring meeting held recently in Washington, D. C.

Harry F. Olson

Born in Mt. Pleasant, Iowa, Dr. Olson received the B.E. degree in 1924, the M.S. degree in 1925, the Ph.D. in 1928, and the professional E. E. degree in 1932, all from the University of Iowa. He has been engaged in acoustical research with the Radio Corporation of America since 1928, and became director of its acoustical laboratory in 1945. A pioneer in the research and development of various types of directional microphones now almost universally employed in broadcasting, television, and sound motion pictures, he is responsible for various important developments involving loudspeakers, phonograph pickups, sound absorbers, room acoustics, underwater sound, and listener preference tests. In addition, he lectured on acoustical engineering at Columbia University, from 1939 to 1942, and is the author of three books on the subjects of acoustics and vibrating systems, 50 articles in scientific journals, and holds 62 patents.

Dr. Olson is a member of Tau Beta Pi, Sigma Xi, and the American Physical Society, and a Fellow of the Acoustical Society of America. In 1949 he was awarded the John H. Potts Medal of the Audio Engineering Society.

His Institute activities have included work on the following Committees: Board of Editors, Electroacoustics, Papers Procurement and Standards; he served also as Institute Representative on several ASA Sectional Committees.

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Raymond W. Andrews (A'35–M'46), formerly merchandising manager of the radio tube and television picture tube divisions of Sylvania Electric Products Inc., has been promoted to the position of manager of factory sales, according to a recent announcement. In his new capacity he directs tube merchandising, sales planning, and customer service activities, and is responsible for maintaining sales department liaison with tube plants.

R. W. Andrews

A Commander in the Naval Reserve, Mr. Andrews had control of allocation and distribution of the Navy's fire control radar equipment during World War II. Later he was in charge of a Naval activity which produced major assemblies of the famous "Bat" guided missile. He joined the Sylvania Electric Products staff in 1945.

A native of Buffalo, N. Y., he attended Lehigh University and the University of Buffalo. He is a member of the American Radio Relay League, the Society of American Military Engineers, the Radio Club of America, and the Reserve Officers of the Naval Services.

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Keron C. Morrical (SM'44), Secretary-Treasurer of the St. Louis Section of the IRE, died recently of a heart attack, it has been learned.

Born in 1908 in Illinois, Mr. Morrical was graduated from the University of Illinois in 1933, and received the Ph.D. degree there in 1936. He began his career as acoustical development engineer at the RCA Victor Division in Camden, N. J., and Indianapolis, Ind., where he was concerned mainly with electroacoustical measurements with regard to theaters, broadcast and recording studios. Later he was employed in this capacity with the United States Gypsum Company in Chicago, Ill. During World War II, Mr. Morrical served at the Underwater Sound Laboratory at Harvard University, whereupon he returned to RCA in Indianapolis. His last position was that of professor at Washington University in St. Louis, Mo.

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H. O. Peterson (A'22–M'31–F'41), research section head in charge of the Radio Reception Laboratory, RCA Laboratories Division, at Riverhead, L. I., N. Y., was awarded the honorary degree of Doctor of Engineering by the University of Nebraska on June 4, 1951.

Dr. Peterson received the B.S. degree in electrical engineering from the University of Nebraska in June, 1921. After a short period with General Electric, he joined RCA in 1922. Since that time he has been actively engaged in the field of research and advanced development as applied to communication receivers, receiving antennas, and propagation. He has been in charge of the Radio Reception Laboratory at Riverhead since 1929.

Dr. Peterson has served on the following IRE Committees: Board of Editors, Industrial Electronics, Radio Receivers, Standardization, Transmitters and Antennas, and Wave Propagation. His contributions to the field of communications receivers, receiving antennas, measuring techniques, and propagation are extensive.

Over 130 United States patents have been issued up to the present time in his name.

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Robert D. Huntoon (A'40–SM'47) has recently been appointed Associate Director of the National Bureau of Standards, in charge of the Bureau's newly established Corona Laboratories, near Corona, Calif. Formerly Chief of the NBS Atomic and Radiation Physics Division, he will assume his new duties immediately and will be concerned with various phases of electronic research, development, and engineering. Dr. Huntoon's special fields of interest include atomic beam measurement, special amplifiers, atomic particle counting, electronic ordnance devices, the phasing of oscillators, and the study of the deuteron-deuteron nuclear reaction.

R. D. Huntoon

Dr. Hunton joined the staff of the National Bureau of Standards in 1941, and was one of the principal NBS scientists to design and develop the proximity fuze. In 1944 his services were loaned to the War Department, where he served as consultant on proximity fuzes and related problems in the Office of the Secretary of War. He was appointed Chief of the NBS Electronics Section in 1945, and directed fundamental research on electronic circuits, control devices, and other electronic ordnance components. In 1947 he became Assistant Chief of the Atomic and Radiation Physics Division, and in 1948, Chief. He has served also as Coordinator of Atomic Energy Commission Projects, National Bureau of Standards.

Born in 1909 at Waterloo, Iowa, Dr. Huntoon received the B.A. degree from Iowa State Teachers College in 1932. In 1944 he was awarded a graduate assistantship in the physics department of the State University of Iowa, where he received the Ph.D. degree in 1938. From 1938 to 1940 he was a member of the faculty of New York University, and during the period 1940–1941, a research physicist in the vacuum-tube division of Sylvania Electric Products Inc.

Dr. Huntoon is active in the work of the Research and Development Board, and holds membership in Sigma Xi, the American Physical Society, and the Philosophical Society of Washington. Recently he received the Washington Academy of Sciences Award in Physical Sciences for his contributions to the field of electronics.

Jay E. Browder (A'38–M'44) was appointed chief of the radio communications engineering section of Kollsman Instrument Corporation, now a subsidiary of the Standard Coil Products Company, Inc., it was announced recently.

Jay E. Browder

Mr. Browder has done engineering research on automatic direction finders, airborne radar equipment, microwave communication systems, and microwave instrument landing systems. He has various patents to his credit, and is the author of several technical papers in the aircraft radio communication and navigation field. A member of the Institute of Aeronautical Sciences, he also participated in the work of the Radio Technical Commission for Aeronautics, and was on its famous "Special Committee 31" for air traffic control. In addition, he has been active in the radio aids group and the flight technical group of the International Air Transport Association.

Prior to joining the Kollsman Instrument Corporation, Mr. Browder spent 11 years on the engineering staff of the Sperry Gyroscope Corporation where he last held the position of engineering section head for aeronautical radio equipment. Previously he was associated with TWA in its radio engineering laboratory at Kansas City, Mo.

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Warren B. Burgess (A'20–M'31–SM'43), radio engineer at the Naval Research Laboratory, was designated, upon the nomination of the Department of the Navy, as a member of the United States Delegation to the sixth assembly of the International Radio Consultative Committee (CCIR), which was held in Geneva, Switzerland, from June 5 through July 6, it was recently announced. The CCIR, a permanent organ of the International Telecommunications Union, studies technical questions in the field of radio. At the Geneva Meeting, the results of studies in the various participating countries on technical radio matters were considered, and recommendations drawn up regarding them.

Although this was the sixth assembly of the CCIR, it was the first meeting since its establishment at Stockholm in 1948 as a continuing international body, with an elected director, Dr. Balthus van der Pol.

Mr. Burgess received the B.S. and E.E. degrees from the Worcester Polytechnic Institute, Worcester, Mass., in 1916 and 1917, respectively. He has been actively associated with amateur radio since 1908, having held the First Grade Commercial Operator's license since 1910. During World War I he served as a radioman and ensign, engaged in the inspection and testing of shipboard radio equipment, and also the installation and development of radio direction finders. After the war, he became a radio inspector and aid at the Navy Yard, in Boston, Mass., with similar duties.

In 1921 he joined the staff of the Naval Research Laboratory, where he was directly in charge of research and design for improvement of naval radio direction finders. He has made a number of inventions in this field, for two of which patents have been issued. Mr. Burgess has served on the Papers Procurement and Navigation Aids IRE Committees, and was Secretary of the Washington, D. C., IRE Section during 1935–1936, and Chairman in 1937.

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H. T. Kohlhaas (SM'46), aged 68, former assistant vice-president of the International Telephone and Telegraph Corporation, died recently of injuries received in an automobile accident near Jacksonville, Fla.

Mr. Kohlhaas had been associated with International Telephone and its predecessor companies for 42 years, and was editor of *Electrical Communication*, the company's technical magazine, which, under his guidance, became a leading contemporary journal of communication development. Prior to his affiliation with IT&T he had belonged to the staff of the Western Electric Company Laboratories, now Bell Telephone Laboratories, Inc.

Born in Brooklyn, N. Y., Mr. Kohlhaas was graduated from Cooper Union in 1907. Later he received engineering degrees from Cooper Union and the Polytechnic Institute of Brooklyn.

He was a Fellow of the American Institute of Electrical Engineers, and a past president of the Brooklyn Polytechnic Alumni Association.

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I. S. Coggeshall (A'26–M'29–F'42) President of The Institute of Radio Engineers, and general traffic manager of Western Union Telegraph Company's overseas communications, was awarded the honorary degree of Doctor of Engineering by Worcester Polytechnic Institute at commencement exercises held June 17, 1951.

I. S. Coggeshall

Dr. Coggeshall received recognition for his work in the adoption of electronic methods and devices to the telegraph and submarine cable field. The most recent of these devices is a traffic capacity-doubling amplifier now being inserted into transatlantic cables and sunk on the ocean bottom.

A holder of one of the early amateur radio licenses in 1911, Dr. Coggeshall studied electrical engineering at Worcester Polytechnic Institute before joining Western Union in 1917. For the past 14 years he has specialized in ocean cables, and during World War II he represented his company on the cable committee of the Board of War Communications.

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Rodney D. Chipp (A'34–SM'43), director of engineering at the DuMont Television Network, has been elected a director of the Technical Societies Council of New York, an affiliation of the local sections of 17 technical societies dedicated to service to the member societies and to technical men and women of the New York area. Mr. Chipp represents the New York Section of the IRE on this council.

Born in New Rochelle, N. Y., in 1910, Mr. Chipp attended the Massachusetts Institute of Technology. Prior to his affiliation with DuMont, he held the post of radio facilities engineer at the National Broadcasting Company. In 1945, after five years of active duty in the Navy, he was awarded the Commendation Ribbon with a citation for his work in radar development.

Mr. Chipp is a member of the Veteran Wireless Operators Association and also is an associate member of the United States Naval Institute.

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Beardsley Graham (S'39–A'41–SM'47), formerly of the Bendix Aviation Corporation, has recently been appointed a member of the engineering staff of the Stanford Research Institute, as an assistant chairman of the engineering department. Mr. Graham, a specialist in radar systems and television development, will take charge of the Institute's equipment engineering and advanced techniques program.

Beardsley Graham

His previous professional experience includes service as an engineer with the RCA Manufacturing Company, television development engineer with the National Broadcasting Company, and staff member of the MIT Radiation Laboratory, where he worked on shipborne radar systems during the recent World War II.

A graduate of the University of California, Mr. Graham has completed two and one-half years of graduate work in electrical engineering physics, and in ultra-high-frequency electronics at the University of California and Columbia University. He is a registered professional engineer in the State of California, a member of the American Physical Society, the American Rocket Society, and the Eastern Association for Large Scale Computing Machinery. He is active as well on the IRE Professional Group on Nuclear Science.

Books

Movies for TV by John H. Battison

Published (1950) by the Macmillan Co., 60 Fifth Ave., New York 11, N. Y. 369 pages+6-page appendix+xv pages. 66 figures. 5½×8½. $4.25.

This is a discussion of film production use and techniques as these are related to television, which, because of the amount of material to be covered, cannot investigate any one particular problem in great technical detail. As the author himself points out, "This book is not intended to produce engineers, producers, or even technicians, but after reading and studying it the reader should be well prepared for any job in the film department of a television station that does not require specialized technical knowledge."

There is of course a vast need in the growing television industry for a broader understanding of the tools of the trade. This book quite excellently fills such a need for general education in terminology and the description of specific equipment and industry practices relating to the production and use of film in television stations. The chapters on "Fundamentals" in the first portion of the book are written particularly for the nontechnical reader. In addition to discussions of methods, equipment, and terminology, this section includes chapters on editing, special effects, and suggestions for shooting films for television. The second portion of the book deals with "The Program Angle" and includes not only the choice of, and various uses of film for television programming, but also discusses problems of planning, directing, and shooting film for such programming. Some of these chapters repeat material previously discussed, but this is justified if the book is used as a reference text where certain chapters may be considered separately.

The liberal use of examples and expressions of opinions by the author has been effective in presenting material that is easily readable as well as instructive.

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Semi-Conductors by D. A. Wright

Published (1950) by John Wiley & Sons, Inc., 440 Fourth Ave., New York 16, N. Y., 128 pages+2-page index. 32 figures. 4¼×6¾. $1.75.

This pocket-sized book is one of a series of Methuen's Monographs on Physical Subjects. It deals mainly with the basic theory of electron flow and emission phenomena in semiconductors. The material is presented on a fairly elementary level and is largely descriptive in nature. Some of the more theoretical discussions in the text, however, will be found difficult to follow by a communication engineer unfamiliar with modern physics.

The reader who expects to find in this book a detailed discussion of transistors will probably be disappointed. Only one chapter is given over to the subject of electron flow across the boundary between a semiconductor and a metal, and of this chapter only four pages are devoted to the transistor.

The scope of the book is indicated by the chapter headings which are as follows: "Electrons and Metals"; "Electrons in Crystals"; "Electron Emission from Semi-Conductors"; "Determination of Electron Density in Semi-Conductors"; "Secondary Emission"; "Metal—Semi-Conductor Contacts"; "Thermionic Cathodes"; and "Photo-Electric Cathodes."

Despite its small size, the book contains a wealth of factual information which should be of value both to the research worker in the field of semiconductors, and to the student of the electron physics of solids.

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Antennas by John D. Kraus

Published (1950) by McGraw-Hill Book Co., 330 W. 42 St., New York 18, N. Y. 480 pages+50-page appendix+4-page references+xii pages. 371 figures. 6×9. $8.00.

The purpose of this new book is the presentation of the basic theory of antennas with emphasis on their engineering applications. The book is primarily intended as a text at the senior or first-year graduate level, and appears to be well suited for this purpose. It will also prove useful as a reference and refresher for the practicing engineer. An introductory course in electromagnetic theory is desirable as a prerequisite, although not necessary if the student is willing to accept a few basic points on faith.

The book begins with a detailed treatment of the radiation patterns and directivities of arrays of point sources and of apertures. Then the self-impedance, pattern, directivity, and other properties are studied for linear, loop, helical, and biconical antennas. The various methods of analysis, such as the emf method, the integral-equation method, and the biconical transmission-line method, are all utilized, although for the last two cases the treatments are not carried through completely. In a graduate course, the instructor may wish to provide supplementary reading in the literature. Following the treatment of these basic types of antennas is a chapter on the mutual impedance of linear antennas, and the input impedance and other properties of linear-antenna arrays. Additional chapters deal with reflector, slot, horn, lens, long-wire, and other types of antennas. The final chapter covers many kinds of antenna measurements in an introductory manner. The book is well documented with footnote references as an aid to the student or engineer desiring additional information on any antenna subject. A lengthy appendix is packed with useful tables, graphs, and other material. Problems are included at the end of each chapter.

As must be expected in the first printing of a new book, there are a number of typographical errors. Most of these, however, will be obvious to the reader. There are also places where nonrigorous derivations have been used that can be improved upon with little or no additional difficulty for the student, and it is suggested that this be done in a subsequent edition. On the whole, Professor Kraus's book has fulfilled its objective very well, and it is a valuable contribution to the technical literature on antennas.

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Transmission Lines and Filter Networks by John J. Karakash

Published (1950) by the Macmillan Co., 60 Fifth Ave., New York 11, N. Y. 365 pages+5-page index +39-page appendix+xii pages. 388 figures. 6×9¼. $6.00.

Misled by the title, the reader might expect this book to treat its subject matter in two sections. Actually it consists of three, the first of which is a 145-page discussion of transmission lines, the second an 80-page treatment of network analysis, and the third a 140-page analysis of elementary filter theory. In addition, there are two appendixes devoted to Maxwell's equation and its application to various propagation devices such as waveguides, and to the principles of matrix algebra.

The book is intended for undergraduate students, and in the author's own words, its "subject matter is generally treated on the quantitative basis with a minimum of descriptive matter."

The first section is a nicely condensed version of the mathematics of transmission lines. It covers the derivations of the constants of the line, propagation along the line, impedances, distortionless lines, composite lines, telephone lines, cables, uhf lines, and many associated items. The first two pages of the text start with a discussion of vector propagation, and the first equation given is a vector equation for Poyntings' vector. After that, vectors are not used until the appendixes. It would have been less confusing if these two pages had been omitted.

In the very brief second and third sections the discussion does not appear to familiarize the reader adequately with the material. Network constants, the four terminal networks, T, Pi and lattice sections, transmission parameters, and artificial lines are included in section two, and the third section deals with fundamental filter equations, the constant-K filters, elementary synthesis, M-derived filters, coupled-tuned circuits, lattice filters, and various types of transmission-line filters. Even with this very large scope to cover, the author devotes too much space to reference charts redrawn from various reference works. At one point he includes a full eight pages of charts redrawn from a handbook.

As a set of condensed notes which are to be enlarged upon by the instructor, the book could probably be used in some of our universities where a great many subjects are being crammed into the undergraduate students in a very limited time.

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Jew Publications

High Speed Photography, Volume 3, he third in a series of technical reprints, has ust been announced by the Society of Mo-.ion Picture and Television Engineers. This 60-page paper-bound volume contains 11 .rticles, a 19-page bibliography listing nearly 00 references on cameras, lighting, oscillo-graphy, Schlieren optics, application of sci-ntific photography including X-ray, plus a umulative index to the 44 articles appearing n all three of the Society's volumes that ave been published so far.

This is a composite reprint of articles of special interest to the general field of pho-ography in engineering and research which ave appeared in the Journal of the SMPTE rom January, 1950, through January, 1951. The price is $2.00 (postage prepaid) per single copy, and the publication may be ob-tained from Society headquarters, 40 West 40 Street, New York 18, N. Y.

Quartz Vibrators and Their Applications by P. Vigoureux and C. F. Booth

Published (1950) by His Majesty's Stationery Office for the Department of Scientific and Industrial Research, Charles House, 5-11 Regent Street, London S. W. 1, England. 329 pages+6-page index+27-page appendix+9-page bibliography. 186 figures. 64 plates. 6×9½. $6.75.

While giving credit for advances made in the United States and in Germany, the present volume is largely an account of British practice during and since World War II. As such it will be welcomed in this country, especially by those concerned with the uses of quartz crystals in communication engineering.

As the title indicates, the treatment is confined to quartz, a fact which restricts the value of the chapters on filters, ultrasonics, and other applications of transducers. On the other hand the physical properties of quartz, including the piezoelectric and elastic relations, are well presented in Chapters 2, 3, and 4.

Chapters 5 and 6 on the theory of the resonator and of the crystal-controlled oscillator are clear and well written. The treatment of the determination of the equivalent network, largely by use of graphical methods, is valuable because it presents a summary of some of the chief features in Dye's classical 1926 paper

More than half of the book is devoted to the selection, orientation, and processing of quartz plates, aging phenomena, mounting, temperature control, frequency standards, transmitting circuits, filter theory and filters, and related matters. These subjects are well presented, and illustrated with many drawings and beautiful photographs.

The statement on page 11 that quartz shows only conchoidal fracture should be considered as applying only to the *natural* crystal. For on the same page and on page 16, the fracture of plates in planes parallel to the major or minor rhombohedral faces is mentioned. This too is illustrated in Plate 4.

On page 56 it should have been stated that C_f is the capacitance at constant stress, for only thus can equation (45) on page 63 be understood.

The only misprint noted by this reviewer is on the last line of page 70, where the last expression should have a minus sign.

In the main, the book uses the notation recommended in the IRE 1945 report on piezoelectric standards. It is unfortunate that the authors could not use the 1949 report, especially with regard to the conventions for right and left quartz and for rotated cuts.

The presswork is excellent. There is a good index, and a bibliography with 204 references to books, periodicals, and patents.

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Understanding Radio, Second Edition, by H. N. Watson, H. E. Welch, and G. S. Eby

Published (1951) by McGraw-Hill Book Co., 330 W. 42 St., New York 18, N. Y. 646 pages+14-page index+55-page appendix+lx pages. 522 figures. 6×9. $5.50.

Written for the beginning student of radio, who is assumed to be without either mathematical or scientific training, this book presents a problem to the reviewer hardly less complex than that of the authors who undertook such a task. How would *you*, for example, undertake to explain electricity, magnetism, the detailed operation of vacuum tubes, waveforms, detectors, amplifiers, superheterodyne receivers, and even FM discriminators, to someone like, for instance, your mother-in-law?

A careful reading and rereading of the book make it appear that such an objective cannot be achieved; yet the progression from the explanation of fundamentals and operation of simple circuits to the more complex ones leads logically enough to an understanding of the general operating principles of the more complex circuits. In this respect, the authors are to be congratulated for having accomplished a task most scientific people would immediately consider hopelessly impossible.

Any such work must, of necessity, use explanatory terms which are oversimplifications, and which compromise the scientific truth severely. In most cases, however, the trained reader's objections to this should be overruled, for, after all, the untrained novice is entitled to a brief nontechnical explanation of scientific matters in a language he can comprehend. It is in this spirit, and with leniency, that the reviewer offers the observations below:

Intended as a textbook for beginning students who are, it is judged, without a high-school education, this book covers all of the usual radio circuits, vacuum-tube operation, amplifiers, speakers, power supplies, antennas, and touches upon vhf and frequency modulation. Obviously there are omissions of topics considered too complex which must be introduced by the teacher as they arise. Each chapter begins by outlining the things to be learned about its subject, i.e., how to build a typical circuit, and why it works; then it supplies a series of questions, and concludes with a list of definitions or explanations of new terms used. Television is confined to the concluding chapter entitled, "Looking Ahead in Radio," where brief mention is made also of radar.

Realizing fully the tremendous problem of describing the complex subject matter in simple language, the reviewer is forced, nevertheless, to point out several instances of overindulgence in this respect. For example, in the explanation of self-inductance and reactance of a choke coil, the effect is described in terms of "back voltage." The description concludes with the statement that "the technical name for this back voltage is inductive reactance." In describing the action of a two-section filter consisting of chokes K and L and condensers X, Y, and Z, connected in the conventional manner, in smoothing the pulsating current, the following statement occurs: "When a surge of electrons from the rectifier tube and the high voltage secondary flows as shown by the arrow at A, it will meet the opposition of the choke K. Since the back voltage set up by the choke coil K opposes the surge, the electrons will flow into side 1 of the condenser X. As the surge overcomes the opposition of the choke coil, the current flow starts through the coil. When this surge dies down, the flywheel action of the choke coil tries to keep the current flowing but there is no current remaining in this surge of electrons.

"Now the electrons in side 1 of the condenser flow out and through the choke coil. The result of the action of the condenser and the coil on the first surge of electrons is to reduce the strength of the current somewhat but to deliver a more even flow. This action repeats in condenser Y and choke coil L, which even the flow further and reduce the pulsation. Any remaining pulsations are killed by the action of the condenser Z, leaving a steady flow of direct current."

In comparing a push-pull amplifier with a single-tube amplifier, that the former has two tubes is ignored, and the increased output is described as due entirely to the fact that "the single-tube tank circuit gets a push from the plate once every two surges, while the push-pull tank circuit receives a push from the plate at each end of the circuit during each surge."

In the concluding chapter, "Looking Ahead in Radio," the authors make some very questionable statements. For example, the coaxial cable, for distribution of television programs is said to be " . . . not only . . . too expensive but (has) introduced so much distortion into the program as to be impracticable for television use over long distances." They suggest that small communities investigate an automatic repeater station to pick up programs from the nearest network repeater, whereby "Commercial announcements and other soundcasts can be inserted into the television-program breaks by a local station operator." And finally, a brief paragraph on "Atomic Energy and Nuclear Physics" contains the following which completely stops this reviewer: "It was difficult to study the atomic structure of molecules by means of infrared light because the particles moved too fast for easy study. When microwave frequencies were used, the motion of the particles was slow enough for easier study."

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Abstracts and References

Compiled by the Radio Research Organization of the Department of Scientific and Industrial Research, London, England, and Published by Arrangement with That Department and the *Wireless Engineer*, London, England

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 Annual Index to these Abstracts and References, covering those published in the PROC. I.R.E. from February, 1950, through January, 1951, may be obtained for 2s.8d. postage included from the *Wireless Engineer*, Dorset House, Stamford St. London S.E., England. This index includes a list of the journals abstracted together with the addresses of their publishers.

The number in heavy type at the upper left of each Abstract is its Universal Decimal Classification number and is not to be confused with the Decimal Classification used by the United States National Bureau of Standards. 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.

ACOUSTICS AND AUDIO FREQUENCIES

016:534 1541

References to Contemporary Papers on Acoustics—(*Jour Acous. Soc. Amer.*, vol. 23, pp. 121–129; January, 1951.) Continuation of 1042 of June.

534.21 1542

On the Propagation of Sound over Great Distances—J. Veldkamp. (*Jour. Atmos. Terr. Phys.*, vol. 1, no. 3, pp. 147–151; 1951.) The surface energy density of the abnormal sound from big explosions is calculated for a particular vertical temperature distribution in the atmosphere. The calculations indicate that waves with frequencies as low as 0.1 cps can be reflected strongly from an ionosphere layer at a height of 170 km. Measurements by Cox et al. on abnormal sound from the Heligoland explosion are in agreement with this conclusion. Waves with frequencies >70 cps are almost totally absorbed in the upper stratosphere. It is suggested that accurate measurements of the travel time and intensity of abnormal sound could be used to obtain information on the temperature and pressure in the ionosphere.

534.22 1543

Asymptotic Approximation for the Normal Modes in Sound-Channel Wave Propagation—N. A. Haskell. (*Jour. Appl. Phys.*, vol. 22, pp. 157–168; February, 1951.) Asymptotic methods are used to find approximate solutions of the acoustic wave equation in a medium in which the velocity is a continuously variable function of one co-ordinate. When the velocity function has a minimum, undamped normal mode solutions exist and are closely analogous to the internally reflected waves, in the case of a medium made up of discrete layers. By converting the sum of the high order normal-modes into an equivalent integral, superposition of these modes leads to geometrical ray theory, modified by diffraction in a manner that may be computed from the incomplete Fresnel and Airy integrals.

534.321.9 1544

Ultrasonic Generators for High Powers—B. E. Noltingk. (*Jour. Brit. IRE*, vol. 11, pp. 11–19; January, 1951. Discussion pp. 20–21.) Some common piezoelectric and magnetostrictive types of ultrasonic generator are described; theory is discussed briefly. Applications for which generators have been specially designed include flaw detection, aluminium soldering, and therapy.

534.321.9:534.22–16 1545

Improved Methods for Measuring Ultrasonic Velocity—G. W. Willard. (*Jour. Acous. Soc. Amer.*, vol. 23, pp. 83–93; January, 1951.) Full paper. Summary abstracted in 516 of April.

534.231:534.321.9 1546

Piston Source at High Frequencies—A. O. Williams, Jr. (*Jour. Acous. Soc. Amer.*, vol. 23, pp. 1–6; January, 1951.) For a circular plane piston of radius a, producing an ultrasonic beam with propagation constant k ($=2\pi/\lambda$), an expression is derived for the velocity potential, averaged with respect to magnitude and phase over a "measurement circle" of area equal to that of the piston and centered in the beam. The expression should be highly accurate for $ka \geqq 100$, at distances z from the source such that $(z/a)^3 \geqq ka$; it agrees well with results computed in another way by Huntington, Emslie, and Hughes (2479 of 1948). The assumption that relatively near the source there is a collimated beam of plane waves is shown to be not very accurate; the averaged radiation pressure falls off monotonically over all distances considered. The velocity potential at the rim of the "measurement circle" is also computed and compared with the value deduced on the plane-wave assumption.

534.321.9:534.231–14 1547

Experimental Determination of Acoustic Wave Fronts—P. Tamarkin, G. L. Boyer, and R. T. Beyer. (*Jour. Acous. Soc. Amer.*, vol. 23, pp. 7–11; January, 1951.) Full paper. Summary abstracted in 517 of April.

534.321.9:534.373–14 1548

On the Dispersion and Absorption of Supersonic Waves in Water—B. B. Ghosh. (*Indian Jour. Phys.*, vol. 24, pp. 1–12; January, 1950.) Theoretical expressions for dispersion and absorption in water are derived and found in good agreement with experimental results. Calculation indicates that the dispersion region for liquids, such as water, is in the neighborhood of 1 to 10 kmc.

534.321.9:534.373–14 1549

Ultrasonic Absorption Measurements in Aqueous Solutions of Magnesium Sulfate—M. C. Smith, R. E. Barrett, and R. T. Beyer. (*Jour. Acous. Soc. Amer.*, vol. 23, pp. 71–74; January, 1951.) Full paper. Summary abstracted in 530 of April.

534.321.9:534.373–14 1550

Ultrasonic Absorption in Liquids—C. J. Moen. (*Jour. Acous. Soc. Amer.*, vol. 23, pp. 62–70; January, 1951.) Full paper. Summary abstracted in 531 of April.

534.321.9:537.228.1 1551

Crystal Systems with Low Loss—R. B. Fry and W. J. Fry. (*Jour. Appl. Phys.*, vol. 22 pp. 198–200; February, 1951.) Approximate formulas are derived for computing the effect of mechanical resistance on the electrical characteristics of a piezoelectric-crystal system, for the case of a general reactive termination and low loss. Values of quality factor and of $\Delta f/f_r$(Δf = difference between resonance frequency f_r and antiresonance frequency) obtained from the formulas are compared with the results of exact calculations for quartz Hg and ADP-Hg systems.

534.321.9:537.228.1 1552

Variable Resonant Frequency Crystal Systems—W. J. Fry, R. B. Fry, and W. Hall. (*Jour. Acous. Soc. Amer.*, vol. 23, pp. 94–110; January, 1951.) Full paper. Summary abstracted in 525 of April.

534.374:534.75 1553

Acoustic Filters as Ear Defenders—J Zwislocki. (*Jour. Acous. Soc. Amer.*, vol. 23, pp. 36–40; January, 1951.) Special ear plugs incorporating low-pass filters are described, which afford protection against excessive noise while admitting part of the spectrum of speech frequencies.

534.41:621.317.757 1554

Analysis of a Spectrum of Very Low Frequencies by means of Magnetic Tone-Frequency Equipment—Weber. (*See* 1720.)

534.612 1555

Absolute Measurement of Sound Pressures at High Frequency—V. Timbrell. (*Nature* (London), vol. 167, pp. 306–307; February 24, 1951.) Description of a method using a Mach-Zehnder type of optical interferometer with a light beam of limited width.

534.7:611.85 1556

The Coarse Pattern of the Electrical Resistance in the Cochlea of the Guinea Pig (Electroanatomy of the Cochlea)—G. v. Békésy. (*Jour. Acous. Soc. Amer.*, vol. 23, pp. 18–28; January, 1951.)

534.7:611.85 1557
Microphonics Produced by Touching the Cochlear Partition with a Vibrating Electrode—G. v. Békésy. (*Jour. Acous. Soc. Amer.*, vol. 23, pp. 29–35; January, 1951.)

621.395.623.7.094.3 1558
The Origin of Nonlinear Distortion in Electroacoustic Transducers—E. Hüttmann. (*Elektrotechnik Berlin*, vol. 4, pp. 322–325; September, 1950.) Discussion shows that in af amplifiers the distortion can be kept small without much difficulty, and that in order to radiate sound with minimum distortion, division of the frequency range among several loudspeakers is necessary. Even for modest requirements, separation of the high-tone and low-tone loudspeakers is essential.

621.395.625.2/.3+681.85 1559
Recorders and Reproducers—J. Moir. (*FM-TV*, vol. 11, pp. 14–15, 28; January, 1951.) An account of recording equipment and techniques used by the BBC and short descriptions of two British lightweight pickups, the EMI Type 12 and the Decca Type XMS.

621.395.625.3 1560
Magnetic Recording Systems in Product Design—A. E. Javitz. (*Elec. Mfg.*, vol. 45, pp. 74–81, 204; February, 1950.) A survey of design progress to date and a discussion of applications, including experimental use for recording signals other than sound, e.g. signals for controlling the operation of production machinery.

534 1561
Fundamentals of Acoustics [Book Review]—L. E. Kinsler and A. R. Frey. Publishers: John Wiley and Sons, New York, N. Y. 1950, 499 pp., $6.00. (*Jour. Soc. Mot. Pic. & Telev. Eng.*, vol. 56, p. 130; January, 1951.) "The authors have maintained a very good balance between the fundamental aspects of the physics of the problems and the engineering applications. . . . Should serve as a very useful text in senior college and graduate classes."

534 1562
Wave Motion and Sound [Book Review]—R. W. B. Stephens and A. E. Bate. Publishers: Arnold and Co., London, Eng., 448 pp., 45s. (*Brit. Jour. Appl. Phys.*, vol. 2, p. 27; January, 1951. "This book provides a comprehensive, reliable, and attractive degree course."

ANTENNAS AND TRANSMISSION LINES

621.392.21.011.2:621.3.013.78† 1563
Radiation from Resonant Quarter-Wave Transmission Lines—F. M. Leslie. (*Wireless Eng.*, vol. 28, pp. 70–72; March, 1951.) The input conductance at resonance of three balanced $\lambda/4$ transmission lines of different spacings was measured at 100 mc, and the effect of screening by metal sheets and troughs at various distances from the lines was observed. Results are shown graphically.

621.392.26†:621.3.09 1564
Propagation Characteristics in a Coaxial Structure with Two Dielectrics—A. Baños, Jr., D. S. Saxon, and H. Gruen. (*Jour. Appl. Phys.*, vol. 22, pp. 117–123; February, 1951.) For a given TM mode, the pertinent parameters are the ratio of the dielectric constants of the two media, the ratio of the two radii, and the operating frequency. As previously pointed out by Frankel (22 of 1948) and by Bruck and Wicher (334 and 1555 of 1948), it is found that the phase velocity for a given mode may lie between the phase velocities corresponding to each of the two unbounded dielectrics and that by proper choice of parameters, it is always possible to obtain a phase velocity of a preassigned value, higher than the lower of the velocities for the two unbounded media. This suggests the possibility of using such structures in linear accelerators. The results of extensive computations are given in families of curves, showing the dependence of the propagation constant and of the phase velocity on the parameters of chief interest. The power flow and field distribution are also discussed.

621.392.26†:621.317.34 1565
Power Adjustment for Plane Waves in Waveguides—H. F. Mataré. (*Frequenz*, vol. 4, pp. 321–328; December, 1950.) Description of attenuators and methods developed for their calibration and for sensitivity measurements in rectangular waveguides at 3-cm wavelength. Sliders form a capacitive (or inductive) discontinuity in the waveguide cross-section and are decoupled at either side by means of an adjustable resistance foil. A detector at the input measures the reflected energy; output power is measured by means of a bolometer. Reduction in power is plotted for progressive adjustments of the slider gap and comparison is made with theoretical values.

621.392.43 1566
Theoretical Limitations to Impedance Matching—R. L. Tanner. (*Electronics*, vol. 24, pp. 234, 242; February, 1951.) It is shown that many antennas can be represented with adequate accuracy by a simple RLC circuit. The optimum matching of such a circuit is considered in terms of the attainable voltage swr. A worked-out example shows good agreement with the theory.

621.396.67 1567
Effect of a Metal Mast and Guy Wires on the Performance of the 600-Ohm Multiple-Wire Delta Antenna—H. N. Cones. (*Bur. Stand. Jour. Res.*, vol. 46, pp. 113–120; February, 1951.) Describes measurements on scaled-down models used as receiving antennas. Curves show the terminal impedance, radiation patterns, and efficiency with and without guys for full-scale antennas at 1 to 25 mc. A metal mast causes little change in the vertical radiation pattern (14 to 25 mc) except for large side lobes at 15 mc. Continuous guys improve the radiation pattern; with wooden masts, the guys have no measurable effect on the input impedance, except where half a wavelength long, and the radiation efficiency is unaffected.

621.396.67:621.397.6 1568
Television Totem Pole—F. G. Kear and O. B. Hanson. (*Electronics*, vol. 24, pp. 66–70; February, 1951.) Existing television stations in New York City have been troubled by shadow areas and ghost signals due to nearby high buildings. This is being overcome by the use of five antenna arrays mounted one above the other on a 222 foot tower on top of the Empire State Building, initially for the use of five stations. Other stations may later share the same antennas, or new uhf antennas may be added. Each array has a power gain of about 4. Mutual interaction is small. The transmitters are located in the upper storeys of the building. See also 1261 of June.

621.396.67.012:621.317.3.087.6 1569
Polar Diagram Plotter—(*See* 1705.)

621.396.677 1570
On the Directional Patterns of Polystyrene Rod Antennas—R. B. Watson. (*Jour. Appl. Phys.*, vol. 22, pp. 154–156; February, 1951.) Increase in length from about 4λ to 10λ, while maintaining the ratio between the lengths of tapered and untapered parts, gives experimental values of apparent refractive index which are lower than indicated by theory and decrease with increasing length. The directional patterns are consequently less sharp than expected for the longer rods.

621.396.677:621.396.11 1571
The Theory of Parallel-Plate Media for Microwave Lenses—E. A. N. Whitehead. (*Proc. IEE* (London), Part III, vol. 98, pp. 133–140; March, 1951.) The reflection and refraction of a plane em wave at a plane interface, formed by the edges of an infinite set of equidistant parallel metal plates of negligible thickness and perfect conductivity, are studied. The solution presented extends the scope of those already published and the method is simpler than that previously used. From the formulas derived, curves of the power-transmission and power-reflection coefficients, and of the phase changes across the interface, have been computed. The formulas have also been tentatively applied to certain lenses, to derive practical limits to the design of efficient lenses.

621.396.677.029.64 1572
Parabolic-Cylinder Aerials—D. G. Kiely. (*Wireless Eng.*, vol. 28, pp. 73–78; March, 1951.) The method of design and the technique of construction of open parabolic-cylinder antennas is described. The maximum side-lobe level is 30 db below that of the main beam and represents a considerable improvement in performance over that of the cheese antenna, which is commonly used for navigational-radar sets. Comparison is also made with the performance of a more complicated type of "corrected" parabolic-cylinder antenna, which has been investigated by other workers.

621.396.677.5† 1573
Optimum Dimensions of a Stranded-Wire Loop Aerial for Reception—A. Colombani. (*Compt. Rend. Acad. Sci.* (Paris), vol. 232, pp. 708–709: February 19, 1951.) Results of the study of hf resistance of a helical stranded-wire coil (see 602 of 1950) are applied. Under the optimum conditions, the rf resistance is twice the dc resistance and is given by an expression independent of the nature of the metal used. The distributed capacitance decreases with the ratio of the volume of coil metal to that of its receptacle; to keep this low the deptn of winding may be increased, with corresponding decrease of the length, making the antenna a relatively flat coil of large mean diameter. The signal noise ratio is better for this form than for the antenna form.

621.396.67 1574
Antennas [Book Review]—J. D. Kraus. Publishers: McGraw-Hill, New York, N. Y., 1950, 553 pp., $8.00. (*Electronics*, vol. 24, p. 136; February, 1951.) "Intended for use as a text and reference for senior or graduate courses in antenna theory. As such, it serves admirably."

CIRCUITS AND CIRCUIT ELEMENTS

001.8 1575
The Sense of "Operational Analysis" and the use of "Operators" for the "Operational Diagrams" of "Operational Analysis"—W. Boesch. (*Microtecnic* (Lausanne), vol. 4, pp. 333–341; November and December, 1950.) "Operational analysis" is a method for studying the operation of measuring circuits, especially telemetering and remote-control circuits. A new form of presentation, "functional symbolism," is introduced; this is compared favorably with the block-diagram system of presentation, in respect of the amount of information it can convey regarding the functioning of apparatus. Elementary operational symbols are defined, and equivalent operational forms are given for numerous block-diagram symbols. Several complete systems are given in the new notation. The method facilitates estimation of minimum equipment required and recognition of superfluous apparatus included in a system.

621.3.011.2.018.12 1576
Impedances with Prescribed Variation of Phase Angle—E. Baumann. (*Z. Angew. Phys.*, vol. 1, pp. 43–52; January 15, 1950.) Presentation of a method of calculating impedances

whose phase shift is required to vary with frequency in a particular manner. The method is similar to that used in Cauer's filter theory, and involves the solution of a problem in the division of elliptical functions. Application to negative-feedback amplifiers is indicated.

621.3.015.7†:621.387.4† 1577

A Single-Channel Pulse Analyser for Nuclear Experiments—J. S. Eppstein. (*Jour. Sci. Instr.*, vol. 28, pp. 41–44; February, 1951.) Description, with detailed circuit diagram, of an instrument which accepts all pulses whose height is between two voltage levels, whose mean level is adjustable from 5 to 40 v, while their difference is constant.

621.3.018.4.012.1 1578

Frequency-Coordinate Vector Diagram—P. F. Ordung and H. L. Krauss. (*Jour. Frank. Inst.*, vol. 251, pp. 343–350; March, 1951.) Laplace-transform theory enables the response of a circuit to be expressed as a function of the complex variable $\alpha+j\omega$, where α is the decrement and ω the "real" angular frequency. Physically, the nature of the real-frequency response is usually not discernible by casual inspection of the mathematical function. A method is developed for visualizing this frequency response by means of a vector diagram constructed from the response function.

621.314.2.029.5 1579

Wide-Band High-Frequency Low-Power Transformers with Laminated Cores—H. Wilde. (*Frequenz*, vol. 4, pp. 305–314; December, 1950.) Detailed discussion of the optimum design of wide-band transformers for powers <1w. A toroidal core of high-μ metal tape minimizes leakage and winding capacitances. For sharp attenuation at the lower limiting frequency, this should be below the limiting frequency of the eddy currents in the core. For determination of the upper frequency limit, the leakage inductance and winding capacitance must be known. A method for calculating these approximately gives values in agreement with measurements. Winding capacitance, which reduces the upper frequency limit considerably when the transformation ratio is high, may be reduced by winding the primary in sector fashion.

621.314.25:621.392.4 1580

A Choke-Coupled Phase Invertor of High Accuracy—R. A. Seymour and D. G. Tucker. (*Electronic Eng.* (London), vol. 23, pp. 64–65; February, 1951.) A triode is used with mutually coupled balanced chokes in the anode and cathode, enabling the voltage-handling capacitance to be extended, while preserving good voltage balance and phase accuracy. Expressions are derived for gain, phase shift, and the distortion due to the flow of grid current.

621.314.3† 1581

A Theoretical and Experimental Study of the Series-Connected Magnetic Amplifier—H. M. Gale and P. D. Atkinson. (*Proc. IEE* (London), Part I, vol. 98, pp. 41–43; January, 1951.) Discussion on 2729 of 1949.

621.314.3† 1582

Magnetic Amplifiers—A. G. Milnes. (*Proc. IEE* (London), Part I, vol. 98, pp. 40–41; January, 1951.) Discussion on 2728 of 1949.

621.316.729:517.942.93 1583

Concerning Mathieu's Equation—J. Haag. (*Compt. Rend. Acad. Sci.* (Paris), vol. 232, pp. 661–663; February 19, 1951.) The stability of the periodic solutions of a Mathieu equation including small perturbation terms is investigated. The problem is a particular, simple case of Haag's general theory of synchronization of linear oscillators.

621.316.86 1584

The Quality of Resistors in Printed Circuits—G. Matthaes. (*Elektrotech. Z.*, vol. 71, pp. 105–107; March 1, 1950.) Printed circuits have been in commercial production in Italy for some time. Resistors manufactured by this process, with silicones as binder, are compared with carbon-layer resistors in respect of load-carrying capacitance, variation with humidity, temperature, time and applied voltage, noise, hf operation, and practical tolerances. The quality of the two types is found to be about the same.

621.318.572.015.7 1585

Multiple-Output Predetermined Counter—D. L. Gerlough and H. R. Kaiser. (*Electronics*, vol. 24, pp. 122, 184; February, 1951.) Detailed description of a counter circuit with a counting capacity of 2^n and with m predetermined points during the counting process where output signals can be obtained.

621.319.4 1586

Factors Affecting the Life of Impregnated-Paper Capacitors—H. F. Church. (*Proc. IEE* (London), Part III, vol. 98, pp. 113–122; March, 1951.) Discussion of various causes of failure in service.

621.392.43.012.3 1587

Minimum-Loss Matching Pads—J. C. Bregar. (*Electronics*, vol. 24, p. 118; February, 1951.) An abac simplifies determination of the two resistance values required in an L-type network for matching unequal impedances with minimum loss of power, and also gives, directly, the amount of loss in decibels.

621.392.5 1588

Ladder Networks. An Algebra for their Solution—W. E. Bruges. (*Electrician*, vol. 146, pp. 381–384 and 475–478; February 2 and 9, 1951.) Although the paper deals mainly with the application of ladder networks to the determination of the impedance of conductors in the rotor slots of induction motors, the method of solution is of general interest. Use is made of two simple mathematical series which can be written down and evaluated from easily constructed tables.

621.392.52 1589

The Properties of the Double-T Quadripole between Finite Resistances—Günther. (*Funk u. Ton*, vol. 4, pp. 628–642; December, 1950.) This type of circuit is particularly useful in feedback af amplifiers. Formulas are derived for input resistance as a function of frequency and output load, coupling coefficient, and so forth. These characteristics are presented as families of curves so that simple calculation is possible.

621.392.54† 1590

Single-Input Attenuators with Multiple Outputs—C. W. Ulrich. (*Electronics*, vol. 24, pp. 200, 209; February, 1951.) Design data for a network to feed multiple outputs from a single source so that all branches present equal impedances. The insertion loss involved is calculated.

621.395.813:621.392.53 1591

The Design of Equalizers by Synthesis Technique—C. F. Campbell. (*Strowger Jour.*, vol. 7, pp. 151–161; November, 1950.) A general discussion of the methods of design of networks with specified insertion-loss frequency characteristics. A new method of approximating to the required characteristic by use of the ratio of two polynomials with real coefficients is presented.

621.396.6.002.2:621.961 1592

Mass-Production Component and Circuit Die Stamping—R. G. Peters. (*TV Eng.*, vol. 2, pp. 10–11, 25; January, 1951.) Describes and indicates the advantages of die stamping of circuit wiring and various components from a sheet of conducting material, which is simultaneously cemented to a sheet of insulating material.

621.396.611.21:621.317.3 1593

Measurement of the Electrical Behaviour of Piezoelectric Resonators—Floyd and Corke. (*See* 1704.)

621.396.611.4 1594

Basis of Approximate Calculation of Electromagnetic Oscillations in Cavity Resonators—H. J. Mähly. (*Helv. Phys. Acta*, vol. 23, pp. 864–865; December 10, 1950. In German.) Brief discussion of variational methods.

621.396.611.4:537.525 1595

A Cylindrical Cavity Filled with a D.C. Discharge—G. W. Stuart, Jr. and P. Rosen. (*Jour. Appl. Phys.*, vol. 22, p. 236; February, 1951.) Treating the discharge as a lossy dielectric, formulas for the resonance frequency and for the Q of the cavity are derived.

621.396.611.4:621.316.726.078.3:
621.396.615.14 1596

A New Centimetre-Wave Discriminator and its Application to a Frequency-Stabilized Oscillator—K. C. Johnson. (*Proc. IEE* (London), Part III, vol. 98, pp. 77–80; March, 1951.) The construction of this discriminator is considerably simpler than previous types, in that a single resonant cavity and a single coupling loop are used. The discriminator action is obtained by arranging that the loop is self-inductive, so that the total impedance across it varies linearly with frequency near the cavity resonance. The performance is free from serious errors, due to changes of input power and drift effects in the crystal detector.

The application to the stabilization of a reflex klystron is described, in which errors in frequency are detected by the discriminator and are used to adjust the reflector voltage. Frequency stability to within 1 part in 10^6 for periods up to half an hour has been achieved.

621.396.615:512.83 1597

Determinantal Solution of Phase-Shift Oscillators—J. D. Tucker. (*Jour. Brit. IRE*, vol. 11, pp. 22–24; January, 1951.) The method is demonstrated by application to the calculation of a four-section RC oscillator using a high-slope hf pentode.

621.396.615.14.029.66 1598

Millimeter Waves—J. R. Pierce. (*Electronics*, vol. 24, pp. 66–69; January, 1951.) From many points of view, incoherent sources of radiation in the millimeter wavelength range, such as spark-excited oscillators or mass radiators, are unsatisfactory. The various types of source, at present available, are reviewed, and the difficulties in the way of producing powerful coherent sources of millimeter waves are pointed out. Some of the best results for the mm range have been obtained with pulsed magnetrons, but their efficiency falls very considerably for wavelengths below about 6 mm. Various possible alternative methods of producing such waves, including the double-stream tube, the multi-resonator klystron, and the "relativistic Doppler" method, are briefly discussed.

621.396.645:621.387.464† 1599

Distributed Coincidence Circuit—G. Wiegand. (*Rev. Sci. Instr.*, vol. 21, pp. 975–976; December, 1950.) "A coincidence circuit using the traveling-wave principle, as applied to distributed amplification, is described. The resolving time is about 10^{-8} seconds, when the device is used in connection with scintillation detectors."

621.396.645.015.7 1600

Transmission-Line-Reflection [voltage-] Doubling Amplifier—J. Marshall. (*Rev. Sci. Instr.*, vol. 21, pp. 1010–1013; December, 1950.) Description of a pulse amplifier using voltage doubling by reflection at the open end of a section of coaxial transmission line. Cathode followers are used to couple the output of one line to the input of the next. An amplifier, using in each stage the four triode units of two Type 6J6 tubes in parallel, driving 16 feet of 100-Ω line, has a theoretical voltage gain of 1.32 per stage. With appropriate grid series compensation, the gain can theoretically be made constant up to 100 mc. Experimentally, the gain is 1.3 per stage, with a rise time $<6\times10^{-9}$ sec. Below 50 v, distortion is not serious for pulses shorter than twice the transmission time of the line section. Pulses up to 150 v can be handled.

621.396.645.018.424†:621.396.828.1 1601
Wide-Band Amplifier for Central-Antenna Installations—J. B. Crawley. (*Electronics*, vol. 24, pp. 210, 214; February, 1951.) A two-stage high-gain wide-band amplifier, with cathode-follower output, is described for supplying multiple receivers in locations where high-noise level prevents satisfactory operation of ordinary receivers with built-in antennas.

621.396.645.35 1602
Determination of the Sensitivity Limit of Mains-Driven D.C. Amplifiers—H. Etzold and H. Jahn. (*Funk u. Ton*, vol. 4, pp. 605–618; December, 1950.) Noise sources in a tube amplifier and the effect of fluctuations on the output level in a compensated circuit are discussed. The measured sensitivity limit for a 1:1 signal noise ratio was 30 μv for the Brentano circuit and 28.4 μv for the feedback bridge circuit. These values, referred to previous sensitivity measurements, give equivalent input fluctuations of 3.5 μv and 3.8 μv, respectively, the theoretically calculated figure being 1.1 μv.

621.396.645.35.087.6 1603
A Frequency-Compensated Direct-Coupled Amplifier for Use with a Four-Channel Pen Recorder—J. A. Tanner and B. G. V. Harrington. (*Jour. Sci. Instr.*, vol. 28, pp. 33–35; February, 1951.) The amplifier has balanced stages giving a voltage gain of about 500; a phase inverter is used to provide single-ended output. A method of adjusting the individual dynamic response characteristics of several pen units, by frequency-compensation networks, is described; this results in response characteristics flat to within ± 3 per cent over the range 0 to 100 cps.

621.396.822 1604
Valve and Circuit Noise. Radio Research Special Report No. 10 [Book Review]—(*See* 1808.)

GENERAL PHYSICS

519.2:621.3.015.2 1605
On the First-Passage-Time Probability Problem—Siegert. (*See* 1694.)

535.214 1606
Radiation Pressure in a Refracting Medium—R. V. Jones. (*Nature* (London), vol. 167, pp. 439–440; March 17, 1951.) The radiation pressure of a beam of light incident on a mirror, suspended in a liquid, is found to be proportional to the refractive index of the liquid.

535.215.5 1607
Quenching Action of Long-Wave Radiation in Electrophotoluminescence: Part 1—Experimental Results—F. Vigean. (*Compt. Rend. Acad. Sci.* (Paris), vol. 232, pp. 819–820; February 26, 1951.)

535.215.5 1608
Quenching Action of Long-Wave Radiation in Electrophotoluminescence: Part 2—Tentative Explanation—F. Vigean and D. Curie. (*Compt. Rend. Acad. Sci.* (Paris), vol. 232, pp. 955–957; March 5, 1951.)

535.222+621.396.11 1609
Proposed New Value for the Velocity of Light—L. Essen. (*Nature* (London), vol. 167, pp. 258–259; February 17, 1951.) Evidence is accumulating, particularly from rf measurements, that the velocity of light may be significantly greater than the value hitherto accepted, which is derived mainly from optical measurements. The value 299 to 790 km is suggested for adoption until measurements of still greater accuracy have been made. See also 1751 of 1950 and 1225 of June.

535.623:621.397.5 1610
Color Fundamentals for TV Engineers—D. G. Fink. (*Electronics*, vol. 23, pp. 88–93; December, 1950. Vol. 24, pp. 78–83 and 104–109; January and February, 1951.) Reproduced from the forthcoming second edition of "Principles of Television Engineering." An elementary explanation of basic principles, including discussion of additive and subtractive methods of color matching, the trichromatic system, the RGB color diagram, and its transformation to the XYZ form, color specification by dominant wavelength, specification of color distortion in a television system and its relation to brightness distortion, the color standards in the "reference receiver," congruence requirements in the primary images, and the color-transfer process.

536.7:621.315.615 1611
Thermodynamics of a Fluid Dielectric—W. B. Smith-White. (*Nature* (London), vol. 167, pp. 401–402; March 10, 1951.)

537.221 1612
The Volta Effect as a Cause of Static Electrification—W. R. Harper. (*Proc. Roy. Soc. A*, vol. 205, pp. 83–103; January 22, 1951.) An experimental investigation of the static electrification of metal-metal surfaces indicates that the apparent inconsistency of earlier results was due to ignorance of some of the operative factors. On paying attention to surface topography and allowing for the transfer of electrons by tunnel effect, a precise theory of separation charging is derived, giving results in agreement with experiments.

537.311.5+537.311.62 1613
Skin Effect and Current Distribution in Conductors at High Frequencies—F. Benz. (*Elektrotech. u. Maschinenb.*, vol. 67, pp. 366–374; December, 1950.) In conductors with non-circular cross section or with current-carrying leads in proximity, the effect of the current distribution, as distinct from the skin effect, leads to an increase in resistance, due to the higher current density in the region of greater field strengths at the conductor surface. Practical examples are discussed.

537.315.6:517.942.9 1614
On Some Dual Integral Equations occurring in Potential Problems with Axial Symmetry—C. J. Tranter. (*Quart. Jour. Mech. Appl. Math.*, vol. 3, pp. 411–419; December, 1950.) Hankel transforms are used to reduce the solution of Laplace's equation under given conditions, to the solution of a pair of dual integral equations. A formal solution of these equations is given and, as an example, is applied to find the potential due to a circular disk, parallel to and equidistant from two earthed parallel plates.

537.52 1615
The Part Played by the Electrodes in the Starting of Electric Discharges in Gases—F. L. Jones, E. T. de la Perrelle, and C. G. Morgan. (*Compt. Rend. Acad. Sci.* (Paris), vol. 232, pp. 716–718; February 19, 1951.) Measurements were made of the electron emission, for different surface conditions, from cold Ni and W cathodes subjected to a mean field of the order of 10^6 vcm in air at atmospheric pressure. Oxidation increased the emission. Effects due to residual positive ions were also studied.

537.527 1616
The Influence of the Electrodes on the Discharge between Coaxial Cylinders in Compressed Air—W. Bright. (*Compt. Rend. Acad. Sci.* (Paris), vol. 232, pp. 714–716; February 19, 1951.) Experiments show that the breakdown voltage depends on the metals of which the two cylinders are composed; it is generally lower when the inner is cathode. The theoretical explanation of the mechanisms involved is under investigation.

538.51 1617
Electromagnetic Induction in a Semi-infinite Conductor with a Plane Boundary—A. T. Price. (*Quart. Jour. Mech. Appl. Math.*, vol. 3, pp. 385–410; December, 1950.) The general theory is considered; solutions of the fundamental equations are of two types, corresponding to two distinct modes of decay of currents. Detailed calculations are made of the induced field and currents produced by a periodic and aperiodic line current, parallel to the face of the conductor, and the method of calculation for other inducing fields is indicated. Applicability of the theory to geophysical problems is discussed.

538.56:535.312:537.525.5 1618
Radio Reflexions from a Column of Ionized Gas—D. Romell. (*Nature* (London), vol. 167, p. 243; February 10, 1951.) The reflection of 30-cm radio waves from a mercury-discharge tube of diameter 3.2 cm was studied. With the electric vector perpendicular to the axis of the column, strong reflections were obtained when resonance occurred at a discharge current of about 0.9 a, compared with a predicted value of 0.8 a. The reflected power was 80 per cent of that reflected from a metal strip of width $\lambda/2$. With the plane of polarization parallel to the column, no reflected wave could be detected. See also 96 of February (Denno et al.).

538.56:535.42 1619
Note on Diffraction by an Edge—D. S. Jones. (*Quart. Jour. Mech. Appl. Math.*, vol. 3, pp. 420–434; December, 1950.) If the electric current and charge on a perfectly conducting sheet are integrable in the neighborhood of the edge, the line distributions of electric current and charge on the edge are determined by the boundary conditions on the sheet. Assuming that there are no line distributions of magnetic current and charge, the field is then determined uniquely. When the currents near the edge can be expanded in power series, the index of the first term is $(2p+1)/2$, where p is an integer, greater than -2 or -1, according as the current parallel or perpendicular to the edge is considered. The diffraction, in three dimensions, of a plane wave by a semi-infinite plane is also considered.

538.56:535.42.001.11 1620
A Method for the Exact Solution of a Class of Two-Dimensional Diffraction Problems—P. C. Clemmow. (*Proc. Roy. Soc. A*, vol. 205, pp. 286–308; February 7, 1951.) Representation of the scattered field as an angular spectrum of plane waves leads directly to a pair of dual integral equations, which replace the single integral equation of Schwinger's method. The unknown function in each of these dual integral equations is that which defines the angular spectrum, and when this function is known, the scattered field is presented in the form of a definite integral. As far as the radiation field is concerned, this integral is of the type which can be approximately evaluated by the method of steepest descents, though in

certain circumstances it is necessary to generalize the usual procedure.

The method is appropriate to two-dimensional problems in which a plane wave, of arbitrary polarization, is incident on plane perfectly conducting structures; for certain configurations, the dual integral equations can be solved by the application of Cauchy's residue theorem. The three problems considered are those for which the diffracting plates, situated in free space, are, respectively, a half-plane, two parallel half planes, and an infinite set of parallel half planes, the second case being illustrated by a numerical example. Several points of general interest in diffraction theory are discussed, including the question of the nature of the singularity at a sharp edge, and, it is shown, that the solution for an arbitrary three-dimensional incident field can be derived from the corresponding solution for a two-dimensional incident plane wave.

538.566 1621

On the Propagation of Energy in Linear Conservative Waves—L. J. F. Broer. (*Appl. Sci. Res.*, vol. A2, nos. 5/6, pp. 329–344; 1951.) The method of stationary phase is applied to show that whenever this method is applicable, the rate of energy propagation in a system of waves without dissipation is equal to the group velocity. The reason why this equality can be established by this kinematical method is examined by discussion of simple harmonic waves. The choice of an expression for the energy density, to be used in connection with a given wave equation, is shown to be restricted by the conservation of energy in such a way that the ratio of the average rate of work to the average energy density always equals the group velocity. Examples of wave motion are discussed to illustrate the use of the formulas derived.

538.566.001.11:523.72 1622

The Interpretation of the Magneto-ionic Theory—K. C. Westfold. (*Jour. Atmos. Terr. Phys.*, vol. 1, no. 3, pp. 152–186; 1951.) Discussion of the implications of the Appleton-Hartree formula when collisions are neglected, and study of particular approximate solutions of the quadratic equation that determines the complex-refractive index and the polarization. The four cases considered are those where the collision frequency or the gyrofrequency is small or large, compared with the wave frequency. The most fruitful approximation is for the case where the collision frequency is small, compared with the wave frequency. The solutions then provide sufficient data for a set of curves to be drawn, which present a complete mathematical picture of the associated complex dielectric constant and the polarization. From these curves others are derived, showing the effects of different physical conditions on the refractive index and absorption coefficient.

Applications of the magneto-ionic theory to propagation in the solar atmosphere are critically discussed, and formulas are given for the intensity of the emergent thermal radiation, whose frequency is large compared with the gyrofrequency of the medium.

621.3.001.4:513.344 1623

Calculation of the Electrical Capacitance of a Cube—D. K. Reitan and T. J. Higgins. (*Jour. Appl. Phys.*, vol. 22, pp. 223–226; February, 1951.) "The basic theory of calculation of the capacitance of a given geometrical configuration by the use of sub-areas is advanced and applied to solve the long-standing problem of the accurate evaluation of the capacitance C of a cube of side a. The best previously published determination is $0.62211a < C < 0.71055a$. The value obtained of $C \approx 0.655a$ esu is both a lower limit and very close to the exact value."

537.523.3/.4 1624

Corona and Breakdown at Frequencies up to 12 Mc/s. Technical Report Reference L/T229 [Book Review]—A. W. Bright. Publishers: The British Electrical and Allied Industries Research Association, London, Eng., 1950, 24 pp., 12s. (*Beama Jour.*, vol. 58, p. 57; February, 1951.) Previous work on breakdown in air at frequencies up to about 100 mc is reviewed, experiments on hf breakdown between a thin wire and a concentric tube are described, and critical-gap phenomena with 2-cm spheres in air, N_2, O_2, and freon are studied. A simple theory is put forward to explain some of the results.

GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

523.5:621.396.9 1625

Deceleration and Ionizing Efficiency of Radar Meteors—D. W. R. McKinley. (*Jour. Appl. Phys.*, vol. 22, pp. 202–214; February, 1951.) Improved instrumentation of the radar system permitted more accurate analysis of range-time records of meteor echoes. The meteor deceleration could be determined in special cases, mean decelerations of 0.48, 1.1, and 1.5 km per sec. being deduced. Velocity data from a cw Doppler system were compared with radar data for the meteors observed. Using existing atmospheric-density values and Lovell's scattering formula, the ionizing efficiency is computed to be 10^{-6} for a 60-km meteor and 10^{-8} for a 20-km meteor. For lower values of air density efficiency figures are increased and rates of electron production, deduced from loss-of-mass considerations, agree better with rates calculated from radar data. Further statistical data and information, concerning the ionizing properties of 1-kev atoms, are required before definite conclusions can be drawn.

523.72:538.566.001.11 1626

The Interpretation of the Magneto-Ionic Theory—Westfold (*See* 1622.)

523.72+523.8]:621.396.822 1627

Noise from Extra-Terrestrial Sources—H. Rakshit. (*Sci. Culture*, vol. 16, pp. 293–297; January, 1951.) A brief survey of experimental evidence accumulated and explanatory theories advanced by workers from 1931 onwards.

523.72:621.396.822 1628

Equivalent Path and Absorption for Electromagnetic Radiation in the Solar Corona—H. C. Jaeger and K. C. Westfold. (*Aust. Jour. Sci. Res., Ser.*, vol. 3, pp. 376–386; September, 1950.) "Calculations of the trajectories, equivalent path, and absorption of rays in the frequency range 20 to 100 mc in the solar corona have been made, neglecting possible magnetic fields and assuming spherical symmetry. Interpreting the double-humped burst of solar noise, as the superposition of a direct and an echo signal, inferences are made as to the height in the corona and location on the solar disk of its source."

523.72:621.396.822 1629

Observations of the Spectrum of High-Intensity Solar Radiations at Metre Wavelengths: Part 1—The Apparatus and Spectral Types of Solar Burst Observed—J. P. Wild and L. L. McCready. (*Aust. Jour. Sci. Res., Ser.*, vol. 3, pp. 387–398; September, 1950.) "An apparatus for recording the dynamic spectrum of high-intensity solar radiation (in particular the sudden bursts) in the frequency range 70 to 130 mc is described. The spectra are displayed on a cathode-ray tube at intervals of about one third of a second. Solar bursts observed with the apparatus were found to have widely different spectra. However, analysis of a number of bursts indicated the common occurrence of three distinct spectral types. These types are described and illustrated by samples. One type of narrow bandwidth was exhibited by short-lived bursts that occur in large numbers during periods of high intensity ("noise storms"); these bursts are presumed to be circularly polarized and associated with sunspots. A second type, characterized by a slow drift of spectral features towards the lower frequencies, was exhibited by sporadic "outbursts" associated with solar flares. Other sporadic bursts had diverse spectra, but some of them conformed to a third spectral type, in which the frequency of maximum intensity drifts rapidly towards the lower frequencies. The result, that outbursts seem to exhibit a distinct type of spectrum, is considered to provide a possible means of recognizing these phenomena with certainty."

523.72:621.396.822 1630

Observations of the Spectrum of High-Intensity Solar Radiation at Metre Wavelengths: Part 2—Outbursts—J. P. Wild. (*Aust. Jour. Sci. Res., Ser.*, vol. 3, pp. 399–408; September, 1950.) Observations of the spectrum of outbursts of solar rf radiation in the range 70 to 130 mc are described. In accordance with part 1, an "outburst" is defined as a burst having a particular type of "dynamic" spectrum, characterized by a drift of spectral features, with time, towards the lower frequencies at a rate of the order of $\frac{1}{4}$ mc per second. The observed outbursts have a close connection with solar flares and their geophysical accompaniments. The spectra are tentatively interpreted in terms of the motion of a physical agency in the solar atmosphere. The possible identification of the agency with "surge" prominences, and the corpuscular streams that cause a type of terrestrial magnetic storm is discussed. The evidence is quite consistent with the hypothesis that the agency corresponds to the magnetic-storm particles.

523.746:550.385 1631

Solar Activity and Geomagnetic Storms, 1950—H. W. Newton and H. F. Finch. (*Observatory*, vol. 71, pp. 45–47; February, 1951.) A brief review of conditions during 1950. Provisional mean daily sunspot numbers for each month, details of the larger sunspot groups (which suggest the existence of favored regions for spot formation or continuance) and data dealing with the magnetic storms recorded at Abinger, are included. The decline of solar-flare occurrence is noted.

523.752.001.572 1632

Ultra-Violet Emission from the Chromosphere—R. v. d. R. Woolley and C. W. Allen. (*Mon. Not. R. Astr. Soc.*, vol. 110, no. 4, pp. 358–372; 1950.) A model of the quiet chromosphere is described, which has been fitted to eclipse data and to observations of solar radio noise and of the ionosphere, all taken at minimum solar disturbance. The model has spherical symmetry and a single value of the kinetic temperature at every height. There is a sharp division at 6,000 km; below this height the temperature is 5,040° K, above it the temperature increases with height, very rapidly at first. The number of quanta emitted by the chromosphere, capable of ionizing terrestrial gases, is estimated at 7×10^{14} per cm² per sec. About half have an energy >13.6 v, and are, therefore, capable of ionizing O to O+ in the ionosphere. The energy is supplied by conduction inwards from the corona.

523.78 1633

Alaskan Eclipse Expedition—(*Tech. Bull. Nat. Bur. Stand.*, vol. 35, pp. 17–19; February, 1951.) Modern radio-astronomy methods were applied in this expedition to observe the total eclipse of the sun by the moon, September 11, 1950. A mirror of diameter 10 feet and focal length 3 feet was used to collect the solar rf energy. Automatic records were made of intensity time for wavelengths of 3, 10, and 65 cm, starting about 2 hours before the eclipse. Delay of the recorded minimum with respect to time of total optical eclipse is attributed to sunspot asymmetry of the corona. The cause of unexpected increases of intensity at first and fourth contact is under investigation.

523.854:551.510.535:621.396.822 1634
The Effects of the Terrestrial Ionosphere on the Radio Waves from Discrete Sources in the Galaxy—M. Ryle and A. Hewish. (*Mon. Not. R. Astr. Soc.*, vol. 110, no. 4, pp. 381–394; 1950.) "Observations of the discrete sources of radio waves in the galaxy have shown the existence of irregular refraction processes, in the terrestrial ionosphere. These irregularities cause rapid fluctuations in the intensity of the radiation at the ground, while observations with antennas of high resolving power have shown, in addition, that the apparent position of a source may vary irregularly by 2 to 3 minutes of arc. The incidence of these irregularities shows a marked diurnal variation, having a maximum at about $01^h 00^m$ local time.

It does not seem possible to account for the irregularities in the ionosphere in terms of solar emissions, and an alternative mechanism is proposed, which is based on the interception of interstellar matter moving under the gravitational attraction of the sun. If this hypothesis is correct, further experiments may provide information of interest in theories of the accretion of matter by the sun."

537.591:523.752 1635
The Change of Cosmic-Ray Neutron Intensity following Solar Disturbances—J. A. Simpson, Jr. (*Phys. Rev.*, vol. 81, pp. 639–640; February 15, 1951.)

537.591:621.396.812.3 1636
The Cosmic-Ray Intensity and Radio Fade-Outs—Dolbear, Elliot, and Dawton. (*See* 1763.)

538.12:521.15 1637
Gravitation and Magnetism—E. A. Milne. (*Mon. Not. R. Astr. Soc.*, vol. 110, no. 4, pp. 266–274; 1950.) "It is shown, by the methods of kinematic relativity, that there should be a connection between gravitation and magnetism of the type suggested by the empirical formulas of Blackett and Wilson, multiplied, however, by certain dimensionless ratios."

550.38 1638
The Terrestrial Magnetic Field—H. Manley. (*Research* (London), vol. 4, pp. 43–44; January, 1951.) Results of research on the thermomagnetic properties of rocks are reported briefly and discussed. They support the theory that the earth's magnetic field has remained substantially constant throughout geological time. The reverse magnetization of intrusions is usually attributed to the action of a local secondary field in the remote past, possibly due to a Bullard-Elsasser core vortex.

550.385+551.594.5 1639
Notes on Aurorae and Magnetic Storms—S. Chapman. (*Jour. Atmos. Terr. Phys.*, vol. 1, no. 3, pp. 189–199; 1951.) Suggestions for a laboratory experiment to throw light on the production of auroras and magnetic storms.

551.510.4:546.214 1640
Vertical Distribution of Atmospheric Ozone—V. H. Regener. (*Nature* (London), vol. 167, pp. 276–277; February 17, 1951.) Results obtained from free-balloon flights to heights of 30 to 32 km are reported and discussed.

551.510.5:621 396.11:532.517.4 1641
Spectrum of Atmospheric Turbulence—L. F. Richardson: E. C. S. Megaw. (*Nature* (London), vol. 167, p. 318; February 24, 1951.) Comment on 973 of May (Megaw) and author's reply.

551.510.53:001.4 1642
Upper Atmospheric Nomenclature—S. Chapman. (*Jour. Atmos. Terr. Phys.*, vol. 1, no. 3, p. 201; 1951.) Addendum to 876 of May.

551.510.53:001.4 1643
Nomenclature of the Upper Atmosphere—N. C. Gerson and J. Kaplan. (*Jour. Atmos. Terr. Phys.*, vol. 1, no. 3, p. 200; 1951.) Comment on 876 of May (Chapman).

551.510.535 1644
Effects of the Atmospheric Scale-Height Gradient on the Variation of Ionization and Short-Wave Absorption—M. Nicolet. (*Jour. Atmos. Terr. Phys.*, vol. 1, no. 3, pp. 141–146; 1951.) A discussion of layer formation processes in an atmosphere in which the scale height varies linearly with height, the electron recombination coefficient also varying with height. The rate of variation of the critical frequency, and of the total hf absorption of the layer with the solar zenith angle are tabulated. The experimental data available may be explained on the basis of some of the atmospheric models considered.

551.510.535 1645
Thermal Splitting of Ionosphere Layers—In 879 of May the journal reference should read "*Arch. Met. Geoph. Bioklimatol. A.*"

551.510.535:535.361.2:621.396.81 1646
The Scattering of Radio Waves—Dieminger. (*See* 1762.)

551.510.535:550.385 1647
Investigation of the World-Wide Ionospheric Disturbance of March 15, 1948—In 889 of May the journal reference should read "*Arch. Met. Geoph. Bioklimatol. A.*"

551.578.1:621.396.9 1648
Radar Observations of Rain and their Relation to Mechanisms of Rain Formation—E. G. Bowen. (*Jour. Atmos. Terr. Phys.*, vol. 1, no. 3, pp. 125–140; 1951.) The two types of radar echoes from rain, one showing the "bright band," the other of the columnar type, were investigated using both ground and airborne radar. The results suggest that two distinct types of rain formation are concerned.

551.594.5:551.510.535 1649
Reflexion of High Frequencies during Auroral Activity—D. Davidson. (*Nature* (London), vol. 167, pp. 277–278; February 17, 1951.) Observations have been made at Concord, Massachusetts, of auroral echoes at frequencies of 17.3 and 3.5 mc. Echoes corresponding to long slant ranges are obtained, indicating that the centers of reflection are at a considerable horizontal distance rather than at great heights.

551.594.51:537.13 1650
Protons and the Aurora—C. W. Gartlein. (*Phys. Rev.*, vol. 81, pp. 463–464; February 1, 1951.) Spectrograms of an auroral arc over Arnprior, Ontario, obtained simultaneously on September 30, 1950, at Arnprior and at Ithaca, New York, 330 km apart, indicate that hydrogen, in the form of protons, was approaching the earth with a mean velocity of 675 km and a maximum velocity of 1,350 km. See also 636 of April (Meinel).

551.594.6 1651
•The Statistical Action of Atmospherics on a Receiver Tuned to 27 kc/s—F. Carbenay. (*Compt. Rend. Acad. Sci.* (Paris), vol. 232, pp. 949–950; March 5, 1951.) The rhythm of succession of atmospherics capable of actuating a recorder tuned to 27 kc appears to be independent of the bandwidth of the recorder, provided the threshold level is defined by a pulse of duration small compared with the period of the receiver (37 μs). See also 2282 of 1950 and back references.

551.594.6:621.396.11 1652
The Propagation of a Radio-Atmospheric—K. G. Budden. (*Phil. Mag.*, vol. 42, pp. 1–19; January, 1951.) The propagation of an atmospheric may be considered theoretically in terms of multiply reflected rays or in terms of the permitted modes in the waveguide, formed by the earth and the ionosphere. Using the latter method, the assumption of perfectly conducting boundaries results in the "zero-order" mode being unattenuated at all frequencies, in disagreement with observation. By considering the ionosphere as a homogeneous medium of complex-refractive index, the mode propagation is modified in such a way that the zero-order mode undergoes marked attenuation, at frequencies below about 8 kc. General agreement between the theoretical and observed curves of attenuation versus frequency, at a distance of 1,000 km, is obtained at all frequencies below 16 kc, using only two modes. The frequency below which attenuation increases rapidly depends markedly on the height of the ionosphere, and the attenuation is found to change diurnally and during sudden ionospheric disturbances in agreement with observed height changes.

The deduced waveform at 1,000 km is oscillatory, with decreasing frequency and amplitude, corresponding to types often observed in practice. The equivalence of the ray and mode pictures is demonstrated for the case of a large ionospheric reflection coefficient. Possible extensions of the theory are suggested to deal with various ion-density distributions and to include the effect of the earth's magnetic field.

LOCATION AND AIDS TO NAVIGATION

621.396.9 1653
Visibility of Radar Echoes—A. W. Ross. (*Wireless Eng.*, vol. 28, pp. 79–92, March, 1951.) A method is described for calculating the ratio of signal power to noise power required for any given probability of detection, particularly when the number of signal pulses producing the echo paint is small. The approximations used to obtain a simple solution are unlikely to introduce serious errors, though the "criteria of recognition," employed to express the behavior of an operator in mathematical terms, is open to discussion. Numerical solutions are included which cover a fairly wide range of conditions, and these are used to examine the influence of various factors on the detectability of a weak signal. Biasing and limiting are also considered.

621.396.9:523.5 1654
Deceleration and Ionizing Efficiency of Radar Meteors—McKinley. (*See* 1625.)

621.396.9:526.9 1655
Datum Stabilizer for Radar-Altimeter Surveying—B. I. McCaffrey. (*Electronics*, vol. 24, pp. 100–103; February, 1951.) To correct for altitude fluctuations when using an airborne profile recorder, an electronic circuit operating from an aneroid element senses deviations of the aircraft from level flight, and automatically applies corrections to the radar record of terrain elevation.

621.396.9:551.578.1 1656
Radar Observations of Rain and their Relation to Mechanisms of Rain Formation—Bowen. (*See* 1648.)

621.396.9:551.594.22 1657
A Radar Echo from Lightning—I. C. Browne. (*Nature* (London), vol. 167, p. 438; March 17, 1951.) An echo on λ3.2 cm of duration 2 ms was observed during a thunderstorm. If it was due to reflection from a column of electrons, the concentration must have exceeded 5×10^{13} electrons per cm length.

621.396.933 1658
Automatic G.C.A.—J. T. McNaney. (*Electronics*, vol. 24, pp. 82–87; February, 1951.) Using conventional-precision approach radar, a fully automatic system capable of controlling the approaches of five aircraft simultaneously can be developed. A full analysis of the system is given, describing the scanning systems, target selection by virtue of range and speed, and presentation of elevation, azimuth, and range indications as visual displays or as control signals to an automatic pilot.

621.396.933 1659
Distance-Measuring Equipment for Civil Aircraft. Airborne Apparatus and Ground Beacons—D. G. Lindsay, J. P. Blom, and J. D. Gilchrist. (*AWA Tech. Rev.*, vol. 9, pp. 1–41; January, 1951.) Substance of a lecture at the Radio Engineering Convention, Melbourne, Australia, May, 1950. The main features of the system are outlined and its operation is explained. The receiver, video, and monitor sections of the ground beacon are described, with details of the arrangement and performance of the various units and description of the modulator and transmitter, and also of the antenna system. An account is included of test instruments developed for periodic checks of the equipment. See also 645 of 1950 (Busignies).

MATERIALS AND SUBSIDIARY TECHNIQUES

531.788.7 1660
The Physical Basis of the Residual-Vacuum Characteristic of a Thermionic Valve—G. H. Metson. (*Brit. Jour. Appl. Phys.*, vol. 2, pp. 46–48; February, 1951.) The residual-reverse grid current in a hard tube is found to be due to the effects of soft X rays, which give rise to low-energy photoelectric emission and positive ions. An electrode system is described which enables the photoelectric emission to be suppressed. Using the principles described, the minimum pressure measurable by an ionization gauge has been reduced from 10^{-6} mm to 10^{-9} mm of Hg.

533.56 1661
The Design of Molecular Pumps—R. B. Jacobs. (*Jour. Appl. Phys.*, vol. 22, pp. 217–220; February, 1951.) A theoretical treatment of the operation characteristics of molecular drag-type pumps. The treatment is based on kinetic theory and is limited to high-vacuum operation. The formulas derived are simple and may be applied readily to practical pump design.

535.215.1:[546.32+546.33 1662
New Aspects of the Photoelectric Emission from Na and K—J. Dickey. (*Phys. Rev.*, vol. 81, pp. 612–616; February 15, 1951.) "The external photoelectric effect for Na and K in the form of thick evaporated layers was investigated with photon energies $h\nu$ up to 6.71 ev. When $h\nu$ was within 1.5 ev of the threshold, the energy distributions of the emitted electrons were like those derived by DuBridge for a simple photoelectric effect involving an ideal metal. For larger $h\nu$, there was a growing preponderance of low-velocity electrons."

535.215.1:546.431-3 1663
The Enhanced Photoelectric Emission Effect in Barium Oxide—B. D. McNary. (*Phys. Rev.*, vol. 81, pp. 631–632; February 15, 1951.) The spectral distribution of the photoemission from BaO, before and after irradiation at a wavelength of 3,700 Å, is shown graphically. Curves showing the decay of enhanced photoemission at 300°K at various wavelengths indicate that wavelengths near the threshold, for which the enhancement effect is most pronounced, have the greater decay rate. See also 365 of March (Dickey & Taft).

535.37:546.472.21 1664
Photoluminescence Efficiency of ZnS-Cu Phosphors as a Function of Temperature—R. H. Bube. (*Phys. Rev.*, vol. 81, pp. 633–634; February 15, 1951.) Additional results are given for the variation of the efficiency with temperature during excitation by ultraviolet light of wavelength 3,650 Å. See also 648 and 649 of April.

537.311.33 1665
Electronic Conduction in Non-Metals—G. Busch. (*Z. Angew. Math. Phys.*, vol. 1, pp. 3–31 and 81–110; January 15 and March 15, 1950.) A comprehensive review of the subject from both the theoretical and experimental viewpoint, including recent researches. There are 131 references given.

537.311.33 1666
Solid State Electronics—(*Tech. Bull. Nat. Bur. Stand.*, vol. 35, pp. 22–27; February, 1951.) General account of research on semiconductors in progress at the National Bureau of Standards. Measurements of Hall coefficient and conductivity of *p*-type germanium as functions of temperature are plotted down to low values of temperature. Work on TiO_2 and "grey tin" is mentioned.

537.311.33:541.183.26 1667
Adsorption on Semiconductors—P. Aigrain, C. Dugas, and J. Germain. (*Compt. Rend. Acad. Sci.* (Paris), vol. 232, pp. 1100–1101; March 12, 1951.) The adsorption of certain gases on certain semiconductors can be explained by purely electrostatic phenomena. The calculated adsorption energies and the number of adsorbable atoms are in good agreement with experimental results.

537.311.33:546.289 1668
Measurement of Hole Diffusion in *n*-Type Germanium—F. S. Goucher. (*Phys. Rev.*, vol. 81, p. 475; February 1, 1951.) Measurements were made of the voltage picked up by a probe located at a point along a Ge rod when a spot of light was focused on to the rod at various distances from the probe. From the results, the lifetime of holes is calculated to be 18 μs, a value in agreement with other determinations.

537.311.33:621.315.592† 1669
Theory and Experiment for a Germanium *p-n* Junction—F. S. Goucher, G. L. Pearson, M. Sparks, G. K. Teal, and W. Shockley. (*Phys. Rev.*, vol. 81, pp. 637–638; February 15, 1951.) The junctions investigated were produced in *n*-type Ge by the addition of Ga, so that one portion of the single crystal was *p* type. Germanium *p-n* junctions closely obey the theoretical rectification law $I = I_s[\exp(ev/kt) - 1]$. Measurements of photo response, as a function of the distance of the illuminated point from the junction, yield values for the lifetimes of injected electrons and holes. Confirmatory values are obtained by measurements of the junction admittance as a function of frequency. See also 1682 below (Teal et al.).

537.533:[546.78.26+546.77.26 1670
Electron Emission Measurements on Carburized Tungsten and Molybdenum—E. Baş-Taymaz. (*Z. Angew. Math. Phys.*, vol. 2, pp. 49–51; January 15, 1951.) The results of measurements at temperatures up to about 2,300° K are shown graphically. Richardson's law is followed, though in the case of mo a sudden increase of emission to a value over ten times greater occurred at 1,810° K, with a corresponding sudden decrease at 1,940° K. With falling temperature, the increase at 1,810° K was maintained to 1,720° K. Corresponding resistivity changes were noted. These jumps are probably related to phase changes of the crystal structure.

538.221 1671
Isothermal Remanent Magnetization of Finely Granulated Magnetite—J. Roquet. (*Compt. Rend. Acad. Sci.* (Paris), vol. 232, pp. 946–948; March 5, 1951.)

538.221 1672
Precipitation and the Domain Structure of Alnico 5—A. H. Geisler. (*Phys. Rev.*, vol. 81, pp. 478–479; February 1, 1951.)

538.221 1673
Magnetic Properties of Mixed Ferrites of Magnesium and Zinc—C. Guillaud. (*Compt. Rend. Acad. Sci.* (Paris), vol. 232, pp. 944–946; March 5, 1951.) Results for ferrites with graded proportions of MgO and ZnO are shown graphically, and illustrate the effect of heat treatment on the magnetic properties and on the Curie temperature.

538.221 1674
Ferromagnetic Resonance in Various Ferrites—W. A. Yager, F. R. Merritt, and C. Guillaud. (*Phys. Rev.*, vol. 81, pp. 477–478; February 1, 1951.) Presentation and discussion of results of experiments on spherical polycrystalline specimens of ferrites of Mg, Ni, Co, Mn, and Mn-Zn. An operating frequency of 24.164 kmc was used.

538.221:539.23 1675
Thin Ferromagnetic Films—M. J. Klein and R. S. Smith. (*Phys. Rev.*, vol. 81, pp. 378–380; February 1, 1951.) The dependence of the spontaneous magnetization of these films on temperature and thickness is studied by means of the Bloch spin-wave theory. Summary noted in 2486 of 1950.

538.221:548.2:539.374 1676
The Effects of Plastic Deformation on Magnetic Properties of Polycrystalline Metals—U. M. Martius. (*Canad. Jour. Phys.*, vol. 29, pp. 21–31; January, 1951.)

538.221:621.318.323.2:621.396.662.22
.029.62 1677
Cores for U.S.W. Variometers—M. Kornetzki and J. Brackmann. (*Frequenz*, vol. 4, pp. 318–320; December, 1950.) Measurements on variometer coils with cores of different ferrite materials are reported and the results shown graphically.

538.632:546.48-31 1678
Hall Constant of Cadmium Oxide—C. A. Hogarth. (*Nature* (London), vol. 167, pp. 521–522; March 31, 1951.) An account of measurements made on compressed-powder specimens, of dimensions 4 cm × 1 cm × 0.1 cm and apparent density 5.5 gm per cm^3. The results indicate that conduction is by electrons. The influence of temperature and oxygen pressure was investigated.

541.183.26 1679
Gases and Metals—H. Lepp. (*Le Vide*, vol. 3, pp. 433–441 and 463–468; May and July, September, 1948.) The phenomena of adsorption and absorption of gases by metals are discussed from the physico-chemical viewpoint. The usefulness of the method of thermodynamic analysis for investigating metal-gas systems is examined, with particular reference to problems encountered in vacuum technique.

546.281.26:[537.323+536.41+537.311.32
1680
Physical Properties of Carborundum below 1000°C—A. Schulze. (*Elektrotechnik* (Berlin), vol. 4, pp. 326–328; September, 1950.) Results of measurements of (a) thermoelectric force of SiC/cekas (a high-resistivity Ni-Cr-Fe alloy) and SiC/Pt couples, (b) linear expansion coefficient, (c) electrical resistivity, are shown graphically for temperatures from room temperature up to 1,000°C. Anomalies are found in all cases around 560°C. and are probably related to the occurrence of an allotropic modification near this temperature.

546.289 1681
Germanium.—Recorder II—(*Metal Ind.* (London), vol. 78, pp. 151, 153; February 23, 1951.) A short general account of the properties of Ge. Until recently, this material was obtained mainly from sludges produced during the purification of the electrolytes used in the manufacture of electrolytic zinc in the United States, but during the past few years methods have been developed for recovering Ge from the flue dusts of certain British gasworks. The effect of the method of preparation on the electrical properties is discussed, and current theories of the dependence of I per V characteristics on the distribution of impurities are questioned.

546.289:548.55 1682
Growth of Germanium Single Crystals containing *p-n* Junctions—G. K. Teal, M. Sparks, and E. Buehler. (*Phys. Rev.*, vol. 81, p. 637; February 15, 1951.) A note describing methods of producing Ge single crystals by progressive pulling from the melt. One type thus produced is a crystal in which the magnitude and type of conductivity in the direction of crystal growth is controlled by addition of an impurity such as Ga (acceptor) or Sb (donor), to the melt from which the crystal is being grown. In this way *p-n* junctions have been formed in Ge single crystals, which are exceptional in their agreement with theory and in their electrical properties. See also 1669 above (Goucher et al.)

548.0:537.228.1:539.32 1683
Elasticity of Piezoelectric and Ferroelectric Crystals—F. Jona. (*Helv. Phys. Acta*, vol. 23, pp. 795–844; December 10, 1950. In German, with English summary.) The diffraction of a beam of monochromatic light by a crystal, subjected to ultrasonic excitation, was used to determine the elastic behavior of various crystals. The elastic constants for Rochelle salt and $NaClO_3$ measured between −50° and +30°C. are in agreement with previous results. A theoretical argument shows that the values obtained for ferroelectric crystals by this method refer to constant electric field, and a marked anomaly for KH_2PO_4 at the Curie point is confirmed. Differences in intensity of the observed diffraction patterns are discussed.

621.3.042.2:538.144 1684
H. F. Magnetization of Ferromagnetic Laminae—O. I. Butler and H. R. Chablani. (*Wireless Eng.*, vol. 28, pp. 92–97; March, 1951.) "An attempt is made to improve the accuracy and consistency of calculations based on the classical theory of ac magnetization of ferromagnetic laminas by the simple, but logical, expedient of using a value of permeability, which differs from the ratio of the peak values of B and H. The chosen value of permeability is, to some extent, dependent upon the shape of the B/H curve, which justifies its use in the case of high-frequency magnetization of the laminas.

It is found that calculated results of the power loss and apparent permeability of silicon-steel samples give fairly close agreement with experiment. A similar accuracy is obtained by a more rigorous and laborious solution which replaces the B/H curve by a Legendre polynomial series. It appears that there is a definite limitation in the accuracy of calculations based on the dc characteristics of the material, and the inherent supposition that the material is homogeneous."

621.314.632:546.289:537.533.9 1685
Electron-Bombardment-Induced Conductivity in Germanium Point-Contact Rectifiers—A. R. Moore and F. Herman. (*Phys. Rev.*, vol. 81, pp. 472–473; February 1, 1951.) A report of preliminary experiments. The surface of the Ge is scanned as in television, using an electron-beam voltage of several kilovolts, and the signal generated in the circuit of the point contact is used to provide modulation in a kinescope or oscilloscope indicating tube. Displays obtained are shown and discussed. The records provide evidence regarding the charge-carrying processes in Ge.

621.314.634 1686
Origin of a Time-Lag Effect in Selenium Rectifiers—K. Lehovec. (*Nature* (London), vol. 167, pp. 522–523; March 31, 1951.) To decide whether resistance drift after application of voltage is caused by ionic migration or by trapping and thermal release of electrons, a Se rectifier with a transparent Cd electrode was prepared, and the influence of light on the effect was studied. It is concluded that the recovery, after pulsing with a small voltage in the blocking direction, is an electronic process.

621.315.613.1:538.214 1687
Magnetic Susceptibility and Anisotropy of Mica—J. T. Kendall and D. Yeo. (*Proc. Phys. Soc.* (London), vol. 64, pp. 135–142; February 1, 1951.) Measurements on natural muscovite and synthetic fluorphlogopite show that the mean susceptibility varies approximately linearly with the total iron content, and that the paramagnetic anisotropy is proportional to the ferrous-iron content. At low field strengths, ferromagnetic impurities cause anomalous results. These impurities account for the apparent paramagnetism of synthetic mica.

621.315.616.96:621.317.333.6 1688
The Deterioration and Breakdown of Dielectrics Resulting from Internal Discharges—J. H. Mason. (*Proc. IEE* (London), Part I, vol. 98, pp. 44–59; January, 1951.) The deterioration of polythene and other materials, when subjected to internal discharges, was investigated under controlled conditions. The discharge-inception voltage in voids of different dimensions was determined and the progressive physical deterioration under the action of repeated discharges was observed; typical results are reproduced in tables and photographs. Under equivalent conditions, polytetrafluoroethylene and perspex are less resistant to discharges than polythene.

621.396.822:539.23:621.315.616.9 1689
Random Noise in Dielectric Materials—J. H. Mason. (*Jour. Appl. Phys.*, vol. 22, pp. 235–236; February, 1951.) Electrophotographs show that individual discharges between dielectric surfaces involve only a small area, although continued application of stress may affect many separate areas until the whole area is discharged. The variation in magnitude of the noise observed with different dielectrics and varying humidity is explained by variation in surface and volume conductivity, which control the area which discharges as a single unit, and the time constant of charge leakage.

621.775.7 1690
Metallkeramik. [Book Review]—F. Skaupy. Publishers: Verlag Chemie, Weinheim/Bergstr., Germany, 4th edn 1950, 267 pp., 19 DM. (*Metal Ind.* (London), vol. 78, p. 149; February 23, 1951.) A completely revised and enlarged edition of the first book to be published on powder metallurgy. "It is a most excellent review of the subject."

MATHEMATICS

517.514 1691
Notes on Laplacian Stationary Aleatory Functions—A. Blanc-Lapierre. (*Compt. Rend. Acad. Sci.* (Paris), vol. 232, pp. 934–936; March 5, 1951.) Properties of these functions are enumerated, and a theorem relating to elliptically polarized light is deduced.

517.514:517.942.9 1692
Harmonic Analysis of Laplacian Stationary Aleatory Functions—A. Blanc-Lapierre. (*Compt. Rend. Acad. Sci.* (Paris), vol. 232, pp. 1070–1072; March 12, 1951.) The properties of a monochromatic or quasimonochromatic component of a Laplacian stationary aleatory function are defined, using Poincaré's representation. The method is useful for describing the statistical properties of radiation of arbitrary polarization.

517.93 1693
A Differential Equation occurring in Physics—N. Minorsky. (*Compt. Rend. Acad. Sci.* (Paris), vol. 232, pp. 1060–1062; March 12, 1951.) Stable periodic solutions are sought for the equation $\ddot{x}+b\dot{x}+x+(a-ex^2)\ x\ \cos\ 2t+ex^3=0$, using the calculus of perturbations. The general case and 14 particular cases are considered.

519.2:621.3.015.2 1694
On the First-Passage-Time Probability Problem—A. J. F. Siegert. (*Phys. Rev.*, vol. 81, pp. 617–623; February 15, 1951.) An exact solution is derived for the first-passage-time probability of a stationary one-dimensional Markoffian random function, from an integral equation. A recursion formula for the moments is given for the case where the conditional probability density, describing the random function, satisfies a Fokker-Planck equation. Various known solutions for special applications (noise, Brownian motion) are special cases of the general solution. The Wiener-Rice series, for the recurrence-time probability density, is derived from a generalization of Schrödinger's integral equation, for the case of a two-dimensional Markoffian random function. See also 1127 of June (Stumpers).

681.142 1695
A Process for the Step-by-Step Integration of Differential Equations in an Automatic Digital Computing Machine—S. Gill. (*Proc. Camb. Phil. Soc.*, vol. 47, part 1, pp. 96–108; January, 1951.) "It is advantageous, in automatic computers, to employ methods of integration which do not require preceding function values to be known. From a general theory given by Kutta, one such process is chosen, giving fourth-order accuracy and requiring the minimum number of storage registers. It is developed into a form which gives the highest attainable accuracy and can be carried out by comparatively few instructions. The errors are studied and a simple example is given."

681.142 1696
A Simple Analogue Computor for Fourier Analysis and Synthesis—J. H. Bowen and T. E. Burnup. (*Electronic Eng.* (London), vol. 23, pp. 67–69; February, 1951.) The computer was designed to solve integrals encountered in the study of the responses of electromechanical systems to applied impulses. It is simpler, though less accurate, than previously described computers, and incorporates only components that are commercially available.

681.142 1697
Binary Counter with Additive or Subtractive Operation controlled by Pulses—A. Peuteman. (*Compt. Rend. Acad. Sci.* (Paris), vol. 232, pp. 1082–1084; March 12, 1951.)

681.142:061.3 1698
Calculating Machines and Human Thought—D. M. MacKay. (*Nature* (London), vol. 167, pp. 432–434; March 17, 1951.) A short report on a colloquium held in Paris. The main sections were: the design of computers, their application to mathematics, and analogies between the functions of computing elements and those of the brain.

681.142:512.831 1699
An Improved Electrical Network for Determining the Eigenvalues and Eigenvectors of a Real Symmetric Matrix—A. Many. (*Rev. Sci. Instr.*, vol. 21, pp. 972–974; December, 1950.) Description of an enlarged and improved form of the computer previously de-

scribed [131 of 1949 (Many & Meiboom)]. The time required for solution has been shortened, and the accuracy made ten times greater. A matrix of the tenth order can be solved in 4 hours.

517.43 1700

Operational Calculus—based on the Two-Sided Laplace Integral [Book Review]—B. van der Pol and H. Bremmer. Publishers: Cambridge University Press, London, Eng., 415 pp., 55s. (*Brit. Jour. Appl. Phys.*, vol. 2, pp. 57–58; February, 1951.) "Mainly concerned with [the two-sided Laplace and the Fourier] integrals . . . This book is likely to become, and should become, a classic. There is no doubt that those interested in applying mathematics will find it of the greatest value and interest."

517.51 (083.5) 1701

Tables of the Function sin ϕ/ϕ and of its First Eleven Derivatives [Book Review]—Staff of the Computation Laboratory, Harvard University. Publishers: Harvard University Press, Cambridge, Mass., and Oxford University Press, London, Eng., 1949, 241 pp., 63s. (*Nature* (London), vol. 167, p. 375; March 10, 1951.)

MEASUREMENTS AND TEST GEAR

621.314.58.001.4 1702

Test Procedures for Checking Performance of Vibrator Power Supplies—M. S. Roth. (*FM-TV*, vol. 11, pp. 24–25; January, 1951.) Various possible causes of faulty operation or failure of vibrator units are discussed, and oscilloscope tests for voltage and current waveforms are described. Short vibrator life is generally due to faults external to the vibrator itself, so that the checking of the components in the associated circuit is of primary importance in the case of vibrator failure.

621.317.029.5/.6 1703

Second Conference on High-Frequency Measurements—(*Tech. Bull. Nat. Bur. Stand.*, vol. 35, pp. 39–43; March, 1951.) New Developments from laboratories throughout The United States were reported at the conference held in Washington, D. C., January, 1951. Brief notes are given of the subject matter of the papers presented, which were grouped under the headings: frequency and time; impedance; power and attenuation; transmission and reception.

621.317.3:621.396.611.21 1704

Measurement of the Electrical Behaviour of Piezoelectric Resonators—C. F. Floyd and R. L. Corke. (*Proc. IEE* (London), Part III, vol. 98, pp. 123–132; March, 1951.) The equivalent circuit of a crystal resonator is determined from measurements of, (a) the series- and parallel-resonance frequencies, (b) the insertion loss as a series element in a suitable low-impedance transmission test-set, (c) the capacitance at 1 kc. A new type of holder is used in which specimens can be accurately clamped at any desired point on their major surfaces. The results of measurements obtained on various types of quartz, ADP and EDT resonators, are tabulated and the effects of off-node mounting are illustrated.

621.317.3.087.6:621.396.67.012 1705

Polar Diagram Plotter—(*Wireless World*, vol. 57, p. S4; February, 1951.) The polar diagram of a scaled-down (1:10) model of an antenna array is obtained in about 15 seconds. The transmitter, operating on a 10-times higher frequency, is fixed while the receiving antenna array is rotated about its mast, at the same angular velocity as that of a circular chart about its center. Simultaneously, the pen of a recorder is deflected radially across the chart, the amount of deflection being controlled by the rectified antenna voltage.

621.317.333.6:621.315.616.96 1706

The Deterioration and Breakdown of Dielectrics Resulting from Internal Discharges—Mason. (*See* 1688.)

621.317.335.3.029.64†:546.217 1707

Dielectric Constant and Refractive Index of Air and its Principal Constituents at 24,000 Mc/s—L. Essen and K. D. Froome. (*Nature* (London), vol. 167, pp. 512–513; March 31, 1951.) A high-accuracy method, developed at the National Physical Laboratory, consists basically in observing the resonance frequency of a cavity (a) when filled with the gas, and (b) when evacuated. All frequency measurements are referred directly to quartz-standard oscillators, and a high-Q cavity in a bridge circuit is used to obtain the necessary sensitivity. The value of the dielectric constant found for dry CO_2-free air at NTP is 1.0005764 $\pm 2\times 10^{-7}$. Results are compared with those obtained from other recent researches.

621.317.34:621.392.26† 1708

Power Adjustment for Plane Waves in Waveguides—Mataré. (*See* 1565.)

621.317.341:621.315.212 1709

The Use of a Piston Attenuator for Cable Testing in the Frequency Range 1-30 Mc/s—E. C. H. Seaman, D. A. Crow, and C. G. Chadburn. (*P.O. Elect. Eng. Jour.*, vol. 43, part 4, pp. 192–197; January, 1951.) The instrument described has been designed for insertion-loss measurements on the more recently developed types of hf coaxial cables. Crystal-controlled spot frequencies in the range 1 to 30 mc are available from a portable signal generator in which piston attenuators are incorporated. These permit accurate measurement of the ratio of the input and output voltages of a cable.

621.317.342:621.392 1710

The Calculation of Phase Constant for Small Differences of Open and Closed Impedance—P. R. Bray. (*P.O. Elec. Eng. Jour.*, vol. 43, part 4, pp. 200–201; January, 1951.) A note on the use of a bridge for the measurement of the phase constant of a transmission line. The derivation of the phase constant is simplified by using, directly, the difference between the bridge settings for the open-circuit and closed-circuit conditions.

621.317.41 1711

An Oscillation Type Magnetometer—J. H. E. Griffiths and J. R. MacDonald. (*Jour. Sci. Instr.*, vol. 28, pp. 56–58; February 1951.) Description of a method developed for measuring the saturation magnetization of small Ni disks of thicknesses down to 0.1 μ. Measurements were also made on Ni and supermalloy disks of the thickness 0.11 mm. The sample under test is made to oscillate in a uniform field parallel to the plane of the disk, the axis of rotation being a diameter normal to the field.

621.317.727.029.63 1712

Radio-Frequency Micropotentiometer—(*Tech. Bull. Nat. Bur. Stand.*, vol. 35, pp. 33–34; March, 1951.) Brief account of a coaxial type of device developed by M. C. Selby and making available, without the use of attenuators, accurate voltages of 1 to 10^5 μv at impedances of the order of milliohms and at frequencies up to 1 kmc.

621.317.729.088 1713

Factors Limiting the Accuracy of the Electrolytic Plotting Tanks—P. A. Einstein. (*Brit. Jour. Appl. Phys.*, vol. 2, pp. 49–55; February, 1951.)

621.317.73.029.64 1714

U.H.F. Measurements with the Type 874-LB Slotted Line—R. A. Soderman and W. M. Hague. (*Gen. Rad. Exper.*, vol. 25, pp. 1–11; November, 1950.) The equipment comprises oscillator, coaxial slotted line with traveling electrostatic probe, and crystal or receiver detector. The frequency range is 300 to 4,000 mc. The use of the line for impedance measurements is described, and sources of error are discussed.

621.317.733 1715

Automatic A.C. Bridges—J. F. Graham (*Electronics*, vol. 24, pp. 110–116; February, 1951.) The design of bridge and detector circuits for production-line measurements of inductance, capacitance, and resistance is discussed. These depend on a phase discriminator, which gives zero output for input voltages 90 degrees out of phase. In one system, the standard-variable circuit elements are adjusted by servo-motors, until a balance is obtained. Graphical methods are used for determining the phase and amplitude of the bridge unbalance voltage.

621.317.733 1716

New Version of Schering Bridge—J. H. Jupe. (*Electronics*, Vol. 24, No. 2, pp. 214, 218; February, 1951.) Suitable for voltages up to 200 kv. The detector unit uses a 3-stage RC-coupled amplifier with a thermionic rectifier and dc moving-coil milliammeter for indication of balance.

621.317.733.011.4.029.53/.55 1717

A Capacitance Bridge for High Frequencies—J. S. Mendousse, P. D. Goodman, and W. G. Cady. (*Rev. Sci. Instr.*, vol. 21, pp. 1002–1009; December, 1950.) A bridge for measurements in the megacycle range is described. The unbalance voltage is rectified by a Ge-crystal rectifier and measured as a small direct voltage. For examining the vibration modes and measuring the Q of piezoelectric crystals, the variable capacitor is set at a value equal to the parallel capacitance of the crystal, and the output voltage is recorded graphically, while the frequency is slowly varied over the resonance range. Typical records are shown illustrating the performance of crystals at various frequencies and under different mechanical loads. Theory of the bridge is given and sources of error are discussed.

621.317.755:621.3.015.3 1718

An Oscilloscope for the Observation of Long-Duration Transients—A. E. Ferguson. (*Jour. Sci. Instr.*, vol. 28, pp. 52–56; February, 1951.) A cro with a very-long-persistence screen for the observation of transients, of duration 30 seconds or more, is described. The linear timebase provides either repetitive or single sweeps.

621.317.755:621.317.772 1719

Aids to C.R.O. Display of Phase Angle—L. Fleming. (*Electronics*, vol. 24, pp. 226, 234; February, 1951.) Some simple methods of applying passive circuits to phase-angle indication on a cro, for use at audio frequencies.

621.317.757:534.41 1720

Analysis of a Spectrum of Very Low Frequencies by means of Magnetic Tone-Frequency Equipment—K. H. R. Weber. (*Funk. u. Ton*, vol. 4, pp. 619–627; December, 1950.) The principle of the method is the multiplication of the original frequencies so that an analyzer with small bandwidth may be used. The signal is recorded magnetically at a low tape velocity, 2.57 cm, and reproduced at the normal velocity of 77 cm, so that frequencies are multiplied by 30. The complete apparatus is described, including a test oscillator generating very low frequencies (1 to 300 cps) by electromechanical means.

621.317.761 1721

Production-Line Frequency Measurement—G. J. Kent. (*Electronics*, vol. 24, pp. 97–99; February, 1951.) A description of equipment for the accurate measurement of the frequencies of crystals used in monitors and oscillators. Frequencies up to 10 mc can be measured rapidly, by relatively unskilled operators, to within 1 or 2 cps. The frequency standard is a 100-kc oscillator, the 50th harmonic of which is kept at zero beat with the 5-mc signal from WWV. The first digit in the unknown frequency is obtained from the calibration of a communications receiver, the remaining digits being evaluated in turn, using methods of bracketing between known harmonics or subharmonics of the master oscillator, until the last two digits can be obtained by comparison with a calibrated af oscillator, using Lissajous figures.

621.317.761.426 1722

Apparatus for Accurate Measurement of Frequency—W. S. Wood. (*Engineering* (London), vol. 171, p. 216; February 23, 1951.) A description of portable apparatus for the measurement of frequencies near 50 cps or multiples thereof, with an accuracy to within ±0.1 per cent. A standard tuning fork of variable known frequency is used to discharge a capacitor periodically, and the resulting sawtooth voltage is applied to the x plates of a cro. The voltage whose frequency is required is applied to the y plates, and the frequency difference can be obtained by observing the speed of the trace drift, or the fork frequency can be varied to produce a stationary trace.

621.317.761.029.64 1723

A Frequency Meter for Microwave Spectroscopy—J. D. Rogers, H. L. Cox, and P. G. Braunschweiger. (*Rev. Sci. Instr.*, vol. 21, pp. 1014–1015; December, 1950.) Description of the operation of equipment using known absorption lines as frequency standards. By mixing signals of unknown microwave frequency f_x, a standard microwave frequency f_0, and known rf signals nf_c from a crystal-controlled harmonic generator, beats of frequency f_r can be observed by means of a calibrated receiver. The unknown frequency f_x can then be determined from the relation

$$f_r = \pm m(f_x - f_0) \pm nf_c.$$

Measurements by this method of the frequencies of methanol absorption lines, using NH_3 absorption lines as frequency standards, show that frequencies of the order of 25 kmc can be measured to within ±0.03 mc.

621.317.772 1724

A Simple Phase Measuring Circuit—R. A. Seymour. (*P.O. Elec. Eng. Jour.*, vol. 43, part 4, pp. 198–199; January, 1951.) A description and analysis of an addition circuit used for measuring the phase angle between two sinusoidal signals, having frequencies in the range 50 cps to 200 kc. Phase angles between 0° and 60° can be measured to within ±1°. but larger errors occur as the angle approaches 90°.

621.396.615.015.7† 1725

An Incremental-Delay Pulse Generator—G. F. Montgomery. (*Electronics*, vol. 24, pp. 218, 226; February, 1951.) The equipment uses a 100-kc oscillator with a chain of four frequency dividers of the ring-counter type, and provides three 25-cps pedestal outputs. Two of these are variable in phase, in 0.1-ms steps over the 40-ms cycle. Timing marks at intervals of 0.1, 1, 10, and 40 ms are generated, and also a 50-cps output for operating a clock.

621.396.615.11.001.4:621.396.619.13 1726

Complex-Tone Generator for Deviation Tests—F. A. Bramley. (*Electronics*, vol. 24, pp. 184, 196; February, 1951.) Frequency-deviation and modulation-deviation measurements are often inconsistent, when voice-testing is used. A two-tube battery-operated generator is described for producing a standard complex tone for such tests.

621.396.62.001.4 1727

Performance Tests on Radio Receivers—N. S. Smith. (*Telecommun. Jour. Aust.*, vol. 7, pp. 155–168; February, 1949.) An outline of tests for sensitivity, selectivity, gain, distortion, frequency drift, calibration, etc., and description of suitable equipment.

621.396.645.029.4/.5].001.4 1728

A Null Method of Measuring the Gain and Phase Shift of Comparatively Low Frequency Amplifiers—T. Baldwin and J. H. Littlewood. (*Electronics Eng.* (London), vol. 23, pp. 65–66; February, 1951.) Description of a simple practical method requiring only an accurate attenuator calibrated in 0.1-db steps, a cro, a variable decade capacitor, a known resistor, and a vfo with adequate output.

621.396.822(083.74) 1729

Noise Figure Standards—(*Tech. Bull. Nat. Bur. Stand.*, vol. 35, pp. 27–28; February, 1951.) The National Bureau of Standards provides a calibration service for the noise figure of linear electrical networks in the frequency range 500 kc to 30 mc. The apparatus comprises a temperature-limited noise diode, a two-terminal source network, the test network (four or two terminal), an attenuator, and a sensitive voltmeter. The method is valid for a matched or unmatched condition of input impedance.

OTHER APPLICATIONS OF RADIO AND ELECTRONICS

531.775:621.38 1730

A Precision Electronic Tachometer—S. W. Punnett and H. G. Jerrard. (*Electronic Eng.* (London), vol. 23, pp. 55–58; February, 1951.) For another account see 177 of February.

534.321.9:539.32:669 1731

Ultrasonics in Metallurgy—(*Metal Ind.* (London), vol. 78, p. 146; February 23, 1951.) Brief account, taken from an address by Sir Ben Lockspeiser to the Scottish branch of the Institute of Physics, of ultrasonic methods of determining the elastic constants of metals. The methods were developed by G. Bradfield at the National Physical Laboratory and are particularly useful when the dimensions of the sample are small. See also *Engineer* (London), vol. 191, p. 318; March 9, 1951.

534.321.9:621.791.3 1732

Ultrasonic Soldering Irons—B. E. Noltingk and E. A. Neppiras. (*Jour. Sci. Instr.*, vol. 28, pp. 50–52; February, 1951.) Two types are described, one for normal use on sheet aluminium or light alloy, while in the other the vibrations are communicated to a bath of solder in which small pieces of metal, or foil or wire, can be tinned by dipping. A compact electronic unit furnishes 50 w at 22 kc for energizing the magnetostriction transducer.

535.82:621.397.611.2 1733

A Flying-Spot Microscope—J. Z. Young and F. Roberts. (*Nature* (London), vol. 167, p. 231; February 10, 1951.) A cathode ray tube, providing a television raster of high brilliance and short time-constant, is placed in front of the eyepiece of a microscope. The objective produces a minute spot of light that scans the preparation under examination. The amount of light transmitted is determined by the density of the specimen and is picked up by a multiplier photocell, the output of which is used to modulate a projection-type cathode ray tube, the raster of which is locked to the scanning raster. A display some 3 feet square can be obtained with continuously variable magnification, brightness, and contrast. Resolution and quantum efficiency are considerably greater than in an ordinary microscope, and quantitative analysis becomes a possibility.

536.58:621.316.076.7 1734

Electric Controllers for Laboratory Furnaces—M. H. Roberts. (*Electronic Eng.* (London), vol. 23, pp. 51–54; February, 1951.) Detailed descriptions are given of the design and construction of two types of electronic temperature control, giving regulation to within ±1°C. in 1-kw furnaces running at 400 to 1,000°C. The instruments are based on the amplification of the out-of-balance voltage produced in a bridge circuit by an electrical temperature-measuring device, such as a resistance thermometer, forming one arm of the bridge.

In the simpler arrangement, a hot-wire vacuum switch gives two-positional control; in the other, a saturable inductor in the amplifier output circuit gives control proportional to the out-of-balance voltage, and is particularly useful in the regulation of creep-testing furnaces.

537.533.73:621.385.032.2 1735

Electron Diffraction in Valve Technique—H. A. Stahl. (*Schweiz. Arch. angew. wiss. Tech.*, vol. 16, pp. 359–369; December, 1950.) Review of the application of the electron-diffraction camera in examination of electrode surfaces, etc. in high-vacuum and gas-filled discharge tubes. Details and illustrations of diffraction patterns, for various substances, are given and discussed.

621.317.083.7 1736

Telemetering System for Radioactive Snow Gage—J. A. Doremus. (*Electronics*, vol. 24, pp. 88–91; February, 1951.) Details of unattended equipment based on the attenuation of γ radiation. Data from several sites are transmitted via fm repeater stations to a central recording station.

621.38.001:786.6 1737

Electronic Music for Four—L. A. Meacham. (*Electronics*, vol. 24, pp. 76–79; February, 1951.) Description of an "organ" with separate soprano, alto, tenor, and bass oscillators whose frequencies are adjusted by four players, operating control arms over tone-graduated quadrants. The four oscillators feed a single loudspeaker through a common amplifier.

621.384.62† 1738

Measurements on a Linear Accelerator—P. Grivet and J. Vastel. (*Compt. Rend. Acad. Sci.* (Paris), vol. 232, pp. 809–810; February 2, 1951.) Report of preliminary tests of one section of an accelerator designed for 4 to 5 mev. The hf energy is supplied by a magnetron with peak power 0.5 mw. Pulses of 2-μs duration and 100 per sec. repetition frequency give 0.4-mev electrons with a peak current of about 10 ma. The apparatus in its final form will be entirely demountable and the high-power pulses will be produced by a rotating arc.

621.385.38.001.8:621.314.653:621.791.7 1739

Load Sharer for Welder Ignitrons—G. M. Chute. (*Electronics*, vol. 24, pp. 71–73; February, 1951.) A flip-flop thyratron circuit transfers the welding load, automatically, from one pair of ignitrons to another every two seconds. This reduces the duty cycle for the ignitrons, permitting the use of a given welder for a heavier weld or a longer time than was originally specified.

621.385.833 1740

An Objective for Use in the Electron Microscopy of Ultra-thin Sections—J. Hillier. (*Jour. Appl. Phys.*, vol. 22, pp. 135–137; February, 1951.)

621.385.833 1741

The Refractive Index of Electron Optics and Its Connection with the Routhian Function—W. Glaser. (*Proc. Phys. Soc.* (London), vol. 64, pp. 114–118; February, 1951.) The derivation of the electronoptical-refractive index from Hamilton's principle is discussed. Criticisms by Ehrenberg & Siday (1455 of 1949) of the method, first introduced by the author in 1933, are shown to be invalid. To prove the general applicability of the method, the isotropic-refractive index of an axially symmetrical field is obtained.

621.385.833 1742
Optical Properties of the Independent Electrostatic [electron] Lens with Thick Central Electrode—É. Regenstreif. (*Compt. Rend. Acad. Sci.* (Paris), vol. 232, pp. 710–712; February, 1951.) Calculations presented previously (see 1455 of July) are generalized and applied to determine transgaussian-electron trajectories. The theoretical results agree well with values of converging power found experimentally by Heise and Rang.

621.385.833 1743
The Technique of the OSW [Oberspreewerk] Electron Microscope—F. Eckart. (*Elektrotechnik Berlin*, vol. 4, pp. 414–415; December 1950; and vol. 5, pp. 32–35; January, 1951.) Short discussion of basic principles and of the resolving power, as dependent on the electron-optical properties of the various lenses, together with an illustrated description of the Type-OSW2748 instrument and its stabilized power supply unit, which provides voltages of 45, 65, 85, and 100 kv and lens currents constant to within 5 parts in 10^5.

621.387.4†:621.396.823 1744
Line-Noise Interference with Particle Counting—C. G. Goss and F. M. Glass. (*Nucleonics*, vol. 8, pp. 66–69; February, 1951.) The characteristics of noise voltages encountered on laboratory lighting and power wiring are briefly discussed, with particular reference to that due to faulty fluorescent lamps. Precautions to be taken to minimize the effects of the noise are described. Particulars are given of a simple monitor which, when connected to the mains, provides audible and visual warning of the presence of excessive noise voltages.

621.387.464†:621.396.645 1745
Distributed Coincidence Circuit—Wiegand. (*See* 1599.)

621.395.625.3 1746
Magnetic Recording Systems in Product Design—Javitz. (*See* 1560.)

621.398 1747
Radio-Controlled Ship Models—S. G. Lankester and S. G. Dreier. (*Engineer* (London), vol., 191, p. 319; March 9, 1951.) A system using pulse-code tone modulation is described for the control of models used in steering investigations. In order to reduce the delay in transmission of orders which this system involves, a system using variable-frequency AM is being developed.

621.57:621.318 1748
The Magnetic-Fluid Clutch—S. F. Blunden. (*Engineer* (London), vol. 191, pp. 244–246; February 23, 1951.) Description of investigations carried out at RRDE, Malvern, on clutches of the type developed by Rabino (2585 of 1950) at the National Bureau of Standards.

681.142:536.7 1749
Predicting Phase Behavior with Digital Computers—T. J. Connolly, S. P. Frankel, and B. H. Sage. (*Elect. Eng.*, vol. 70, p. 47; January, 1951.) Summary of an AIEE Fall General Meeting paper. An account of the application of punched-card computing equipment, to the evaluation of the conventional thermodynamic properties of pure substances and mixtures.

621.38.001.8 1750
Electronics Manual for Radio Engineers [Book Review]—V. Zeluff and J. Markus. Publishers: McGraw-Hill, New York, N. Y., and London, Eng., 1949, 879 pp., 57s. (*Electronic Eng.* (London), vol. 23, pp. 75–76; February, 1951.) "This manual contains 289 of the more important articles which appeared in . . . *Electronics* during the period 1940 to 1948 . . . [It] is particularly valuable for the broad survey it provides of the most recently developed techniques in electronics; much of the information presented is not available even in the most recently published text-books in this field."

621.385.83 1751
Elektronenoptik: Vol. 1 [Book Review]—A. A. Rusterholz. Publishers: Birkhäuser, Basel, Switzerland, 1950, 249 pp., 65s. (*Electronic Eng.* (London) vol. 23, p. 76; February, 1951.) ". . . a complete restatement of the classical parts of geometrical electron optics, in plane or rotationally symmetrical fields, without space charge."

PROPAGATION OF WAVES

538.566 1752
On the Propagation of Energy in Linear Conservative Waves—Broer. (*See* 1621.)

621.396.11+535.222 1753
Proposed New Value for the Velocity of Light—Essen. (*See* 1609.)

621.396.11 1754
Ionosphere Data Deduced from Direct Tests on Radiotelegraphic Links in the Italian Army Network: Part 2—S. Silleni. (*Ann Geofis.*, vol. 3, pp. 567–578; October, 1950.) Continuation of 1753 of 1950, further analyzing the results obtained in some 60,000 calls received, involving 12 different stations. When the percentage probability of communication is plotted against the ratio of actual frequency to muf, the curve obtained nearly coincides with a Gaussian distribution curve. In order to guarantee 95 per cent probability of communication between any two points, a mean operating frequency $\not>$0.73 times the muf was found to be required. Increasing the power used from 40 w to 1 kw reduced the probability of failure of communications by a factor of 4.

621.396.11:532.517.4:551.510.5 1755
Spectrum of Atmospheric Turbulence—L. F. Richardson; E. C. S. Megaw. (*Nature* (London), vol. 167, p. 318; February 24, 1951.) Comment on 973 of May (Megaw) and author's reply.

621.396.11:551.510.535 1756
Echoes from the D and F_2 Layers on a Frequency of 21 Mc/s—E. Gherzi. (*Nature* (London), vol. 167, p. 412; March, 10, 1951.) Pulses on 21.7 mc were received at Macau between 0900 and 1200 GMT during November and December, 1950, from an unidentified distant source. The components received are thought to be the ground ray, which frequently showed very marked fading, and reflections from the *D* and F_2 layers.

621.396.11:551.594.6 1757
The Propagation of a Radio-Atmospheric—Budden. (*See* 1652.)

621.396.11:621.317.353.3† 1758
Self-interaction of Radio Waves in the Ionosphere—M. Cutolo. (*Nature* (London), vol. 167, pp. 314–315; February 24, 1951.) A description of experiments indicating that, in transmission through the ionosphere, there is a reduction in the percentage modulation of radiation, having a frequency near to the gyro-frequency. The percentage modulation decreases as the modulation frequency increases.

621.396.11:621.396.65 1759
Projected New Radio-Telephone Link from the Mainland to Tasmania: Propagation Measurements—O. M. Moriarty. (*Telecommun. Jour. Aust.*, vol. 7, pp. 281–299; October, 1949.) A detailed account of transmission tests on frequencies of 58 mc and 158 mc on a two-section link between Tasmania and the mainland, via Flinders Island. Observed values of attenuation, during the period 1947 to 1949, were in most cases within a few db of the calculated values. The measurements of received signals enabled the range of fading to be estimated. If frequency diversity is not used, an allowance of 20 db for fading will be adequate, except for a deep fade which occurs on the average less than once a fortnight. An extra allowance of 20 db would probably be sufficient to account also for this type of fading. The results, in general, indicate that a high-quality multichannel link operating on a frequency of about 60 mc, with a power of 50 w, could be established between Wilson's Promontory and Tasmania, with a repeater at Flinders Island.

621.396.11.029.45:551.594.6 1760
The Waveforms of Atmospherics and the Propagation of Very-Low-Frequency Radio Waves—P. W. A. Bowe. (*Phil. Mag.*, vol. 42, pp. 121–138; February, 1951.) The responses of narrow-band receivers to individual radio atmospherics have been studied in order to investigate the propagation over the earth's surface of waves, having frequencies between 2 and 10 kc. Atmospherics from a fixed distance are sufficiently uniform to enable the relative attenuations of different frequencies to be deduced. Frequencies below 8 kc are heavily attenuated during the daytime, particularly during sudden ionospheric disturbances, but at night such frequencies are propagated freely. The results confirm and extend those obtained recently by Gardner (1473 of July); they are consistent with the theory proposed by Budden (see 1652 above.)

621.396.11.029.51 1761
Low-Frequency Radio-Wave Propagation by the Ionosphere, with particular reference to Long-Distance Navigation—C. Williams. (*Proc. IEE* (London), Part III, vol. 98, pp. 81–99; March, 1951. Discussion, pp. 99–103.) A discussion of radio-wave propagation in the 70–300-kc band, with particular reference to the phase or time displacement of the received signal, due to the ionosphere-reflected component. Time-error curves are shown for typical propagation conditions, both for day and night, and at various frequencies. The characteristics of the errors which occur are discussed, and the magnitude of navigational errors arising therefrom are deduced, using suitable examples.

From phase and amplitude observations made with receivers in aircraft during both day and night, the relative values of the ground-reflected and the ionosphere-reflected wave components were obtained as a function of distance from the transmitters. The mean height deduced for the reflecting layer is 70 km for daytime and 90 km at night, with corresponding oblique-incidence reflection coefficients of 0.05 and 0.25.

Collation of data obtained at fixed receiving points, using Consol, Decca, and Post Office position indicator transmitters, shows that navigational accuracy is improved by making the base-line distances as great as possible.

621.396.81:551.510.535:535.361.2 1762
The Scattering of Radio Waves—W. Dieminger. (*Proc. Phys. Soc.* (London) vol. 64, pp. 142–158; February 1, 1951.) A review is given of previous work on the scattering of radio waves returned from the ionosphere. Four types of scatter are briefly described, namely E, F, G, and 2F scatter. The last of these is treated at length, and various experiments are described which lead to the conclusion that the scattering in this case takes place at the surface of the ground and not in the sporadic-*E* layer. The nature of the scattering to be expected from the ground is considered, and, it is shown, that large-scale irregularities are of major importance. The influence of scattering on short-wave communication and direction finding is discussed.

621.396.812.3:537.591 1763

The Cosmic-Ray Intensity and Radio Fade-Outs—D. W. N. Dolbear, H. Elliot, and D. I. Dawton. (*Jour. Atmos. Terr. Phys.*, vol. 1, pp. 187–188; 1951.) Analysis of cosmic-ray data, obtained during the occurrence of 35 fadeouts, indicates an average increase of intensity of about 0.3 per cent on the sunlit side of the earth during a fadeout. No such increase was found corresponding to fadeouts recorded on the other side of the earth.

RECEPTION

551.594.6 1764

Atmospheric Noise Levels at Radio Frequencies near Darwin, Australia—D. E. Yabsley. (*Aust. Jour. Sci. Res., Ser. A*, vol. 3, pp. 409–416; September, 1950.) Between August 25, 1944 and October 31, 1945, a practically continuous record of the average level of atmospheric radio noise, at a frequency of 1.93 mc, was obtained near Darwin, in north-western Australia. A few measurements were also made at a frequency of 5.9 mc. The noise-measurement program is described and the results obtained are presented graphically.

621.396.62.001.4 1765

Performance Tests on Radio Receivers—Smith. (*See* 1727.)

621.396.81:621.396.932 1766

450-Mc/s Mobile Radio Tests—A. J. Aikens and L. Y. Lacy. (*FM-TV*, vol. 11, pp. 26–27, 36; January, 1951.) Reprint. *See* 734 of April.

621.396.828.1:621.396.645.018.424† 1767

Wide-Band Amplifier for Central-Antenna Installations—Crawley. (*See* 1601.)

STATIONS AND COMMUNICATION SYSTEMS

621.39 1768

Inaugural Address [as President of the I.E.E.]—A. J. Gill. (*Proc. IEE* (London), Part I, vol. 98, pp. 1–11; January, 1951.) A review of recent developments in telecommunication engineering, in the public services of Britain.

621.39.001.11:519.21 1769

On the Relation between the Quantity of Information in the Fisher Sense and in the Wiener Sense—M. P. Schutzenberger. (*Compt. Rend. Acad. Sci.* (Paris), vol. 232, pp. 925–927; March 5, 1951.)

621.395.645 1770

A Negative-Impedance Repeater—J. L. Merrill, Jr. (*Elec. Eng.*, vol. 70, pp. 49–54; January, 1951.) An AIEE Fall General Meeting paper. The repeater described comprises a transformer, an amplifier, and associated network, arranged as a complete unit for direct insertion in a 2-wire line. It can be used to decrease transmission loss, to produce a low-loss line without reflection effects, or to neutralize a lumped positive-impedance irregularity.

621.396:355.58 1771

Defense Communication in New York City—A. A. McK. (*Electronics*, vol. 24, pp. 74–75; February, 1951.) A description, illustrated by block diagrams, of the warning facilities available in case of air attack. These include general broadcasts and special links to police and fire stations. Regular testing of the specially installed line connections is carried out. An ultrasonic-modulation system may also be used to activate publicly installed receivers with giant loudspeakers, the receivers being muted except when warnings are broadcast.

621.396 (941) 1772

The Installation of the Radio-Telegraph Network in North-West Western Australia—J. Mead. (*Telecommun. Jour. Aust.*, vol. 7, pp. 303–307; October, 1949.) Description of the equipment of a chain of stations linking Geraldton to Wyndham, a circuit of about 1,800 miles; any section can be quickly brought into service in case of a breakdown of the landline equipment. The transmitters are battery fed and crystal controlled, four frequencies being available in the range 2.5 to 10.0 mc; they give an output of 11w at 2.5 mc to a high-impedance 600-Ω antenna, which is used for both transmission and reception. Superheterodyne receivers, using vibrator hv supplies, cover the range 200 kc to 30 mc, except for a small gap at 535 kc near the IF used. Telegraphy operation is normal practice, using either headphones or loudspeaker for reception, but where no operators are available, telephony is used.

621.396.5 1773

Control-Terminal Equipment for Overseas Radio-Telephone Services—W. O. Gibberd. (*Telecommun. Jour. Aust.*, vol. 7, pp. 232–236; June, 1949.) A short description, with block diagrams, of the principal features of the Western Electric Type-C3 radio control-terminal equipment, four sets of which are installed in Sydney, on the London and San Francisco radiotelephone circuits. The use of such equipment reduces to some extent the effects of variations in the radio circuit.

621.396.5:621.396.931 1774

Mobile Radio-Telephone Services—N. S. Feltscheer. (*Telecommun. Jour. Aust.*, vol. 7, pp. 322–326; February, 1950.) Discussion of the general characteristics of and equipment for a proposed service providing communication facilities between any suitably equipped vehicle, within a prescribed area (initially within capital cities) and any telephone connected to the public network.

621.396.65:621.311 1775

Radio Communications for Power Systems—R. E. Martin. (*Engineering* (London), vol. 171, pp. 114–116; January 26, 1951.) Abridged version of a paper read at the Conférence Internationale des Grands Réseaux Électriques à Haute Tension, Paris, France, July, 1950. Radio links between control centers and repair and maintenance gangs are extremely useful. Either simplex, two-frequency simplex, or duplex operation is used by the British Electricity Authority in the 70 to 90-mc band. Fixed stations, using high vertical $\lambda/2$ dipoles, with 10-, 12- or 50-w rf output, operated from 230-v 50-cps mains, may be remotely controlled by radio link or landline from the control center. Average coverage is approximately 400 square miles. Mobile stations, installed in the maintenance trucks, use a $\lambda/4$ whip antenna and operate from the vehicle battery. AM is in general preferred to FM, because of greater ease of maintenance and superior intelligibility at long range. The speech amplifier of the mobile station may also be used for public-address work. A voice-frequency-operated calling system has been developed.

621.396.65:621.396.11 1776

Projected New Radio-Telephone Link from the Mainland to Tasmania: Propagation Measurements—Moriarty. (*See* 1759.)

621.396.65:[621.396.5+621.317.083.7 1777

Microwave Applications to Bonneville Power Administration System—R. F. Stevens and T. W. Stringfield. (*Elec. Eng.*, vol. 70, pp. 29–33; January, 1951.) Essential test of an AIEE Summer and Pacific General Meeting paper. A description is given of the proposals for the provision of general telephone facilities, telemetering of power output, supervisory control, fault location, etc. by a system which will use microwave radio links where many communication channels are required, and power-line carrier or leased telephone circuits where few channels are needed.

621.396.65.029.6 1778

Very-High-Frequency and Ultra-High-Frequency Radio Links in the Australian Post Office Communication Network—S. J. Ross. (*Jour. Inst. Eng.*, vol. 23, pp. 11–20; January to February, 1951.) A discussion of the problems involved in the introduction of vhf and uhf RT systems into the telecommunication network. The planning and engineering of such systems are considered, reference being made to site-selection problems and to the calculation of propagation-path data. A short account is included of multichannel equipment recently installed in Queensland, for communication between Brisbane and Redcliffe. This operates on a carrier frequency in the 4,600 to 4,800-mc band, pwm being used in conjunction with a time-sharing system to provide eight channels. Two further systems will soon be installed; these will provide 23 channels each, the operating frequency being 2,000 to 2,500 mc.

621.396.712:621.396.619.11/.13 1779

V.H.F. Transmitting Station at Wrotham—(*Engineer* (London), vol. 191, pp. 221–222; February 16, 1951.) A general description is given of the fm and am transmitters operating, respectively, at powers of 25 kw and 18 kw and on carrier frequencies of 91.4 mc and 93.8 mc, of the housing, control, and monitoring of the equipment and of the mast, feeder, antenna system, and power supplies.

621.396.82:621.396.41 1780

Interference in Multi-Channel Circuits—L. Lewin. (*Wireless Eng.*, vol. 28, p. 98; March, 1951.) Correction to paper abstracted in 986 of May.

621.396.93 1781

Radio in the Jungle—"Pronto." (*Wireless World*, vol. 57, pp. 73–74; February, 1951.) The difficulties of maintaining military communications in Malaya due to the dense jungle, high humidity, and high level of atmospherics at night, are briefly reviewed. The ideal equipment should operate on the sky wave at crystal-controlled frequencies between 4 and 10 mc; it should be light yet robust, simple to operate, have facilities for both speech and Morse, and must be fully tropicalized.

621.396.931 1782

Design of Mobile Two-Way Radio Communication Equipment at 152–174 Mcs—R. A. Beers, W. A. Harris, and A. D. Zappacosta. (*Broadcast News*, pp. 56–65; January to February, 1951.) More detailed description of equipment noted in 989 of May.

SUBSIDIARY APPARATUS

621.314.58 1783

Methods of increasing the Power Rating of Vibratory Convertors—K. H. Dixey and C. V. Wilman. (*Proc. IEE* (London), Part III, vol. 98, pp. 105–111; March, 1951.) The power limitations of converters are discussed, and two methods by which the difficulties may be overcome are described. Both make use of special circuits designed to reduce the current carried by the contacts at the moment of breaking. Reference is also made to methods of dealing with certain vibrator defects and with transformer surges. Examples are given of practical applications of the circuits described.

621.314.58.001.4 1784

Test Procedures for Checking Performance of Vibrator Power Supplies—Roth. (*See* 1702.)

621.396.68:621.317.755 1785

E. H. T. from an R. F. Oscillator—C. J. Dickinson. (*Wireless World*, vol. 57, pp. 70–72; February, 1951.) Construction details are given of a simple unit providing a current of 1 ma at 2 kv, or a greater current at a lower voltage, from an oscillator operated at a frequency in the range 20 to 100 kc.

TELEVISION AND PHOTOTELEGRAPHY

621.397.5:535.623/.624 1786
Comparative Analysis of Color TV Systems—A. V. Loughren and C. J. Hirsch. (*Electronics*, vol. 24, pp. 92–96; February, 1951.) General analysis, with particular reference to band sharing and the use of the "mixed-highs" principle in the dot-sequential system. See also 250 of February (Bedford) and 466 of March (Dome).

621.397.5:535.623 1787
Color Television Systems—F. Shunaman. (*Radio and Electronics*, vol. 22, no. 4, Annual Television Number, pp. 20–22, 32; January, 1951.) Description, with illustrations in color, of the CBS field-sequential, the CTI line-sequential and the RCA dot-sequential systems. See also 249 of February.

621.397.5:535.623 1788
Color Fundamentals for TV Engineers—Fink. (*See* 1610.)

621.397.5:535.623 1789
Progress in Dot-Sequential Color TV—D. G. F. (*Electronics*, vol. 24, pp. 80–81; February, 1951.) The newest tricolor tube has higher resolution (600,000 phosphor dots in a screen of diameter 13½ inches), and red and blue phosphors of truer colors and greater brightness. Improved circuits reduce the visibility of the dot structure, and eliminate spurious patterns due to beats between the image structure and the dot structure.

621.397.6 1790
A New Video Distribution System—E. D. Hilburn. (*Tele-Tech*, vol. 9, pp. 28–30, 75; December, 1950.) The "bridged-T" and "parallel-amplifier" arrangements are reviewed, and it is shown how their disadvantages are removed in the system described, where a single video power amplifier is used to feed a parallel pad network, each unit of which is connected to a terminated monitoring or switching unit.

621.397.6:621.396.67 1791
Television Totem Pole—Kear and Hanson. (*See* 1568.)

621.397.611.2 1792
Image Tubes and Techniques in Television Film Camera Chains—R. L. Garman and R. W. Lee. (*Jour. Soc. Mot. Pic. & Telev. Eng.*, vol. 56, pp. 52–64; January, 1951.) A review of techniques used for televising motion-picture films. The flying-spot scanning technique, and other methods using various types of storage and nonstorage pickup devices, are discussed and compared in relation to such factors as signal noise ratio, spurious signals, spectral response, and transfer characteristic.

621.397.62 1793
Matching of the Frame Output Stage to the Deflection Coils in Television Receivers: Part 1—Low-Impedance Deflection Coils—P. D. van der Knaap and J. Jager. (*Philips Tech. Commun.* (Australia), no. 7, pp. 13–19; 1950.) Design calculations are given for frame output stages, using low-impedance deflection coils and a matching transformer between these coils and the output tube. Calculations are based on the assumption that the sawtooth current through the coils is linear, and an output transformer of reasonable dimensions is then chosen. Finally, the current which the preceding stage must be able to supply is computed.

621.397.62.004.64/.67 1794
Television Trouble-Shooting, Alignment Equipment and Procedures—J. R. Meagher. (*Radiotronics*, vol. 16, pp. 4–26; January, 1951.) Reprinted from RCA Service Co. publication. Common faults are described and step-by-step methods of discovering their causes and eliminating them are outlined. Servicing equipment and methods for re-alignment of television receivers are dealt with similarly.

621.397.621.2.002.2 1795
Characteristics of All-Glass Television Picture Bulbs—J. L. Sheldon. (*Jour. Soc. Mot. Pic. & Telev. Eng.*, vol. 56, pp. 65–74; January, 1951.) Methods of manufacturing the glass envelopes of television tubes are described. The relevant properties of the glass used are discussed, with particular reference to a new glass (Corning Code 9010). The trends in the size and shape of tubes over the past few years are examined; the rectangular type will probably continue to be largely used. Methods are mentioned which can be adopted in the design of the envelope, so as to reduce the glare arising from reflection at the tube face of light from external objects.

TRANSMISSION

621.396.619.14:621.385.5 1796
A Beam-Deflection Phase-Modulator Valve—Hopkins. (*See* 1805.)

TUBES AND THERMIONICS

621.383.2 1797
Photoelectric Effect in Cs-O, Cs-S, Cs-Se and Cs-Te Photocathodes—W. Baumgartner and N. Schätti. (*Helv. Phys. Acta*, vol. 23, pp. 869–873; December 10, 1950. In German.) Experimental determinations of spectral sensitivity.

621.385.002.2:668.393:679.52 1798
Cements used in Radio-Valve Construction—C. Biguenet. (*Le Vide*, vol. 3, pp. 451–453; May, 1948.) Nitrocellulose-base cements are used for support (e.g. of cathode carbonates) or for protection of tube elements during assembly. The physico-chemical properties of these cements are discussed briefly.

621.385.2/.5 1799
Principles of the Electrical Rating of High-Vacuum Power Tubes—E. E. Spitzer. (Proc. I.R.E., vol. 39, pp. 60–69; January, 1951.) A rational system of ratings for power tubes is described which enables the rating to be calculated from the results obtained in operating and life tests of the tube, when used as a rf power amplifier, and as a class-C oscillator. Rating factors are tabulated and a system for reducing the ratings of tubes at hf is outlined.

621.385.2.011 1800
Theory of the Parallel Plane Diode—A. H. Taub and N. Wax. (*Jour. Appl. Phys.*, vol. 22, pp. 108; January, 1951.) Correction to paper noted in 777 of April.

621.385.2.032.216 1801
Some Characteristics of Diodes with Oxide-Coated Cathodes—W. R. Ferris. (*RCA Rev.*, vol. 11, p. 568; December, 1950.) Corrections to paper abstracted in 2099 of 1949.

621.385.3:546.289 1802
Physics and Technology of Transistors—E. H. Hungermann. (*Elecktron Wiss. Tech.*, vol. 4, pp. 357–367; October and November, 1950.) Description of the mechanism of transistor action and review of the development in 1949, particularly in The United States, of the transistor amplifier. Illustrations and details of basic circuits are given.

621.385.3:546.289 1803
A High-Performance Transistor with Wide Spacing between Contacts—B. N. Slade. (*RCA Rev.*, vol. 11, pp. 517–526; December, 1950.) Contact spacings between 0.010 and 0.020 inches give transistors with power gains of 20 to 30 db and current gains up to 25, at the same time reducing the average value of equivalent base resistance. Transit-time effects, however, cause the current gain to decrease more rapidly with frequency in transistors with wide-spaced than with narrow-spaced contact points. This disadvantage can be overcome by activating at wide spacing and operating with narrow-spaced contacts. High current gains and good frequency response can then be obtained if the Ge crystal material is properly selected.

621.385.3.012.6+621.385.5.012.6 1804
Valve Input Conductance at V.H.F.—Cathode Circuit Feedback—E. E. Zepler. (*Wireless Eng.*, vol. 28, pp. 51–53; February, 1951.) Mutal inductance between the cathode lead and grid and anode leads of a triode should be taken into account when calculating input conductance, due to effective cathode-lead inductance. When this is done, it is possible to neutralize input conductance by running leads correctly. For pentodes, positive input conductance, due to grid cathode capacitance, can be neutralized by negative conductance due to grid-screen-grid capacitance, without inserting an additional inductance in the screen-grid lead, especially in a single-ended tube. See also 4355 of 1938 (Strutt and van der Ziel).

621.385.5:621.396.619.14 1805
A Beam-Deflection Phase-Modulator Valve—E. G. Hopkins. (*AWA Tech. Rev.*, vol. 9, pp. 53–66; January, 1951.) The intensities of two electron beams of rectangular section are independently modulated by voltages derived from two carriers in phase quadrature. The beams are deflected by a common deflection system, which swings them across an electrode in which a series of apertures having the profile of a rectified sine wave are cut. Each aperture has an anode immediately behind it; alternate anodes are linked together and feed a balanced output circuit. A sine-wave modulation is thus impressed on the beams, and since the equivalent distance between their positions at any instant, expressed in terms of the sine-wave profile, is $\pi/2$ radians, the resulting output is a phase-modulated wave. Investigations with an experimental tube are described and improvements in construction are suggested.

621.385.832 1806
Correction of Deflection Defocusing in Cathode-Ray Tubes—J. E. Rosenthal. (Proc. I.R.E., vol. 39, pp. 10–15; January, 1951.) Formulas are derived which give the shape of the deflection plates for minimum spot distortion in tubes using electrostatic deflection.

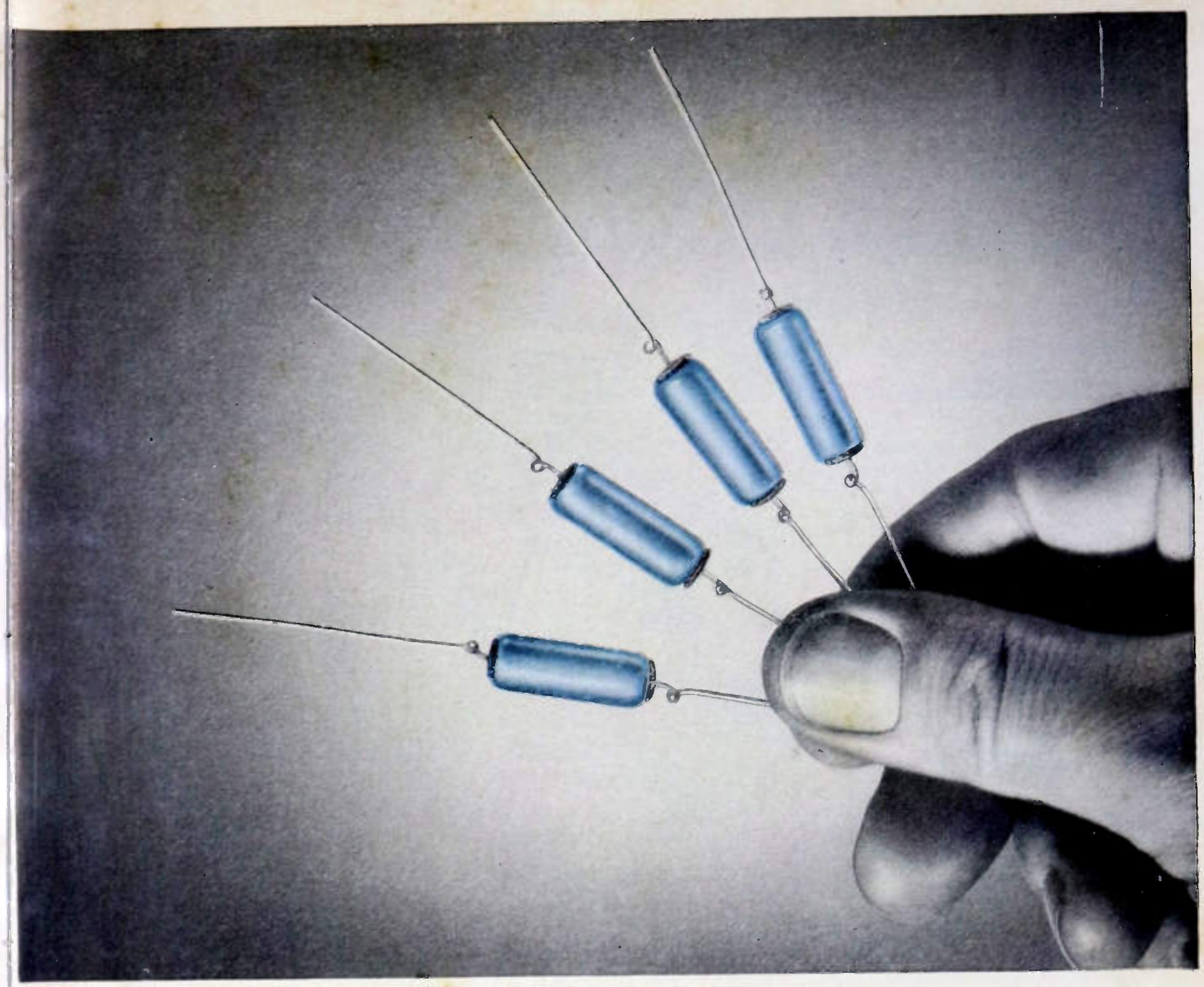
621.396.615.14.029.66 1807
Millimeter Waves—Pierce. (*See* 1598.)

621.396.822 1808
Valve and Circuit Noise. Radio Research Special Report No. 20 [Book Review]—Publishers: H. M. Stationery Office, London, Eng., 19 pp., 9d. (*Wireless Eng.*, vol. 28, p. 97; March, 1951.) "This report of the Department of Scientific and Industrial Research provides a survey of existing knowledge and outstanding problems. . . . A bibliography is included."

MISCELLANEOUS

621.3(083.74/.75) 1809
Good Engineering Practice—(*Jour. Brit. IRE*, vol. 11, pp. 25–32; January, 1951.) A review of radio and electronic standards and specifications, prepared by the Technical Committee of the British Institution of Radio Engineers, and based on a report by F. G. Diver and H. E. Drew. Specific references to numerous relevant published standards are listed in appendixes.

621.396 1810
Radio Research, 1933–1948. Book Notice—Publishers: H. M. Stationery Office, London, Eng., 1950, 2s. (*Govt. Publ.* p. 21; December, 1950.) Report of the Radio Research Board for the period October 1, 1933 to December 31, 1948, with a survey of the investigations carried out during 1934 to 1947 and report of the Director for 1948.



SECTION MEETINGS

Akron

"The Future of the Radio Engineer," by P. L. Hoover, Faculty, Case Institute of Technology; May 15, 1951.

Atlanta

"Results of the Investigation on UHF-TV Transmission and Reception in the Bridgeport, Connecticut Area," by Raymond Guy, National Broadcasting Company; May 8, 1951.

Baltimore

"Control Problems in Nuclear Power Plants," by M. A. Schultz, Manager, Instrumentation and Control, Westinghouse Atomic Power Division; June 13, 1951.

Boston

"Trends in the Field of Magnetic Recording," by S. J. Begun, Brush Development Company; May 3, 1951.

"The Pendulum of the Ammonia Clock," by E. W. Fletcher, Cruft Laboratory, Harvard University and Election of Officers; May 17, 1951.

"Subharmonic Generation in Loudspeakers," by W. J. Cunningham, Faculty, Yale University; June 7, 1951.

Cedar Rapids

"Lateral Flight Path Control," by D. L. Markusen, Minneapolis-Honeywell Regulator Company; May 23, 1951.

Cleveland

"History and Control of Industrial Radio Interference," by G. E. Sterling, Commissioner, Federal Communications Commission; January 25, 1951.

Columbus

"Activities of the Electrical Engineering Department of Ohio State University," by E. E. Dreese, Faculty, Ohio State University; April 26, 1951.

"Focusing Sound Waves with Microwave Lenses," by F. K. Harvey, Bell Telephone Laboratories; May 10, 1951.

"The Weather and Why," by R. C. McMasters, Faculty, Battelle Institute and Election of Officers; June 6, 1951.

Connecticut Valley

"Sonar Development," by C. T. Milner, Head, Underwater Communications Section, U. S. Laboratory, New London, Conn. and "High-Frequency Compensation of Amplifiers Using Feedbacks," by J. E. Shea, Student, University of Connecticut; May 17, 1951.

Annual Outing for Members and Families at Hurd State Park and Election of Officers; June 9, 1951.

Denver

"High Fidelity (?) Audio Systems," by F. J. Todd, Superintendent of Aircraft Radio and Electrical Equipment, United Air Lines; May 8, 1951.

Detroit

Address by IRE President I. S. Coggeshall, "Music and Animation in Motion Pictures," by Samuel Benavie, The Jam Handy Organization of Detroit, and "Summer Styles," courtesy of Hughes and Hatcher; May 26, 1951.

Emporium

"Applications of Reliable Tubes," by R. A. Bachhuber, Sylvania Electric Products Inc.; May 8, 1951.

Evansville-Owensboro

"Education for Action," by W. L. Everitt, Dean College of Engineering, University of Illinois and Election of Officers; May 16, 1951.

Fort Wayne

"Industrial Television," by M. C. Banca, Manager, Industrial Television, RCA; May 14, 1951.

(Continued on page 36A)