Advanced Communications

Nothing is more vital to the future of our business than the development of advanced techniques for communication. Communication systems and techniques that now exist are getting older, and obsolescence comes at a very fast pace. A business like ours must acquire a reputation for outstanding performance, but in addition we must retain vital and dynamic leadership by exploiting new products and new technologies for ever changing markets. In other words, there must be a continuous injection of new ideas and innovations. RCA has been most successful when we have pioneered as we did with color TV.

We cannot foresee the future, but we should be able to perceive more clearly and project more skillfully than our competition can the future significance of the past. We must create our own future by concentrating upon advanced communications concepts and technology so that we can leapfrog competition. Those products which are standard in today's market must be analyzed in terms of 1970 requirements. Specifications reflecting those requirements can then be hypothesized, and special techniques permitting implementation of those requirements can be conceived and developed. For growth and expansion this is where our future lies.

Techniques and system study contracts, together with well planned internal programs, will produce new opportunities for RCA with new customers. This is the basic reason we have initiated the CSD ADCOM-70 program, and a number of the specific projects included in ADCOM-70 are discussed in this issue of the RCA Engineer.

As part of one of the world's most broadly based electronic companies, we in CSD intend to continue our leadership in support of space and defense communications in all its forms.

J. M. Hertzberg

J. M. Hertzberg,
Division Vice President and General Manager,
RCA Communications Systems Division

RCA ENGINEER

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

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Editor's Note: Electronics gives every indication of becoming an industry which affects all other industries in one way or another. The electronics industry continues to grow because of the hard decisions made by engineering managers torn between the competing demands for profits and for technical progress within their respective companies. This is the theme developed by Dr. George H. Brown, in a talk last Fall at an IEEE Tokyo Section meeting; his complete speech is published here. (We have been notified, as we go to press, that Dr. Brown is to receive, at the IEEE 1967 International Convention and Exhibition in March, the Edison Medal . . . "for a meritorious career distinguished by significant engineering contributions to antenna development, electromagnetic propagation, the broadcast industry, the art of radio-frequency heating, and color television.")

The Engineer and the Corporation

PRESENT TRENDS IN ELECTRONICS RESEARCH

DR. G. H. BROWN
Executive Vice President
RCA Research and Engineering, Princeton, N. J.

It is indeed an honor for me to address the Tokyo Section of the Institute of Electrical and Electronics Engineers and the Institute of Television Engineers of Tokyo on my first visit to Japan. Thirty years ago, I was at a point in my professional career where I would have considered it appropriate to talk in great detail about a variety of directional-antenna characteristics. Fifteen years ago, I would have considered it my duty to tell you all about the RCA all-electronic compatible color-television system. Instead, let me begin my presentation today by telling you of a disagreeable American Indian chief who lived over one-hundred years ago in the vicinity of Chicago. Or perhaps he was a legendary *samurai* who lived near Kyoto in the ancient days. In any event, he was a man of great ferocity who terrified the people with his surly disposition and his skill as a swordsman.

One day, so the story goes, a hapless traveler chanced to cross his path, which so enraged the *samurai* that he drew his sword and sliced the man neatly in two from his head to his haunches. So clean was the blow, however, and so quickly administered that the traveler did not realize what had happened and continued on his way. He remained unaware of his predicament, in fact, until he accidentally bumped into a stranger in the next town . . . whereupon he literally fell apart.

I suspect that every engineering manager must feel a sense of kinship with this poor traveler, divided as we often are between the competing demands for profits and for progress within our respective companies. On the one hand, we must be concerned with maintaining and upgrading the established products and services which produce the greatest profits for our companies. On the other, we must be concerned with pioneering and introducing radically new products and services which will assure the future of our companies. If we fail in either task, we run the risk of bumping into a competitor whose ability to do both tasks well may put us out of business.

A classic instance of this failure both to maintain established product lines and to plan innovations for the future occurred in the United States in the railroad business during the 1920's. The engineering managers of the major railroad lines were so absorbed in laying tracks and building trains that they completely forgot their business was transportation, not railroading. People did not buy trains, they bought the right of passage to a destination. Thus, when the automobile came along and then the airplane, travelers switched immediately to these as offering better, or more convenient, or faster passage. Instead of pioneering these new forms of transportation, however, the railroads fought them with the result that railroading, except for one or two lines, is a very sick industry in America even today.

CRISES IN ELECTRONICS

Turning to electronics, we see that we too, like the railroads, have had our *crises of identity*, as modern psychologists like to put it, moments when we have had to decide what our mission is.

The first one came in the last half of the 19th century when the big electric companies came face-to-face with the telegraph of Samuel Morse and, soon after, the telephone of Alexander Graham Bell. These were novel applications of electric power, but had nothing to do with its generation. After mulling the problem over briefly, the electric utilities resolved to forego the chance to pioneer in these new areas. As a result, a separate industry sprang into being—a communications industry based on electric pulses traveling along networks of copper wire.

The next big crisis came in 1901, when Guglielmo Marconi succeeded in transmitting telegraphic information through the air across a vast ocean instead of over wires. The difference in the two forms of transmission was not too great, however, so the new communications industry was able to adapt itself to wireless. Then Lee DeForest produced the first triode in 1905 and subsequently showed that not only dots and dashes but voice and music could be transmitted through the air. Thus, the companies specializing in communications by wire played an appreciable role in the formation of the new industry of wireless communications by code and voice which got under way after 1910. Following the inauguration in the United States of the National Broadcasting Company in 1926, this latter industry diversified to become a home entertainment medium as well as a communications medium. Thereafter, sound motion pictures and the phonograph were added to round out the concept of entertainment as an adjunct of the communications industry.

The need to provide and service the radios, broadcasting stations, phonographs, records and motion-picture sound tracks that were the life's blood of this novel home-entertainment industry made it inevitable that the industry would develop a manufacturing arm as well—as indeed it did in the late 1920's and early 1930's. It was during this period that the electronics industry, as we recognize it today, was born.

DEVELOPMENT OF THE COMPUTER

Oddly enough, the next crisis of identity for our industry did not come with the development of television, as one might suppose. This was a logical extension of the home-entertainment concept, employed the same basic techniques as radio broadcasting, and was a natural, if far more complex, addition to it. The next crisis occurred, rather, with the development of the computer in the late 1940's. Here was no new home-entertainment device, but a machine that could store and process information electronically.

If the tradition set by the electric-power industry in the 1880's with regard to the wire-communications industry had been followed, one would have expected the home-entertainment industry to back away from the computer and to have allowed it to spawn still another quite-separate industry. But we did not let this happen. Rather, we have adopted it as our own and are pioneering its development and application with the greatest vigor.

I firmly believe that our decision to do so represents one of those profit-or-progress decisions to which I alluded earlier. It is costing us a great deal to prosecute computer technology and much of our present profit is going into the effort. I think, however, that it assures our industry of even bigger financial returns and even wider social impact in the future.

The decision to retain development of computers within our industry is only the most spectacular of several similar profit-or-progress decisions which we have had to make in the past decade. In fact, we have turned to new ways of making decisions following development of the transistor in 1948.

In the past ten years, we have had to decide whether to change the components base of our industry from electron tubes to transistors, and then again from transistors to integrated circuits; whether to enter the low-power field with products that convert other forms of energy directly to electric power through such phenomena as the photo-voltaic effect and thermoelectricity; whether to enter entirely new markets with such things as powerful magnets based on the phenomenon of superconductivity or office copying machines based on dry electrostatic processes, and many more. It has been a difficult decade for engineering management, to say the least.

CATALYTIC ROLE OF GOVERNMENT

Of course, if I am to be perfectly frank with you, I must admit that the electronics industry itself has not been complete master of its own fate in these matters. In the United States, for example, the Federal Government has played a very important catalytic role in helping us make many of these decisions. Its space programs, its defense establishment, its commitments to international cooperation and development, and its desire to solve the problems of urban deterioration, traffic congestion, air pollution, water conservation and the rest have developed within the nation a ravenous appetite for technological change and innovation . . . an appetite which has prodded us to try new ideas and new ways of doing things.

Still, we could seek to meet these needs and to fulfill the terms of the various government contracts that go with them without ever attempting to convert the new findings and new techniques so gained to broader social ends. In fact, many American companies do just that and are quite content to make the government their only customer. This is not the case for our industry as a whole, however, especially not for such large diversified companies as RCA. To us, government contract work is not a fulcrum for our established businesses.

For this reason, I would say that the difficult product decisions forced on our engineering management over the past dozen years have derived not from government insistence but from an innerchange which, for various reasons, has caused us to re-interpret our relationship to society, to deepen our grasp of the physical principles and phenomena on which our technologies are based, and to discover a new willingness to innovate and diversify even if such moves threaten to upset established patterns.

BECOMING A NEW BASIC INDUSTRY

Thus, from an industry that was largely concerned with radio communications and home entertainment in 1940, we have become an industry that is concerned with direct energy conversion, electrical controls, information processing, heating and cooling, printing, photography, medicine, education and a host of other human activities too numerous to mention. In fact, I believe that electronics is well on its way to becoming a new basic industry from which all other industries will derive or on which all will depend for their existence.

One easily gathers this impression by realizing that what J. J. Thomson really discovered in 1897 was not the electron so much as that all matter is electronic; that matter consists of positive and negative charges that can be thrown out of equilibrium and made to do work. This insight was later deepened by Planck, Einstein, Schröedinger and others to account for almost all known physical effects that occur in nature except those stemming from nuclear processes.

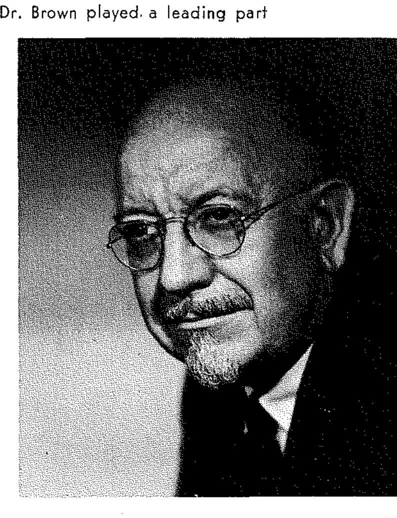
As you know, we built our whole industry on this one tiny insight. For a long time, however, we could not figure out how to use electrons in their natural locale—on the surface or deep in the bulk of various elements. So we began the practice of boiling them off into a vacuum where we could get hold of them with electric and magnetic fields. Of course, this denied us access to the full spectrum of phenomena to which they can give rise in their natural habitats. But, even so, it gave birth to radio, television and radar, not to mention the electron microscope.

Not that we failed to realize what we were giving up. We knew but we had not as yet developed our knowledge of materials and materials-processing to a point where we could tailor matter to produce all the useful effects we sought. The first successful commercial attempt to do so, I would say, led to the thin-film photocathodes and zinc-sulfide type phosphors which brought television into being in the late 1930's.

THE DECADE OF DECISION

However, our ability to exploit the electronic nature of matter in all its variety did not really come until the development

DR. GEORGE H. BROWN studied at the University of Wisconsin, receiving his BSEE in 1930, his MS in 1931, and his PhD in 1933. In 1962, the University of Wisconsin awarded a Distinguished Service Citation to Dr. Brown for his leadership in industry and engineering. In 1933, Dr. Brown joined the RCA Manufacturing Co. in Camden, N.J., as a research engineer. In 1942 he transferred to the new RCA Laboratories research center at Princeton, N.J. During World War II, Dr. Brown was responsible for important advances in antenna development for military systems, and for the development of radio-frequency heating techniques. He and his associates also developed a method for speeding the production of penicillin. At the end of the war, Dr. Brown received a War Department Certificate of Appreciation "for his outstanding work in the research, design, and development of radio and radar antennas during World War II." From 1948 to 1957, Dr. Brown played a leading part in the direction of RCA's research and development of color and UHF television systems. In 1952, he was appointed Director, Systems Research Laboratory, RCA Laboratories. In 1957, he was appointed Chief Engineer, RCA Commercial Electronic Products Division, Camden, and six months later, Chief Engineer, RCA Industrial Electronic Products. In 1959, he was appointed Vice President, Engineering, Radio Corporation of America, and became Vice President, Research and Engineering, in 1961. He was appointed to his present position in 1965. That same year he was elected to the Board of Directors of RCA. A prolific inventor, Dr. Brown holds 79 U.S. patents; he is included in American Men of Science. Dr. Brown is a Fellow of the IEEE and the American Association for the Advancement of Science, and a member of Sigma Xi, the Franklin Institute, and the National Academy of Engineering. He is a Registered Professional Engineer of the State of New Jersey.



of zone refining of semiconductor crystals in the late 1940's and the elaboration of the hole-electron theory in solids. It was at this point that the decade of decision began for the electronics industry.

Our first decision, of course was whether to transistorize all electronic equipment, or as much of it as we could. Concomitant with that was the question of whether to begin phasing out our traditional electron tube technology. Eventually, as you know from your own experience here, it was decided to take a middle course and let transistors seek their own level. Where they proved superior to tubes either for technological or economic reasons, we retired the tubes. Where tubes were superior, we held on and strove to make them even better. This is still pretty much our policy even today.

A far more basic decision that had to be made, however, was whether research and new-product engineering were to be maintained in both the tube and transistor areas, or whether they were to be confined wholly to the latter. This was a difficult decision to make, as much for human as for technological reasons. Eventually, RCA voted in favor of the transistor as did most electronics companies.

As it turned out, this decision led RCA to a complete redirection of its research and development efforts. For example, we expanded our materials research substantially and began rapidly to build a strong staff of solid-state theorists, physicists, metallurgists and chemists. It was during this period also that we opened our research laboratory here in Tokyo as part of an effort to extend our studies of the behavior of electrons in solids.

THREE NEW MARKETS

Not content to simply put us in the solid-state business in markets where we enjoyed a strong sales position—in components manufacture, home instruments, computers, and radio and television broadcasting and transmission equipment—this new staff began to find new materials and new phenomena in such profusion that it seemed we should go out of business just trying to enter all the new markets that were promised. To give you some idea of what I mean, let me mention just three of the new markets into which we have now plunged as a result of these incredibly fertile efforts.

In 1955, a materials research effort at RCA Laboratories in Princeton, New Jersey, led to a technique for incorporating zinc-oxide powders in a specially prepared paper such that electrostatic reproductions of almost any document could be made on it in either black-or-white or color. This work helped to lay the foundations for a vigorous business in office copiers.

In 1960, another materials effort at RCA Laboratories brought into being a vapor-phase transport process for laying down thin films of niobium-tin on a stainless steel ribbon. This seemed like a most unlikely technology until our Electronic Components and Devices organization—formerly a manufacturer of electron tubes almost exclusively—decided it might be used to good purpose in the manufacture of superconductive magnets. After two years of product development, RCA is now a leading manufacturer of such magnets and recently produced one that provides a 137,000-gauss field in a bore 1.9 inches in diameter.

Finally, still another solid-state research effort conducted at RCA Laboratories led to the perfection of a whole new family of field-effect semiconductor devices known as MOS transistors. They control the flow of current by the application of voltage in the classic tradition of the electron tube. It now looks as though these devices, in integrated form, may find immense markets in the home-instrument and computer-memory fields.

These are only three of our most promising recent developments. All have grown out of our solid-state research efforts. To them could be added the perfection of a germanium-silicon alloy with the highest figure of merit for producing electric power from heat, a vapor-phase growth process for growing reliable high-performance gallium-arsenide devices for the first time, and a novel technique for doping silicon solar cells with lithium so that they become virtually invulnerable to the destructive effects of the Van Allen radiation belts above the earth.

I should like finally to mention the laser, probably the most stunning and unforeseen achievement of what I should like to call the new electronics. Nothing illustrates so dramatically the power of research generally and of solid-state research in particular to produce surprises of far-reaching importance in the electronic field than the development of the laser. Its ability to generate coherent light across the spectrum in either pulsed or continuous mode and over a respectable power range has caused us all to consider it seriously for use in a number of areas where radio waves or electron beams would have been used before. Of course, this has also hurt the laser a bit since the technologies for using radio waves and electron beams are already well established.

Nevertheless, we have had to respect its potential. And, even as we have begun to see special applications where it might prove competitive, research has recently found an altogether unique application for which it alone is suited—*holography*, the new technology for using coherent light to produce interference images including those that can be reconstructed in three dimensions. Suddenly, research has catapulted electronics into yet another field outside its traditional domain—the field of visible optics. In fact, I expect lasers and holography to open vast new markets to our industry in the next decade.

ACQUIRING A NEW MATURITY

In summing up, I should like to remind you once again that as engineering leaders and managers in the electronics field we are very much like the traveler who met the *samurai* in the old legend. Our role has been neatly divided into two distinct responsibilities. The first is to help our companies make a profit. The second is to help them make progress.

During the years when the electronics industry was emerging from the telegraph and telephone industries which, in turn, had grown out of the electric power industry, it acquired a unique but somewhat restricted character centered on the technologies of radio communications and home instruments manufacture including television. Later, it acquired still another facet with the invention of the digital computer and its entry into the field of data processing.

Following achievement of a practical semiconductor technology in the late 1940's, however, and the rapid spread of its effects into all phases of electronics, our industry has acquired a new maturity. This is shown, in part, by the moves we have made to re-interpret our role in society, to deepen the intellectual content and sophistication of our products, and to innovate and to diversify our efforts in order to meet the mushrooming demands of society.

As a consequence, we have become richer, healthier and more influential than ever before in our history. In fact, we give every indication of becoming a new basic industry on which all other industries will eventually depend in one way or another.

Like the Prince in the Lady Murasaki's books of long ago, electronics has progressed over the past fifty years from an improvident but charming youth of broadcast antics and home-entertainment escapades to an adulthood of the widest possible interests—a change which may yet make it the keystone of all the industries.

THE CSD IDEA LABORATORY

Ideas may develop gradually over long periods of time or they may blossom spontaneously and unexpectedly. Regardless of how or when ideas or hunches emerge ready for evaluation and test, the innovator requires laboratory space, time, support services and encouragement. This paper tells how CSD's Idea Lab provides these necessities for its engineers.

D. SHORE, Chief Defense Engineer*
Defense Electronic Products, Camden, N.J.

In defense electronics, fierce competition is the watchword. This challenging competition from many electronic manufacturers is the driving force behind a continuous technological race pushing the limits of the state-of-the-art.

The race affects all areas of technology, but very often a single technique or device will forge ahead and promise to outstrip and even replace competitive equipment. For example, the military currently is interested in ultra-portable personal communications gear, movable microwave centers, and computerized switching for entire communications networks.

INNOVATION IS ESSENTIAL

Such concentration of customer interest demands a corollary concentration of effort on the part of the defense industry. One of the prime attributes of any defense-oriented engineering activity is its ability to innovate. If there is one thing that will make an engineering organization fall behind the competition, and even decay, it is lack of innovation.

Innovation, while it can be fostered by management, must spring from the creativeness of the individual engineer. To be truly effective, an engineer must continuously fan his spark of creativity. About one in ten of all employees in CSD is an engineer actively engaged in the design of military communications electronics. The employment of the other nine depends, in a large measure, on how well that engineer does his job. Innovation and creativity are major factors of our success.

Obsolescence

Engineering obsolescence has been a concern in Communications Systems Division for some time. Sometimes it occurs because the engineer does not continuously expand his knowledge. Then too, the dictates of his job require concentrating within the narrow confines of a highly-specialized engineering regime. Premature obsolescence continues to be a major deterrent to effective creativity.

Continuing Education

The processes of continuing education include, of course, the technical literature and symposia. A comprehensive after-hours program of technical courses is attended by many CSD engineers. More than a third of all CSD engineers are currently enrolled in one advanced mathematics course. CSD also conducts an in-plant lecture series at which engineers can learn about emerging techniques from individuals concerned with the development of the technique.

CSD ENGINEERING STRUCTURE

Engineering in CSD is structured to release the engineer from the more routine operations normally associated with his prime work of creating new and better electronics. Support services and direct-support personnel, such as technicians, draftsmen and technical writers, constitute a large percentage of the Engineering Department personnel. They free the individual engineer to exercise the *creative* side of his skill.

CSD's Engineering Department is organized on a project team basis. The individual engineers are assigned to work in a particular regime until the specific project is completed. The engineer's reassignment may then take him into another area of endeavor, contributing to a constant broadening of his capability.

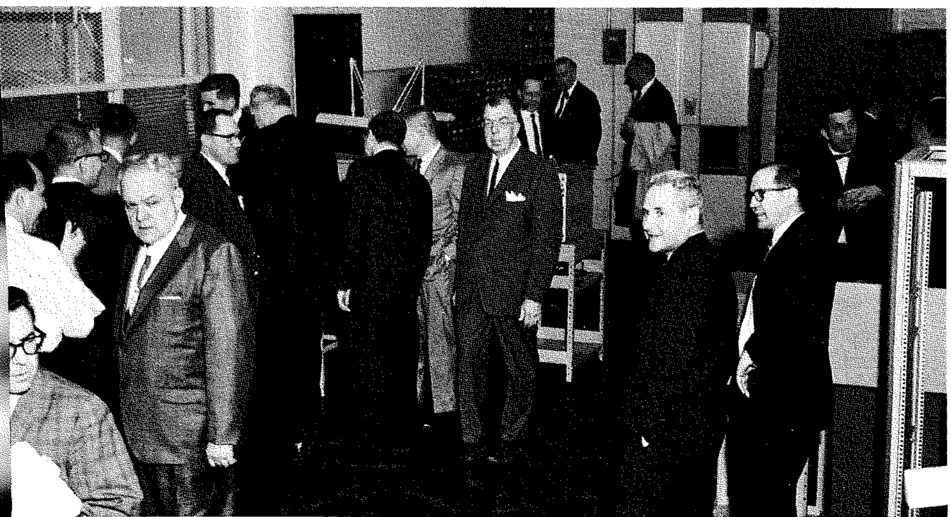
A LOSS OF CREATIVITY

In certain instances the good ideas of some engineers have been going to waste. This happened for a number of reasons, *some* of which are listed here:

1) The idea lost out in competition for funds because other ideas were better —or had better substantiation. Often a modest amount of experimentation might have proven or improved the value of the idea. But no mechanism existed for carrying out that experimentation.
2) The creator of the idea might have gotten out of touch with specific technological advances such as integrated circuits, so he was diffident about proposing the idea for fear of appearing foolish.
3) The idea was outside the product line of Communications Systems Division and therefore not coincident with Engineering Department plans.

Management and individual engineers have been trying for years to overcome these and other creativity constraints. Individual engineers tried experimenting at home. However, our complex

Fig. 1—"Opening day" at the CSD Idea Lab.



Final manuscript received December 8, 1966.

*Mr. Shore was Chief Engineer, Communications Systems Div., when he wrote this paper.

technology very often requires much highly-specialized test equipment to evaluate results. So the home do-it-yourselfer was hampered, to say the least.

Unrelated Ideas

For ideas that appeared to lie outside the Communications Systems Division product-line charter, management ordinarily had but one choice to make. Effort was discouraged; and in no event could it be funded. Sometimes ideas were also lost because management didn't always have the clear vision granted to the inventor on where an idea could lead had it been pursued.

"TLC" Needed

Ideas—it has been said—are a dime a dozen. Certainly there has been no dearth of ideas in the communications profession. Any brainstorming session offers proof that this tenet applies equally in all fields of endeavor. One of the most important ground rules for brainstorming is that no idea be evaluated or spoofed prematurely. This is because ideas—especially the most advanced ones—are like very delicate flowers. They wilt in cigar smoke—and are sometimes soluble in pure logic. It takes tender loving care (TLC) to translate an idea—a thought—into an innovation or an actuality.

DAVID SHORE received his BS in Aeronautical Engineering at the University of Michigan in 1941 and his Master's in Physics at Ohio State University in 1950. Prior to joining RCA, he was with the Air Force's Wright Air Development Center, where he became Civilian Chief of the Systems Analysis Group and was responsible for the initiation of several new aircraft and guided missile systems. Since joining RCA in 1954, he was responsible for the overall systems design of BMEWS, the Ballistic Missile Early Warning System; and has directed the development of satellites. In 1962, he founded SEER (Systems Engineering, Evaluation and Research), with responsibility for directing company-wide efforts on major new systems. In August, 1965, Mr. Shore was appointed Chief Engineer of the Communications Systems Division in Camden and in December, 1966 was appointed Chief Defense Engineer, Defense Electronics Products, RCA, Camden, New Jersey.

ENTER THE "IDEA LAB"

To help provide TLC for innovation and to offset creativity constraints, CSD Engineering Management established the Idea Lab (Fig. 1) in 1966. The *Idea Lab* is envisioned as a complete electronics engineering shop where members of CSD's engineering department can work on *their own* ideas, at *their own* pace, on *their own* time. In the Idea Lab, each participant can cultivate his idea at no sacrifice to the extremely important day-to-day task of supplying our government with its military communications.

The need for some restraint was immediately apparent. The basic purpose of the Idea Lab would be defeated if every member of the engineering staff crowded the lab with all kinds of ideas, to the detriment of some truly fine ideas.

The restraints adopted are minimal. Unsound notions such as perpetual motion are discouraged. Then too, the idea must be generally related to the business areas in which Communications Systems Division is chartered. Since the CSD product line includes radios, computers, switching systems, magnetic recorders and television systems as well as many of the incidental mechanical devices, only in a few isolated cases has it been necessary to make a liberal interpretation of the charter.

ADVISORY BOARD FORMED

A Senior Staff Engineer, Dr. T. T. N. (Nels) Bucher, was chosen to organize and direct a review board and operating committee for the Idea Lab. Dr. Bucher's competence and forward-looking basic outlook is understood and respected by the members of the entire CSD engineering organization. He assembled the Idea Lab Advisory Board. Its members are leaders and engineers from each of our major groups, carefully chosen for their initiative and creative abilities.

The Advisory Board established the basic operational rules for the Idea Lab:

1) An engineer with an idea to pursue fills out a single page application listing the pertinent data on his idea.
2) The Advisory Board selects projects (Fig. 2) on the basis of applicability to divisional product line, magnitude of effort or support required, and availability of facilities.
3) An engineer with an approved idea works in the Idea Laboratory at his own pace on weekday evenings (until 10 p.m.) or during the day on Saturdays.
4) At the conclusion of his program the engineer submits a two-page report of his accomplishments or findings.

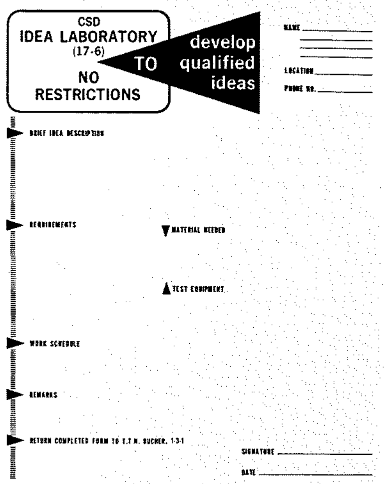
The application form and operating procedures (Fig. 3) are designed for simplicity. The committee felt that ideas should not be smothered by administrative detail.

The Advisory Board prepared and distributed the applications (Fig. 3). The engineers responded and the Board reviewed all the applications for approval. Remember that the Idea Lab Advisory Board, too, is composed of engineers and engineering leaders who still have their day-to-day work to perform.

SETTING UP THE LAB

The Advisory Board also had to physically set up the Idea Lab. The site was selected and equipped with benches and cabinets. A full range of diverse power supplies was made available. Hand tools were procured quickly. A little more difficulty was encountered in acquiring

Fig. 2—The author (David Shore, with pipe, then CSD Chief Engineer) during an Idea Lab planning session with (clockwise), T. T. N. Bucher, J. L. Deppe, R. A. Howell, M. Rosenblatt, and H. Knoll.


CSD
IDEA LABORATORY
(17-6)
TO
develop qualified ideas
NO
RESTRICTIONS

NAME ______
LOCATION ______
PHONE NO. ______

BRIEF IDEA DESCRIPTION

REQUIREMENTS

MATERIAL NEEDED

TEST EQUIPMENT

WORK SCHEDULE

REMARKS

RETURN COMPLETED FORM TO T.T.N. BUCHER, 1-3-1

SIGNATURE ______
DATE ______

Fig. 3—Idea Lab application form.

Fig. 4—Programmable logic idea by S. Krevsky, left, Chairman of New York Idea Lab, and C. Atzenbeck, CSD engineer, New York.

test equipment. Environmental equipment so necessary for the true evaluation of new eletronic and mechanical devices rounded out the permanent facilities.

Stock bins were installed, and filled with resistors, potentiometers, capacitors, choke coils and hardware. Stock also included standard size chassis and blank circuit boards. A CSD Engineering budget allocation was made to set aside funds for special purchases by the Idea Lab Advisory Board. Such purchases might include special tools, electronic components, and various sophisticated metals or plastics required on specific idea projects.

Within weeks, the Idea Lab was fully implemented and formally opened by our Division Vice President and General Manager, Mr. Joseph M. Hertzberg on May 2, 1966 (Fig. 1).

A short time later, the entire process was repeated in CSD's Advanced Communications Laboratory at 75 Varick Street in New York City. One major innovation was made in the ACL Idea Lab, however. Some of the New York engineers wanted to work on their Idea Lab project during their lunch period. Apparently this serves the dual purpose of working on ideas and battling the corpulent bulge simultaneously.

TYPICAL IDEA LAB PROJECTS

Both CSD Idea Labs are now functioning. The typical idea projects (Figs. 4 and 5) described below illustrate the type of ideas being pursued.

One idea concerns a 16-level AM, 9.6-kilobit/second wire-line modem. The objective is to increase the amount of useful information which can be transmitted. Using spare modules, a few additional parts and a liberal helping of personal ingenuity, the idea proposer will explore the feasibility of his approach.

One engineer stated he had been waiting 18 years to prove a theory of his regarding selection of thermally coexpansive materials in gear trains. In 1948 he actually designed a gear train using various materials of complementary thermal coefficients of expansion. This train required no idler gears to compensate for changes in temperature. He was going to experiment and document his findings after his retirement from business; but, now he is proceeding with this work in the Camden Idea Lab.

An idea proposer had a suggestion for a contactless frequency selective relay. This enthusiast had tried at home to prove that the idea was feasible. Now with the Idea Lab he has management's help (regardless of lack of an active project budget)—and best wishes—to do just that.

Another idea concerns a computer program for a delay optimized filter. In line with the ever-growing use of computer-aided design, a shared-time computer console has been made available for this idea and others like it.

Idea Lab programs also include:

1) Dual feedback demodulators
2) MOS array experiment
3) Analysis of an information storage and retrieval system for engineering design data.

One program is being conducted by a nine-man team to provide the benefits of advanced technology (and an objective point-of-view) for the field of medicine. As electronically oriented engineers, these men lack medical knowhow. Fortunately an RCA arrangement with Hoffman-LaRoche, Inc., plus the support of Jefferson Memorial Hospital personnel provide this essential ingredient to our team.

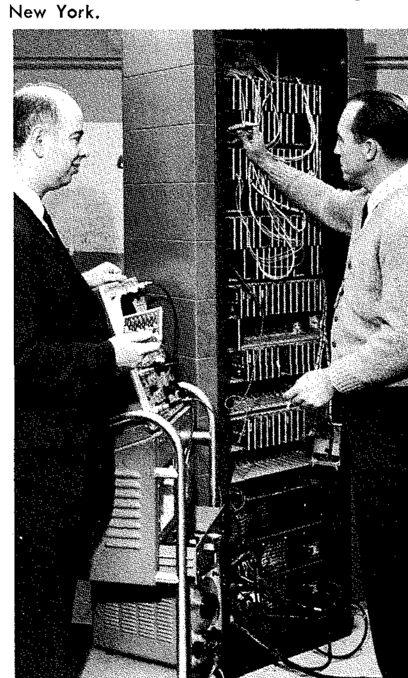
These are representative programs. The participants include inventors, analysts who want to put the rivets in their own dreams, men who want to get up to date in technology, and heretofore frustrated humanitarians.

CONCLUSION

All the Idea Lab engineers have one objective in common; they are working hard at self-improvement. The Idea Lab affords enginering management an extremely valuable, personalized means of participating in that improvement.

There may not be a single, business-oriented idea that bears fruit in the Idea Lab. But every idea is a good one. The best idea in the Idea Lab is that every participant will grow. There is no better way for an organization to advance than through the growth of its individual members.

Fig. 5—9.6 kilobit modem idea by D. P. Goodwin (left) and G. Meslener CSD, engineers, New York.



ADCOM-70
Advanced Communications for the 1970's

ADCOM-70 is the program adopted by the Communications Systems Division to enable CSD to enlarge its share of the military communications market in the 1970's. The comprehensive program affects all areas of the division including the organization of its Engineering, Marketing and Manufacturing activities. In ADCOM-70, major emphasis is put on the development of advanced techniques that will make CSD's products unique in the 1970's.

J. M. HERTZBERG
Division Vice President and General Manager
Communications Systems Division
DEP, Camden, N.J.

The Communications Systems Division has adopted a comprehensive program designed to foster substantial growth for the division. Called AdCom-70—*AD*vanced *COM*munications for the 1970's—the program affects all areas of CSD operations. Its objective is to enable CSD to leapfrog the state of the art in developing major products for the military communications market in the 1970's. Added to CSD's normal business, the new AdCom-70 products will provide the impetus for the projected growth of our division.

In establishing AdCom-70 late in 1965, we identified four types of communication equipments that would benefit most from the AdCom-70 efforts. These are: 1) Microwave/Tropospheric Scatter/Frequency Division Multiplex; 2) Military Information Data System; 3) Small Advanced Radio Sets; and 4) Data Communications Units.

In 1965, our planning data indicated we would maintain a business level of between 100 and 120 million dollars per year through 1970. This is a pretty good business level in anybody's book, but we were not satisfied that this represented as large a share of the market as we could obtain if we did our sales engineering and production work well.

Final manuscript received January 6, 1967.

THE ORGANIZATION

After months of functional analysis of our business and our organization, we evolved a new organizational concept for the major operating elements of the division. Engineering and Marketing were reorganized as of Feb. 1, 1966, around four major product lines—Light Communications Equipment; Heavy Communications Equipment, Digital Communications Equipment; and Recording and Television Equipment. Manufacturing was realigned two months later.

The Marketing Department (Fig. 1) now includes individual sales and contract administration functions in each of the four major product line areas. Additionally, a business planning group reports directly to the Division Vice President, Marketing. This planning group is structured along product lines.

The Engineering Department (Fig. 2) has the same four major product line areas. The similarity in Marketing and Engineering organization fosters better communications between these "business-oriented" organization who work together in "business teams."

BUSINESS TEAMS

In each of the four product-line areas, we have established a "business team" made up of *functional* groups of marketing men, engineers and production specialists. Each man on a team is selected for his ability to think in terms of the overall team objective rather than of problems solely in his specialty area. Further, each man has learned to think not only of today's task, but also how that task relates to the technology, the customer, and the market of the 1970's. The business teams for the four major product areas are composed of specialists and supported by other specialists. The special actions and functions are done by a specialist *but* always as a member of the *team.*

The goal of each of the four major product teams can be stated simply: To

JOSEPH M. HERTZBERG studied Electrical Engineering at the University of Michigan before joining the Stromberg-Carlson Corporation in 1930. During World War II, he was active in the management and technical coordination of numerous Air Force communications, radar and navigation programs. He was awarded the Legion of Merit and the Order of the British Empire for his wartime services. Mr. Hertzberg joined RCA in 1945 as a Senior Communications Engineer. Within two years he was appointed Aviation Marketing Manager. In 1951 he became Assistant to the Manager, Government Marketing; in 1955 Manager, Airborne Systems Department, Defense Electronic Products; in 1957 Manager, Defense Marketing; and in 1958, Vice President, Defense Marketing. In 1957, he received the RCA Victor Award of Merit. During 1960 and 1961, he was with the Philco Corporation as Vice President-Marketing, Government and Industrial Group. He rejoined RCA in 1962 as Division Vice President, Defense Marketing. In 1965, he was appointed to his present position as Division Vice President and General Manager, Communications Systems Division. Mr. Hertzberg is a licensed amateur radio operator. He is a member of the Air Force Association; Armed Forces Communications and Electronics Association; American Ordnance Association; Association of the U.S. Army; American Society of Naval Engineers; and IEEE.

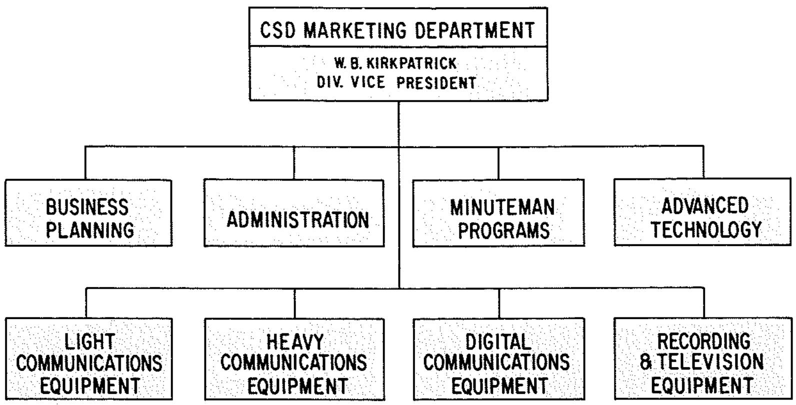


Fig. 1—CSD marketing department.

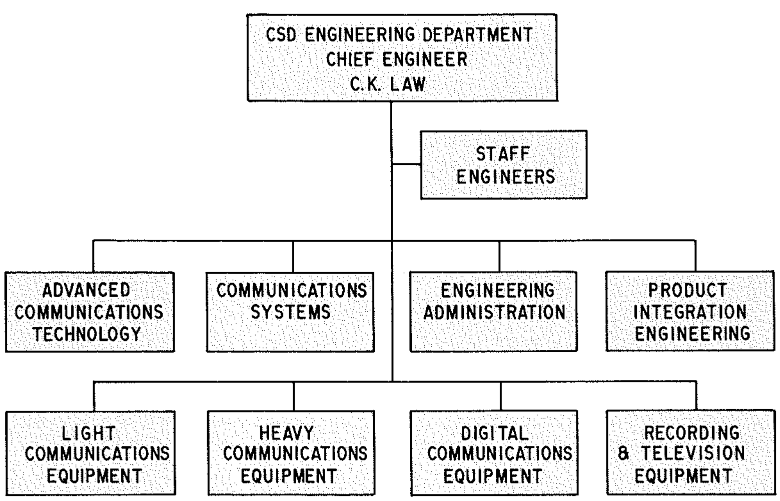


Fig. 2—CSD engineering department.

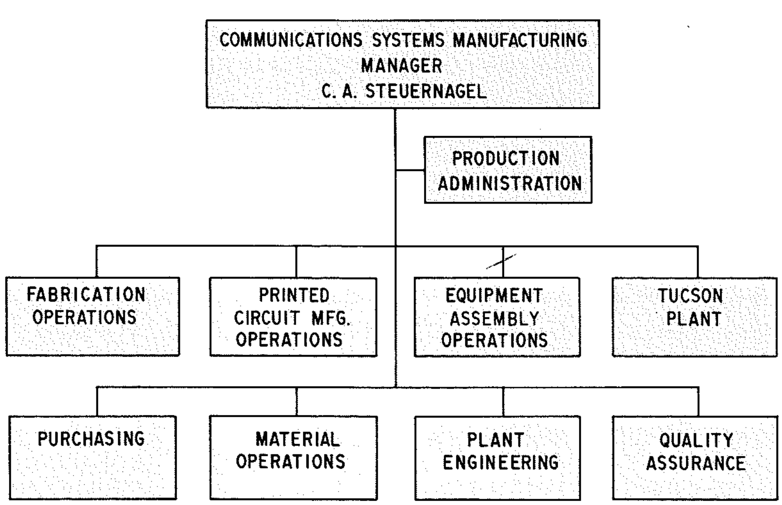


Fig. 3—CSD manufacturing organization.

develop an advanced product in its product area.

For example, the sales manager of the light communications team provides information on the customer's current needs and the field communications radios currently in use and in development. He, the engineers and the light communications planners then evaluate our current light communications products against projected 1970 criteria. The production specialist contributes his estimates on the producibility of the product. The contract administrator adds to his knowledge gained from an intimate day-to-day working relationship with the customer.

The team correlates all of its inputs, estimates the normal advance in technology, and establishes a normal evolutionary product for the 1970's. This product is then evaluated on the basis of the customer's stated and anticipated needs. Anticipated technological breakthroughs or those necessary to make our competitive position unique are then discussed with advanced technology engineers and planned into the program.

After having done this thinking on the most promising products, the team comes to a major decision point. Together, the team members:

1) Select specific targets and establish specifications.
2) Determine the technical objectives and schedule their accomplishment, including basic milestones and decision points.
3) Fund the program and assign the people to pursue the objectives.
4) Obtain supplementary customer support, through study contracts, advanced technique contracts, and design contracts.
5) Monitor the program and revise if necessary.

PROJECT IMPLEMENTATION TEAMS

When a project or contract is undertaken by CSD, a project implementation team is set up, one for each job. These teams are continuously tailored to the changing requirements of the job.

Each project team has a leader who has the maximum practical authority over all elements of effort on his job. He has assigned to him, both functionally and administratively, the personnel he needs to fulfill the job requirements; these include engineers—design, test, reliability, maintainability, etc.—technicians, draftsmen, technical writers, administrators, cost analysts, etc. This team is located together and works together to get the overall job done. When a team member completes his work on the project, he goes to a new team or returns to his technology or engineering integration group for reassignment. While assigned to a project implementation team, each specialist receives direction on job standards and techniques from his technology group or engineering integration group, as applicable.

The project implementation teams also are classified into the four CSD product-line areas. In Engineering, each of these product line areas is headed up by a manager (Fig. 2) who has two deputy managers—one for technical control and the other for business control—to support him in running his product line.

SYSTEMS ENGINEERING GROUP

We have established a Programs Development group to analyze customer needs and to concentrate the resources of CSD engineering in our new business efforts. This group is composed of a cadre of specialists in applications engineering, proposal management, management methods, product planning, publicity and presentations. We recognize that new business, in addition to design and production contracts, must take the form of development and feasibility programs. The programs Development group is flexible so that personnel can shift with projects from initial precontract phases to the proposal and subsequently into design, program management, or production.

SYSTEMS ENGINEERING

The CSD systems group becomes involved in programs in the early stages of the material acquisition cycle, starting with the operational requirements and continuing through concept formulation and program definition. The group's technical effort may include system design studies and analyses concerned with mission parameters such as operations, threat, survivability, vulnerability, maintain ability, logistics support and mission relationships as well as trade-off analyses among system parameters. In the course of this effort, the systems group uses the information gained to guide the development of new, or the modification of existing, CSD equipment to meet proposed specifications. The systems group also prepares data to be presented to the military customer so that future specifications will take into accord advanced developments achieved by our division.

The systems group works closely with the advanced technology, product development and manufacturing groups during the early phases of a program.

Wherever possible the systems group obtains contractual coverage for its effort. This allows CSD to become a "partner" with customers in potential new programs and helps us conserve scarce internal funds.

ADVANCED COMMUNICATIONS TECHNOLOGY

Another area of specialized engineering endeavor is that of providing the new technology to enhance today's product and to ensure tomorrow's. This new technology (and "business-team" participation regarding it) is in the province of the Engineering Department's Advanced Communications Technology Group.

This group maintains our Advanced Communications Laboratory in New York City where almost 100 engineers are responsible for advanced techniques and for the synthesis of advanced concepts. Additionally, the Adavnced Communications Technology has engineers located at the "grass-roots" with project implementation teams. These men serve the dual purpose of getting the advanced technology into projects, and bringing the recurring day-to-day engineering needs to the attention of the advanced thinkers.

MANUFACTURING

With all this advanced thinking, what of the men who must put the rivets in our dreams. The Camden plant, (CSD Manufacturing), has for over 20 years been one of the world's largest producers of military communications gear. How does this tremendous facility fit into AdCom-70?

The organization (Fig. 3) was adjusted slightly for AdCom-70. Production orders at hand for back-pack radios, field radio repeater stations, and the like presently total over $117 million, so you can see that we are not kindly disposed to any sweeping organizational changes.

However, this load together with anticipated contracts indicated that additional plant space and additional personnel would be necessary. In the year since October 1965 our force of production personnel increased to over 5,100 people. Along with this, an expansion and refurbishing program unparalleled in Camden Plant history has been proceeding according to plan.

In the latter part of 1965 we transferred our white room production facilities from the Cambridge plant to Camden. Here we have built up a relatively high production rate, and we have performed on schedule, within planned cost, and with high quality.

We have continued to improve our printed-circuit production area, to serve not only our own needs, but also those of various other elements of the corporation.

Automatic test equipment for integrated circuits, automatic wire-wrap machinery and even small item assembly conveyor lines are but a few of the items contributing to the "new look" in Camden manufacturing.

The manufacturing operation maintains a well-rounded model shop occupying over 40,000 square feet and staffed by more than 600 well qualified personnel. This model shop not only handles short run production jobs but also builds breadboards and prototypes of the products for the 1970's.

Each of the major operating manufacturing units (i.e. fabrication, printed circuit and equipment assembly) maintains its own short-order capability. This permits the performance of short-run tasks without disrupting the flow and scheduling of the day-to-day operations.

AdCom-70 plans for production of the products of the 1970's are on schedule. More importantly, manufacturing personnel, especially those participating in the business teams have accepted the challenge. They too are looking forward.

SUMMARY

AdCom-70 was inaugurated to fill the need for innovation, and we have structured our organization to accommodate the AdCom-70 concept. How effective have the organization and the AdCom-70 concept been?

You can judge that for yourself, somewhat, by reading some of the other articles in this issue. You will notice, I'm sure, an across-the-board representation in techniques, in fields of interest, and even in the responsibility levels of our CSD authors. Mechanical engineers, electrical engineers, program managers, and our former Chief Engineer have articles published in this issue. Each has discussed his own field; but through his own field each has added to the whole.

AdCom-70 concept is progressing in a quite similar fashion. Each unit of the organization is contributing. The people that make up the organization are contributing not only to the product for today, but to the technology from which the product of tomorrow will be drawn.

AdCom-70 led to a significant and substantial change in our division. This change is more than a mere organizational shuffling of people. The role and function of almost every job has been basically altered.

The contributions of the people have made CSD a truly synergistic organization in which the whole is *more* than the sum of its individual parts.

INCREASING SPACE TT&C CAPABILITY
Telemetry, Tracking, and Command Communications

Space telemetry, tracking, and command (called collectively "TT&C") includes all those space-system communications functions other than the familiar ones in which a satellite (for example, like TELSTAR or RELAY) acts as a traffic relay between ground stations. This paper concentrates on a review of the TT&C technology, with emphasis on requirements of the 1970's for earth-orbital, lunar, and planetary missions, both manned and unmanned. The main technical challenge is to increase TT&C capabilities with less—or at least no additional—burden to the spacecraft. To accomplish this, RCA efforts are focussed on microminiaturization of spacecraft TT&C equipment, new concepts and techniques for both spacecraft and ground-station equipment, and means for dealing with re-entry plasma problems.

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Communications Systems Division
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What is *TT&C*? The term *space communications* is commonly used to include all types of radio-frequency information interchanges between spacecrafts, as well as between spacecrafts and earth. Such exchanges fall into two general categories: 1) the space vehicle acts as a traffic relay between ground communication stations (overseas TV, etc.); or 2) the spacecrafts or their subsystems act upon the information either as an originator (*telemetry-communication*), or as a coherent relay (*tracking*), or as a recipient (*commands-communication*)—which we can call, collectively, *TT&C*. Included in TT&C are the following communication functions:

1) *Telemetry,* including data conditioning, coding, and multiplexing.
2) *Command,* including data-reception demodulating and decoding.
3) *Tracking,* including any active continuous wave or pulse transponding to ground as well as on-board RF interrogation.
4) *Voice communications,* including provision for vehicle-to-vehicle as well as extravehicular-to-vehicle exchanges.
5) *Intercommunications,* limited generally to intravehicular voice communications between crew members of manned spacecraft, but may also include data interchanges where selectable terminals are involved.

The satellite which performs the traffic relay function must also be monitored, commanded, and tracked from the earth, and will generally be provided with a TT&C function separate from its traffic relay function.

Final manuscript received November 16, 1966.

THE CHALLENGE OF THE 1970'S

Spacecraft are being or will continue within the next two decades to be built for three general mission classes: 1) earth orbital, 2) lunar, or 3) planetary. Each of these missions subdivide into manned and unmanned subclasses, and further subdivide into experimental-exploratory and military-operational subclasses. For our purposes, experimental-exploratory subclasses will include spacecraft that may already be developed, but whose mission equipment is for experimental or exploratory purposes. The military-operational subclass would include those spacecraft which have been developed and whose mission equipment is both developed and essentially unchanged. Of course, many military-operational spacecraft must move through the experimental phase prior to achieving operational status.

The classification of some of the current as well as possible future programs during the 1970's is shown in Table I.

From a telecommunications standpoint, the mission classes are significant from three standpoints:

1) Distance or range as a function of mission phase.

TABLE I—Current and Future Programs

Class	Subclass		Currently in Use or in Development	Notes	Planned or Possible 1970-1980	Notes
Earth Orbital	Unmanned	Experimental-Exploratory	NIMBUS, OGO	—	lifting bodies	1
	Unmanned	Operational-Military	TIROS-ESSA, SAMOS	—	weather, navigation, communications	—
	Manned	Experimental	GEMINI, MOL-AAP	2	med. to lg. orb. labs	3
	Manned	Operational	—	—	manned observatories	—
Lunar	Unmanned	Experimental	Orbiter, Surveyor	—	LEM Truck	4
	Unmanned	Operational	—	—	Lunar logistic vehicles	—
	Manned	Experimental			MOLAB	5
	Manned	Operational	—	—	Lunar base	—
Planetary	Unmanned	Experimental	Pioneer/Mariner	—	VOYAGER	—
	Unmanned	Operational	—	—	—	—
	Manned	Experimental	—	—	Mars mission module	—
	Manned	Operational	—	—	—	—

NOTES: 1) "Winged" re-entry vehicles, unpiloted
2) MOL—Air Force Manned Orbiting Laboratory
3) 6 to 36 man space stations may use piloted "winged" re-entry vehicles
4) Unpiloted lunar delivery vehicle
5) Lunar Mobile Laboratory

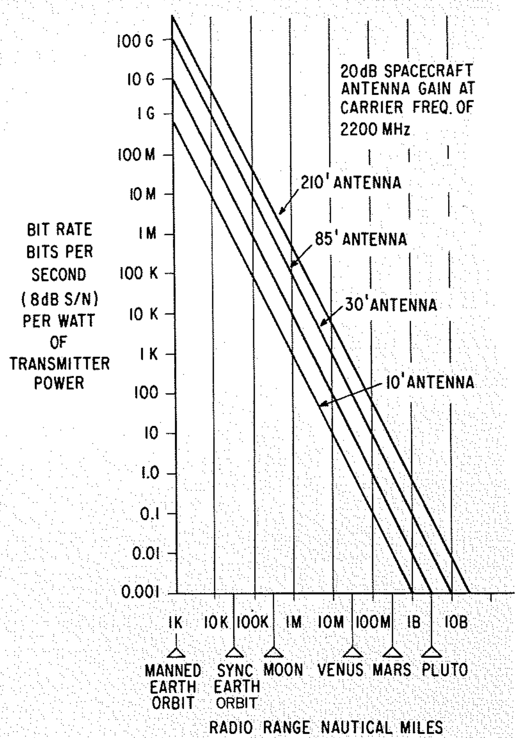


Fig. 1—S-band down-link performance.

2) Communications traffic as a function of mission phase.
3) Position determination accuracy requirements.

The maximum distances for each mission class can be defined approximately as follows:

1) *Earth orbital:* 22,000 nmi (synchronous orbit)
2) *Lunar:* 250,000 nmi
3) *Planetary:* 37 million (Venus) to 47 billion (Pluto) nmi

For our purposes, we can generalize mission phases as follows:

1) Launch
2) Parking orbit (may be final orbit)
3) Transfer (to final orbit)
4) Midcourse (for lunar and planetary missions)
5) Lunar or planetary encounter, orbit and exploration (may also be fly-by)
6) Return
7) Retransfer to parking orbit (primarily for manned systems)
8) Re-entry and recovery (primarily for manned systems)

Within the 1970's, all launches and recoveries will be done on earth. Thus, all spacecraft will have an initial phase where the traffic load for purposes of checkout and monitoring will be quite high but the distances are those for earth orbital. Similarly, a returning probe will likely wind up on earth orbit prior to re-entry and recovery. Thus, data which was not sent during the other phases could be sent during this period. As is the nature of the experimenter, the highest traffic demands fall upon the telemetry information. Were it not for link capacity and ground data-reduction restrictions, the data list on almost any space vehicle program would exceed 1,000 points (some by 10:1) and the resulting telemetry bit rate would be anywhere from 1 to 10 megabits per second (Mbits/s). Typically, the power-weight-volume, state-of-the-art ground support trade-offs have resulted in current system data rates of 100 kbits/s for near-earth, 50 kbits/s for lunar, and 1 to 16 bits/s for planetary distances. Specific mission systems for both operational and experimental spacecraft may require up to 20-Mbits/s transmission capability.

Command requirements are presently modest, and are expected to continue to be for most space missions. It was established very early in space research that close control from earth (which would have required a high command traffic load) was impractical for a number of reasons, mainly the problem of keeping the ground computation elements fed with data that had to be sensed in the spacecraft. A 1 kbit/s command rate is typical of earth orbital and lunar systems, while 1 bit/s has been found satisfactory for planetary missions. In the command case, the higher data will also be used for near-earth phases of planetary mission systems.

Military systems are plagued by a requirement to attempt to operate in a hostile environment, including enemy eavesdropping and electromagnetic jamming. The need in this case is for the maximum obtainable protection.

Data communication between spacecraft is a requirement which provides for emergency telemetry transmission. This data rate is 1.6 kbit/s over a maximum range of 600 nmi. Future spacecraft may well have data exchange rates of 10 times this rate over 100 times the distance.

Voice communication between manned spacecraft having a range requirement of 600 nmi is typical of what has been postulated for a rendezvous operation with earth-orbital space stations, etc. For manned planetary missions, it will be desirable to extend this range as far as possible. It will also be desirable to extend voice communications to earth to the maximum possible range.

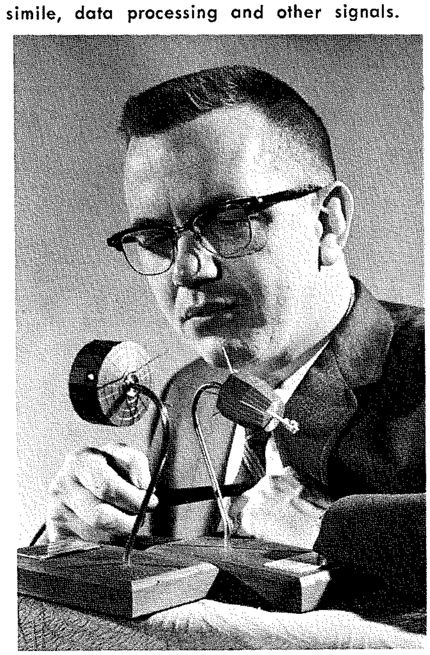
RE-ENTRY COMMUNICATIONS

The so-called "communications blackout" experienced by spacecraft re-entering the planetary atmosphere is of concern in some experimental spacecraft. In the case of lifting bodies, which have lift-to-drag ratios (L/D) on the order of 2:1, this is especially true. The maneuverability range of these "high-

craft. In the case of lifting bodies, which have lift-to-drag ratios (L/D) on the order of 2:1, this is especially true. The maneuverability range of these "high-lift" re-entry craft make them attractive for both civilian and military use. As the L/D increases, the time spent in re-entering increases with a corresponding increase in the interest of real time monitoring and control from the earth. The ionization of the flow field causes a high (50-to-100-dB) effective attenuation (including reflection) of RF energy over a wideband of frequencies. The width of

L. B. GARRETT, JR. received his BSEE from Drexel Institute of Technology in 1953 and his MSEE in 1958. Mr. Garrett joined RCA upon graduating in 1953. He has worked in the design and development of UHF communication equipment, such as the AN/ARC-34 and AN/ARC-62, specializing in frequency synthesizing systems and RF tuners. He was associated with the USAF Time Division Data Link Project, where he participated in the coordination of ground and airborne equipment requirements with government and industry users. He has served as project engineer of a long-range command data link development for advanced interceptors. Prior to his assignment in the Dyna-Soar Program Management Office, he directed company- and government-funded studies on communications for orbital military systems. As a member of the Dyna-Soar Program Management Office, Mr. Garrett was responsible for CSD design analysis and integration, and flight tests. Currently, Mr. Garrett is responsible for the management of certain aerospace communication system studies and analyses, including those for lunar vehicles, lunar bases, LEM and communication satellites.

L. Byron Garrett (author) examines models of two RCA orbital satellites. The TIROS-ESSA (left) is the latest in the series of 13 consecutive successful RCA weather satellites. The RELAY communications satellite (right) established a record of 290 hours of intercontinental television broadcasts and more than 630 demonstrations in transmission of voice, facsimile, data processing and other signals.



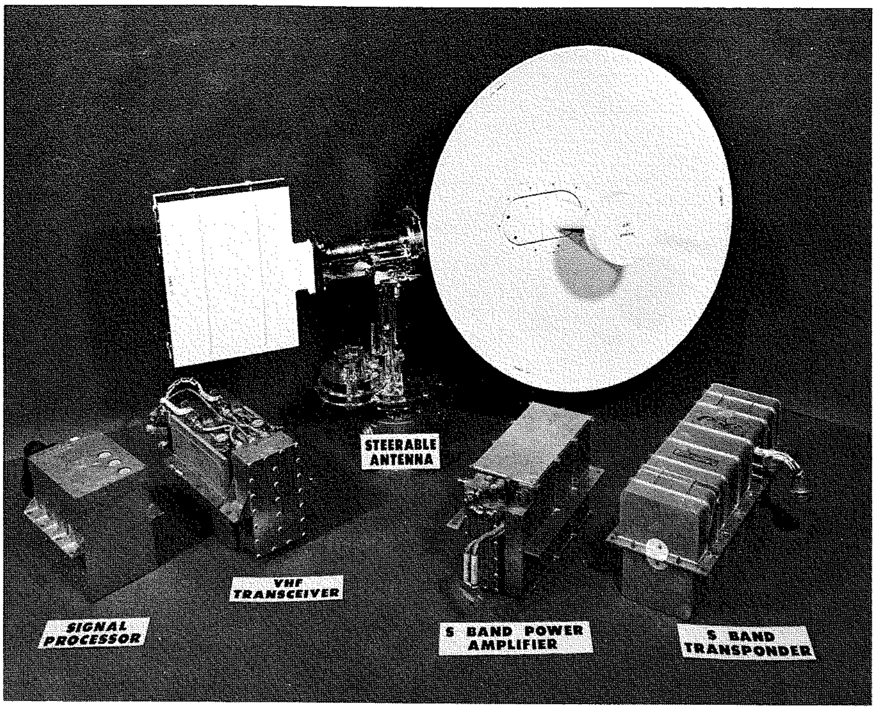


Fig. 2—A composite of representative communications elements.

the band varies with re-entry speed, altitude, and the shape of the vehicle.

Analysis has shown that for typical (24,000-ft/s) re-entry velocities of a high lift vehicle from earth orbit, the blacked-out band is between 100 kHz and 10 GHz extending over a 20-minute period. Re-entry velocities of 34,000 ft/s could extend the blacked-out band up to 100 GHz or higher.

The requirement is to establish a way to achieve minimum communications in the face of these obstacles for experimental efforts that are contemplated in the 1970's.

TRACKING

There is no real limit to the precision desired of earth-derived position and velocity data (typically range, range-rate, and look angles). State-of-the-art limitations in ground antenna precision and geodetic siting accuracies, together with coherent ranging techniques provide position information from a few tenths of a meter near-earth to tens of meters for planetary distances. Depending on the angle between the line-of-sight and velocity vector, velocity determinates can be made virtually independent of distances from a few hundredths to a few tenths of a meter per second.

CURRENT METHODS THAT WILL BE APPLIED TO THE 1970'S

As a result of the other user demands on the radio frequency spectrum in the upper VHF and lower UHF band used for many current spacecraft, TT&C frequencies are being moved higher in the UHF band. NASA and DOD have under development RF equipment which operates in the S-band (roughly 2,100 to 2,300 MHz) for telecommunications purposes. The NASA development is termed *U*nited *S*-*B*and *S*ystem (USBS), and is being implemented in MARINER and PIONEER, as well as some earth orbital programs. DOD is developing the *S*pace *G*round *L*ink *S*ystem (SGLS) which will be implemented in MOL and future military spacecraft. Although presently, compatibility between USBS and SGLS is only partial, it is intended that greater compatibility will be achieved during the 1970's for basic TT&C functions. Both will continue to have differences because of the mission differences.

The S-band unified systems provide full duplex link operation with the spacecraft and provide for the multiplexing of both telemetry, voice, and mission data into a single down-link carrier. Voice command and mission data are also multiplexed on a single up-link carrier. By coherently transponding (locking the transmitter frequency to the received carrier frequency in the spacecraft), doppler information can be extracted to derive accurate range rate information. Through the use of a 1-Mbit/s pseudorandom code originating at the earth and returned by the spacecraft, range can be measured to lunar distances.

Using the same basic receiving and transmitting equipment, but reducing the transmitted and received spectrum, we can extend operation to planetary distances. The relation between the distance and communication capacity for the present S-band equipment is illustrated in Fig. 1, for the several antenna sizes employed.

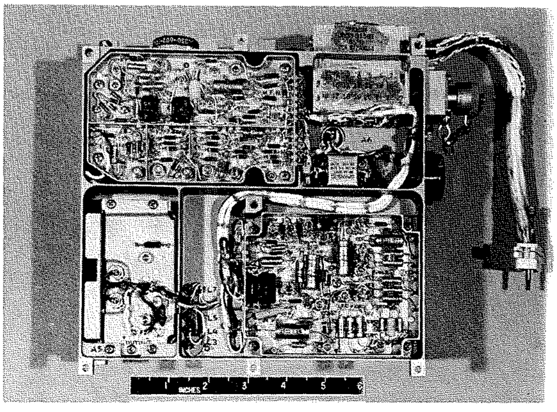
Voice communication between spacecraft continues to be performed in the lower UHF (300-MHz) band and could continue there for lunar distances and beyond. Near-earth operations may have to move to other frequency bands such as L-band (1,400-MHz) because of interference with other services.

Integrated spacecraft communications elements are illustrated in Fig. 2.

The diagram of Fig. 3 illustrates possible TT&C linkage (voice, command, and control) for lunar exploration that can be achieved with equipments currently in development.

A multimanned space station of the MOL variety that will be in use by 1970 will typically have TT&C subsystems which are contained in 3 ft^3 of volume, weigh 200 pounds, require 200 watts of electrical power, and exhibit through use of selected parts and redundancy an effective MTBF of 2 years. Typical unmanned spacecraft may be expected to devote a $1/3$ of these amounts to telecommunications with approximately the same effective MTBF. Although when viewed as a fraction of a 25,000-pound manned or a 1,500-pound unmanned satellite, these do not appear too disturbing. The significance of a substantial reduction in the telecommunications equipment burden and its reallocation to the mission equipment should not be overlooked. Also, without the application of miniaturization techniques, multimanned and multimission space stations will be burdened with a ton or so of telecommunication equipment.

Fig. 4a) Point-to-point wiring in a VHF transceiver.



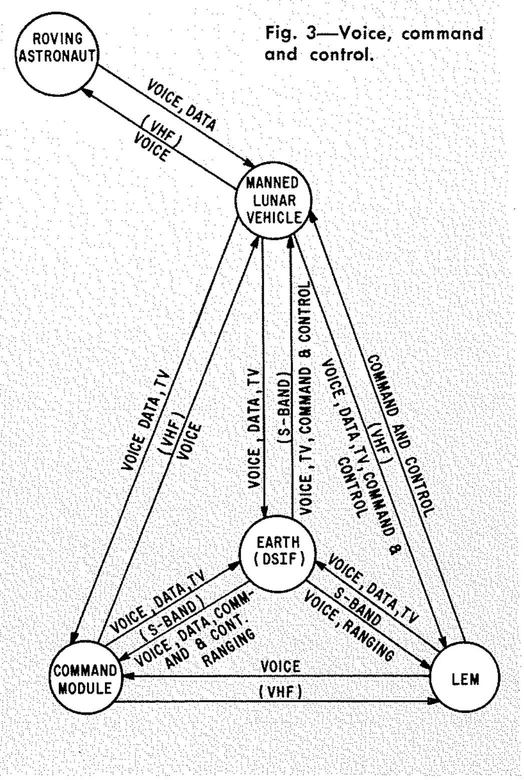


Fig. 3—Voice, command and control.

In the area of re-entry communications, operation above the black band at X-band communications has been successfully achieved. A lifting body at 18,000 ft/s (ASSET) and experiments on GEMINI (24,000 fps) have demonstrated that localized cooling (water) has permitted penetration of the plasma at UHF.

MEETING THE CHALLENGES OF THE 1970'S

The challenge of the 1970's may be essentially summarized as: to provide increased TT&C capability with less—or at least without additional—burden to the spacecraft. To meet this challenge, RCA is proceeding on four major fronts:

1) Microminiaturization techniques in both RF and digital areas.
2) Development of concepts and equipment to obtain greater capability through efficient utilization and organization of spacecraft functions.
3) Development of techniques to achieve higher performance levels in surface equipment.
4) Research in re-entry plasma modification techniques.

Microminiaturization

Microminiaturization through even further application of integrated circuit techniques will permit substantial size and weight reduction. The reduction in size and weight unfortunately does not always go hand in hand with a reduction in power. Integrated circuits for telecommunication functions frequently utilize more active elements, and the total power input requirement may rise slightly rather than decrease. This is particularly true where today's passive elements (i.e., filters) are being replaced by integrated circuit equivalents. RCA's work in this area includes extension of microcircuit technology into the SHF band (Fig. 4).

New Concepts

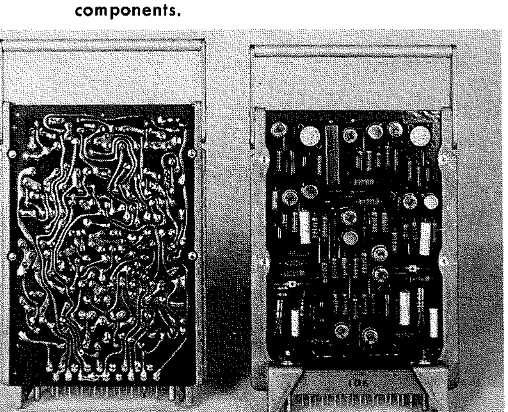
In part, the challenge of reducing the TT&C burden to the spacecraft can be met through further equipment miniaturization while adding capability. This will tend to be offset, however, by the pressure to begin radio link operations in portions of the SHF-band. Equipment in this band tends to have a higher weight per transmitted watt of power than UHF-band equipment (0.6 vs 0.2 lb/w). Another approach lies in making more effective use of equipment through organization of the subsystem.

At the present time and for systems under development, it is common to have separate elements identified for telemetry, tracking, command and data-voice communications often packaged as separate subsystems or equipment groups. In an effort to reduce equipment complexities, it has been found that: 1) equipment may be shared between several functions, and 2) if required functions are similar and if they are properly defined, equipment can perform two otherwise independent operations in a single operation. One approach to achieving this is illustrated in Fig. 5. The relation of the elements shown to the TT&C functions previously discussed for the spacecraft are as follows:

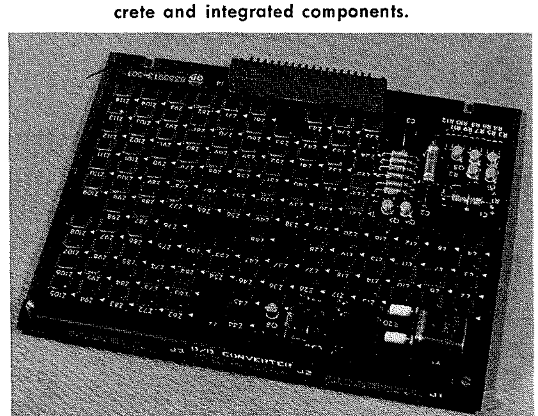
1) *Telemetry function* is performed by the data acquisition elements in conjunction with the RF transmitting equipment. Telemetry multiplexing is a data conversion function which may be aided by data processing. Data collected during periods when the spacecraft is out of network coverage may be stored and transmitted during periods of contact.
2) *Command function* is performed by the RF receiving equipment in conjunction with the data conversion equipment. Routing to a display in the manned system or to a vehicle or mission system (manned or unmanned) may be direct from data conversion elements after decoding or may involve data processing prior to use.
3) *Tracking function* for ground interrogation generally will involve only the RF equipment. When the spacecraft acts as an interrogator, data conversion and processing will be required.

Fig. 4a through 4d—Progression of the technology in RCA's aerospace communications equipment from today's circuits (a, b & c) to tomorrow's integrated circuit (d).

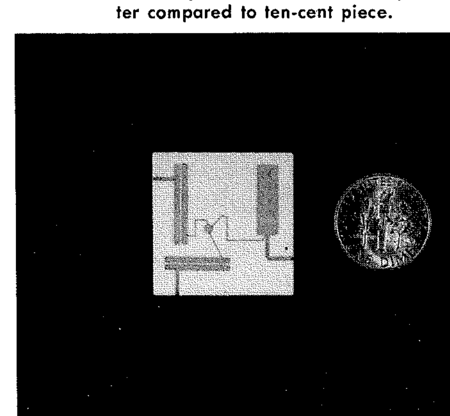
b) A printed-circuit board with discrete components.



c) A multilayer-board converter with discrete and integrated components.



d) An integrated-circuit S-band up-converter compared to ten-cent piece.



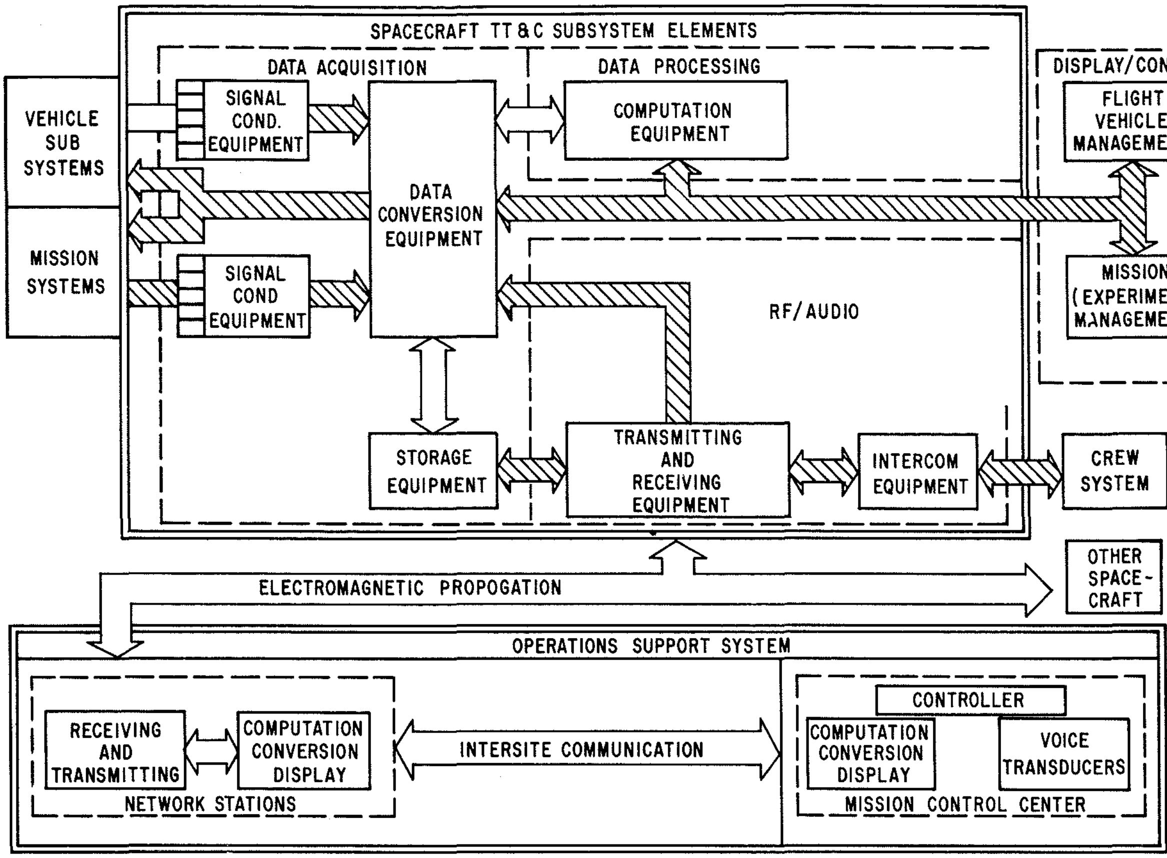


Fig. 5—Elements of space TT&C (telemetry, tracking, and commands communications).

4) *Data exchange* with other spacecraft will involve the Receiving-Transmitting as well as the data conversion equipments.
5) *Voice communication* will involve principally RF audio equipment. Any cryptographic requirements are provided by data conversion equipment.
6) *Intercommunication* will be a separate system for voice and data on small manned spacecraft. For larger spacecraft involving six or more persons, combination of data and voice intercommunication will be justifiable.

As an example of the application of this concept, it is often found that less than 5% of the telemetry data from a test program is of any value. Since telemetry is the biggest problem from a traffic standpoint, it is therefore the most subject to "editing" prior to transmission. This editing can be performed with the data conversion equipment in conjunction with the computation equipment. Such editing can be extended to video information, particularly pictorial data. The RCA *V*ariable *I*nstruction *C*omputer (VIC) concept presently being converted to qualified hardware, represents a reliable processor for this and other space vehicle data processing requirements. Unique to the VIC is the capability of reverting to less sophisticated modes of operation in the event of VIC or some other subsystem failure.

For deep space operations, the opportunity to use a combination of on-board and ground computation facilities to optimize the use of the radio links exists. The editing process can be supplemented by appropriate variations in transmission redundancy and bit rate to optimize the information transfer.

Another area where increased capability can be achieved without weight penalty is that of on-board information collection and distribution. In the traditional design, data for telemetry is hardwired (one circuit per data point) to the multiplexing equipment (PCM and/or FDM). Other data and voice circuits are similarly hardwired to computation and display facilities. For large spacecraft, the sheer weight of cabling can become on the order of hundreds of pounds. For digital data, the use of a number of contrally timed multiplexers materially reduces the cabling required. RCA work on multiple internal communications[1] extends this concept to permit the equivalent handling of a large variety of analog signals including voice.

Ground Status

Telecommunication performance improvement, particularly for deep space operations, can be achieved through improvement of surface RF and detection. RCA's work in masers and threshold extension will provide the basis for new equipments in the ground which will permit an order of magnitude increase in the data rates previously indicated.

Plasma Research

The RCA Victor Ltd. Research Laboratories in Canada have been engaged in the development of methods of modifying an ionized plasma to change its electromagnetic properties. This has been explored from the standpoint of using magnetic fields and material additions. The results have been promising and the feasibility of applying these methods to re-entry spacecraft is in process.

CONCLUSION

A few of the problems of space telecommunications have been reviewed and a few of the many programs in process at RCA have been mentioned. Some important areas of technology such as tape-recording, lasers, and re-entry plasma modifications have not been discussed. It should be evident, however, that RCA having already made a substantial contribution to spacecraft TT&C subsystems, is prepared to continue that contribution during the 1970's.

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THE MULTIPLE INTERNAL COMMUNICATIONS SYSTEM CONCEPT

Single Coaxial Cable Interconnects All System Components by Multiplexing

This system is multiple, because it deals with many forms of information (video, synchro, digital, and voice), and internal because it addresses itself to the total information transfer between all system components. It is the application of systems engineering and multiplexing techniques to the total information-transfer requirements of any ground, aircraft, or shipborne weapon system—an approach that allows one coaxial cable to handle all information transfer between all components of a system, made feasible by the use of integrated circuits in the multiplexing equipment. Discussed is the basic MICS technique and its applicability to a wide variety of communications systems.

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TRADITIONALLY, information transfer between components of a system has been accomplished by multiconductor cable and multipin connectors (Fig. 1). In addition to the cables and connectors, line drivers, and amplifiers, impedance matching and level converter circuitry are required to provide the proper interface. While the multiconductor cable approach was sufficiently effective for the less sophisticated systems of the past, the application of this approach to today's and future systems is impractical.

THE NEW CONCEPT

With the new multiple internal communications system (MICS) concept, one coaxial cable is used to transfer information between all the components of a system (Fig. 2). Components such as a radar or a computer inject or extract information through matched tees as required. The single-cable concept provides the distinct advantage of increased flexibility and expansion. Spare jacks can be provided during the initial installation at a nominal cost, to accommodate changes in layout or additional components for system expansion. This provides a roll-on, roll-off capability where system components can be changed or rearranged without expensive modifications to system cabling. The provision of spare jacks during initial installation can be compared to installing spare AC outlets throughout a house during construction. Installing such outlets after the house is completed can be quite expensive and in aircraft, the cost of modifying cabling is exorbitant. Fig. 3 shows the installation modification required in a modern F-series interceptor to accommodate the AN/ASW-21 Data Link equipment. This modification cost per aircraft was excessive on a production basis.

Two methods of implementing MICS are used. One, the adapter unit method, can be employed for existing systems (Fig. 4). Although it has demonstrated significant weight reduction, it is not the optimal approach. It only eliminates the hardwire between system components, while the multipin connectors and the driving circuitry are retained. The second method is the built-in approach (Fig. 5). This method can be used in new systems by specifying the new approach in the initial equipment procurement. Built-in MICS achieves greater weight savings through the elimination of the cable connectors, and provides increased system reliability through the elimination of redundant driving circuitry.

Final manuscript received December 2, 1966.
*Now with North American Aviation Co., Anaheim, Cal.

Fig. 1—A typical hard-wire interconnector system; inset is a modification required for the AN/ASW-21 Data-Link.

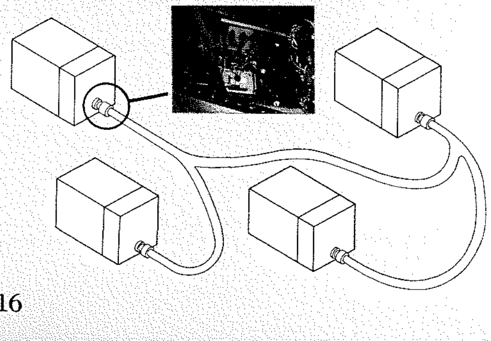


Fig. 2—MICS concept block diagram.

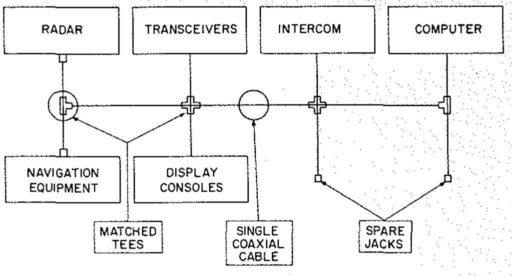
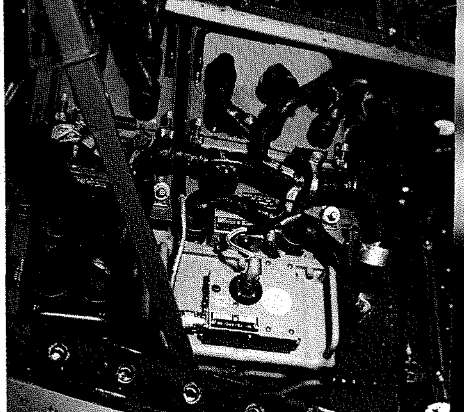


Fig. 3—AN/ASW-21 modification required a modern F series interceptor.



L. Barney D. Wade A. A. Paris L. L. Wolff

LYLE W. BARNEY received his BSEE degree in 1958 from Drexel Evening College, and is currently working toward a MSEM degree. Since joining RCA in 1953, Mr. Barney has held numerous assignments in design, development and test of military equipments. He has been associated with the MICS program since its inception in 1964 and is the Product and Systems Engineer on the MICS-A-NEW program. He has also been responsible for the in-house studies of the application of MICS to other systems. Mr. Barney has been Project Engineer for various programs including the PRESS Airborne Equipment program and programs associated with the Navy's AN/ASW-21 Two-Way Data Link program. He was Project Leader for the F-105B and F-105D Time Division Data Link Test at Eglin AFB in Florida. He was also responsible for preparing detailed test plans for the F-102 and F-106 Time Division Data Link test program at Bedford, Massachusetts.

LARRY L. WOLFF received his BSME degree in 1939 from Prague University, and spent the following three years as a junior engineer designing heavy machinery. He served in World War II as an Engineer and Intelligence Officer in Europe and the Pacific theatres of war. After the war, he was a technical advisor and interrogator at the Nuremberg Trials. After serving again in the Korean War, he joined the Burroughs Corporation Research Center in 1955 and became supervisor of technical publications. With RCA since 1959, he has been assigned to the Systems Support area and the Aerospace Communications Group of ACCD. In 1964 he was assigned to the Autodin PMO and at present is working in the Air/Surface Systems group of CSD. He has published extensively in the field of digital computers and holds one US patent.

DUANE K. WADE received his BSEE degree in 1947 from Ohio State University. Following two years as a junior engineer with the Bell Laboratories, he joined the Ford Motor Company and participated in the design, development and installation of the first application of automation to an industrial complex. Following this, he was a Senior Design Engineer with Convair, San Diego, responsible for weapon system integration for the F-106 and F-102 Interceptors, including the integration of the RCA TDDL. After leaving Convair, he joined the Hughes Aircraft Co., Culver City, as a Systems Engineer. There, he was responsible for Fire Control Systems design, development and their integration into various Weapon Systems. He is presently a leader in the Air/Surface Systems Group of CSD and is responsible for the Multiple Internal Communication System application to all systems.

ALLAN A. PARIS graduated from Brooklyn Polytechnic Institute with an MBA, and from Drexel Institute of Technology with an MSEE; he is a candidate for his PhD at the University of Pennsylvania. Upon joining RCA in 1951, he was assigned the design responsibility for ERIS (Electronic Range Indication System). From 1955 to 1957 Mr. Paris was assigned to the U.S. Army Signal Air Defense Engineering Agency. Since 1957, he was systems project leader on various Time Division Data Link programs (ARIES, F-101, F-102, F-106, and F-108) and later product line manager of an activity responsible for the research and development of numerous military communications projects. Mr. Paris is presently Manager of the Air/Surface Systems Group of CSD with responsibility for Sam "D", AWACS, MICS, LEM, PAN ORB, PRESS, TAC TDDL, AMSA, TACFIRE, and TACOM. He has been the corporate member of the RTCA SC100 Data Link Standards Committee and is currently on the SC100 Committee. He is active in the Military Operations Research Society, working group on Command and Control, and The American Society of Naval Engineers. He is a member of Eta Kappa-Nu and the IEEE.

MICS AND THE A-NEW SYSTEM

The Air-Surface Systems Group of the Communications Systems Division has for years recognized the possibilities of using multiplexing techniques to reduce aircraft cabling. However, until the advent of integrated circuits, the multiplexing equipment was too large and heavy for use in aircraft. In 1964, this group studied the feasibility of using multiplexing to reduce the cabling in the A-NEW system. The A-NEW system was selected because it was one of the largest and most complex airborne systems being developed by the Navy. In that year, the Air-Surface Group was awarded a study and feasibility model contract by the Naval Air Development Center (NADC), Johnsville, Pa., to apply the new concept to the A-NEW-MOD-2 concept. The A-NEW program has as its objective to increase the effectiveness of antisubmarine warfare (ASW) aircraft.

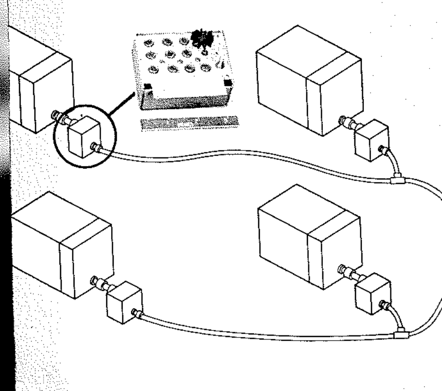


Fig. 4—The adapter approach to the MICS concept reduces weight of existing systems.

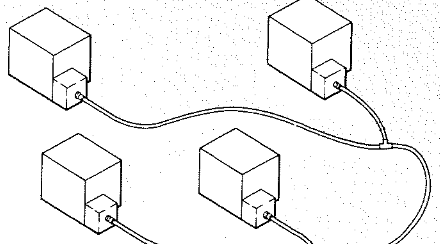


Fig. 5—New systems use the MICS built-in approach to decrease weight and increase system reliability; smaller boxes illustrate size reduction.

The basic concept which will improve the ASW capability centers on a digital computer which services multisensor CRT displays for observation and control of ASW information. The A-NEW-MOD-2 is a feasibility model of an airborne ASW data-processing and display system, designed to operate in the laboratory with a problem generator which furnishes ASW sensor and navigation system outputs required to simulate an operational ASW environment.

The study for the A-NEW-MOD-2 application was completed in January of 1965, and in June of 1965, NADC awarded a follow-on contract for an A-NEW-MOD-3 MICS study, now in progress. In September of 1965, the feasibility model was integrated and demonstrated with the A-NEW-MOD-2 equipment at the Naval Air Development Center, Johnsville, Pa.

Since the A-NEW design was well advanced, the adapter unit approach had to be used. Although this approach did not constitute the ideal method of applying the new technique, the study in-

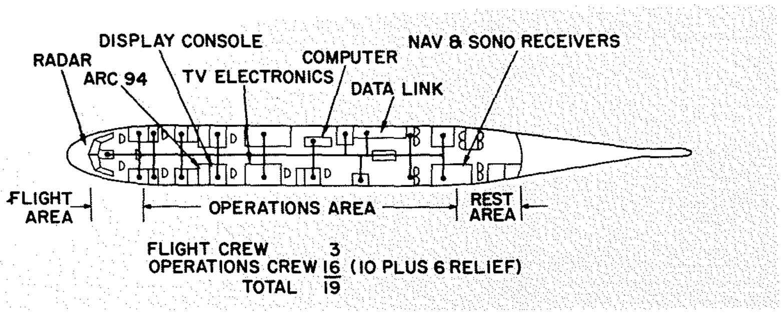


Fig. 6—The A-NEW system interior arrangement.

dicated that over 4,000 wires (total length of 74 miles) could be eliminated in the aircraft with a net weight reduction in excess of 1,000 pounds. The weight reduction provided a gross cost saving of $500,000 per aircraft (based on valuation of weight in this type of aircraft of $500 per pound). The net cost savings could actually exceed this amount, since such other factors as decreased system downtimes, fewer ECP's, lesser training requirements, and decreased documentation could not be evaluated at this time. The study also indicated growth potential through flexibility, and that integrated circuits can be used for the MICS construction to provide a highly reliable system.

In defining the operational elements of the A-NEW system, it was found that the system could be divided into 20 equipment groups or stations (Fig. 6), with an adapter for each of these stations. All intrastation transfer of data was accomplished over conventional wiring. Interstation communication was handled by the MICS, so that 20 MICS adapters and 34 coaxial cables eliminated over 4,000 wires.

In addition, RCA investigated the advantages the built-in MICS method would have had if used during the A-NEW implementation. It was determined that the MTBF of the A-NEW computer output section could have been increased by better than three-to-one through decreasing system complexity and the elimination of unnecessary redundancy. Table I shows a breakdown of how the increased reliability figure was obtained. The circuitry required for the new approach was determined and failure rates for these circuits were computed. This computation yielded an MTBF of 2,500 hours. The driving circuitry which was replaced by MICS included 420 line drivers and 1,840 connector pins. The MTBF of this circuitry was only about 720 hours.

TABLE 1—Reliability Comparison-MICS vs Hardwire: A-NEW computer output section only

Item	Quantity Required		Total Fail. Rate × 10^{-6}	
Basic Group Osc.	MICS	7	14.07	MTBF = 2520 Hrs.
Up Down Conv.		31	116.56	
Comb./Divider		15	36.75	
Basic Group Det.		57	228.50	
Connectors (BNC)		28	1.12	
			397.00	
Data Line Driver	HARD WIRE	420	1218.00	MTBF = 720 Hrs.
Control Line Driver		42	149.94	
Connector Pins		1848	18.48	
			1386.42	

The weight saving of this approach was found to have been far in excess of 1,000 pounds because of the elimination of cable connectors. A comparison between a 91-pin A-NEW hardwire connector and a MICS cable is shown in Fig. 7.

Because the advantages of MICS were so readily identifiable, the Office of Naval Research awarded a contract to RCA in April of 1966 to develop *Multiplexing Guidelines And Standards* for all data on avionics equipment in Naval aircraft.

NADC also recently awarded to RCA a contract to study the applications of MICS to the VFAX aircraft system, and the Marine Corps asked RCA to supply a MICS Data Multiplex System (AN/GCA-1) for use in Tactical Air Operations Center (TAOC) huts.

In the AN/GCA-1 Program, the multiplex system design will result in weight savings of about 10,000 lbs. (now about 13,000 lbs.), increase hut-to-hut distances by 100% (now about 175 feet) and reduction in set-up and take-down time by about 80%. This will be achieved by a combination of FDM-TDM techniques using a single coaxial cable as the transmission medium for radar, video, data, signalling, voice and other forms of information flows between the huts of a TAOC. An MTBF in excess of 5,000 hours and an availability rate of 0.9998 are the realizable design goal for this system.

In addition, the Air-Surface Systems Group has completed or is conducting studies for MICS incorporation into various navy aircraft such as the P3C, VSX, as well as applications to ships, such as the Fast Deployment Logistics Ship Program (FDL) and ASWICS ships.

OTHER APPLICATIONS

At present, additional in-house studies are in progress to apply MICS to several other complex systems such as WAC (Warning and Control), AWACS (Airborne Warning & Control System),

Fig. 7—A 91-pin hard-wire connector (now in use) compared to the MICS single cable (left).

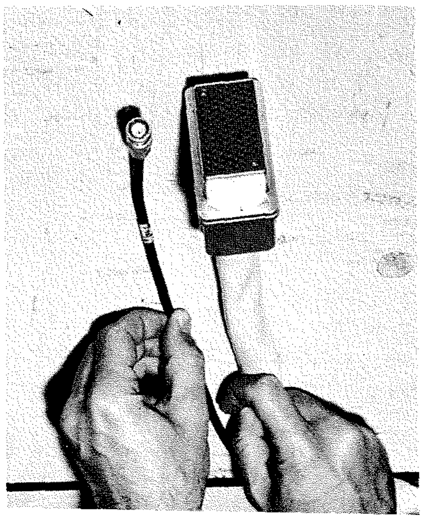
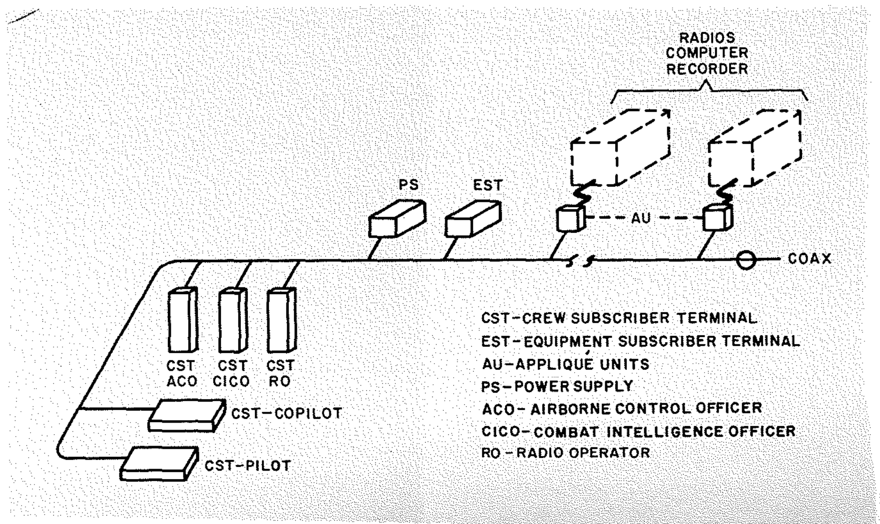


Fig. 8—E-2B ICCS Equipment Diagram.



CBATACS (Carrier Based-Airborne Tactical Aircraft Control System), LOH (Light Observation Helicopter) and ships. Its application to the WAC aircraft indicates that communication between the various system components can be accomplished in the same manner and with similar savings indicated for A-NEW. The study shows that the present WAC aircraft utilizes an in-flight performance monitor which requires 177 wires and 32 coaxial cables to interface with the system components. MICS is able to reduce this to three coaxial cables.

HELICOPTER APPLICATION

In an effort to see how the new technique could be applied to simpler systems, RCA investigated the possibility of using it in the light observation helicopter. The hardwire approach for the communication, navigation and instrumentation (CNI) interface requires 108 wires, a 48-terminal junction box, and 56-pin connectors on the intercom boxes. The intercom boxes themselves contain two integrated circuits and several switches on the front panel. The 56-pin connectors are the largest component on the intercom boxes. The MICS approach accomplishes this interface with two coaxial cables.

INTEGRATED COMMUNICATIONS CONTROL SYSTEM (ICCS)

Another similar system employing many of the MICS concepts for control, routing and transfer of audio and digital signals has been defined by Air-Surface System Engineering and is presently being designed and fabricated by the Light Military Equipment Section of CSD. This system has been designated the Integrated Communications Control System (ICCS). The ICCS is an application of MICS to the E-2B Aircraft. It provides for the control for all interior and exterior communications. An equipment diagram is shown in Fig. 8. The ICCS is

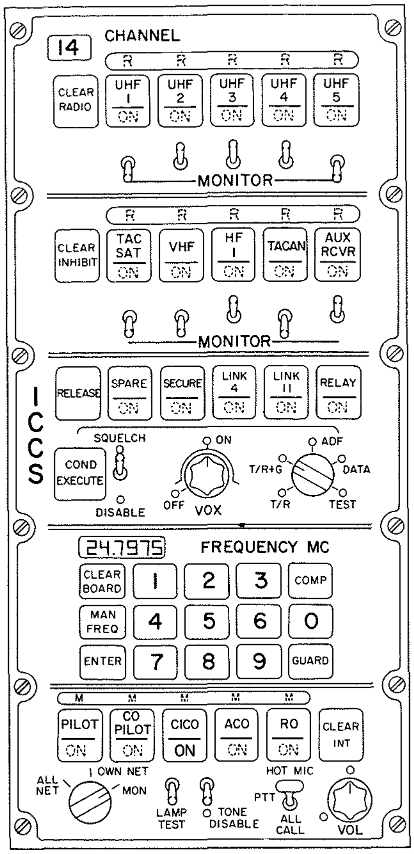


Fig. 9—E-2B ICCS Control panel.

used to interface five intercom stations, twelve radio sets, the Link 4 and 11 data modems, security devices, computer, recorder, emergency tone inputs and tactical satellite communications terminals. The system provides for the functions of net calls, party calls, radio access and control from any subscriber in the aircraft, multiple access to individual radios, and the monitoring capability of all receivers. The ICCS further provides for radio relay operations, transmission over multiple radios, initiation of radio self-tests, and maintenance communications. The control-panel layout is shown in Fig. 9.

Operation is accomplished by multiplexing signals over a common coaxial cable using TDM for signal and control functions and FDM for audio signals. A common timing and sync is used for inherent system stability.

Fig. 10 illustrates an integrated communications control system using a combination of centralized and decentralized circuit functions applied to a shipborne installation.

MICS APPLICATION TO SHIPBORNE RADAR SYSTEMS

The Air-Surface Systems Group is conducting a program which applies the MICS concept to U.S. Navy Surface Missile Ship Radar and fire control systems. The study has already shown that MICS is very advantageous in reducing the wiring problem encountered in building and modernizing these ships. The interface between the AN/SPS-48 Search Radar and the NTDS for instance reduces the requirement of 41 cables and about 600 wires to 6 coaxial cables through the application of MICS. A similar hardwire reduction from 28 cables and approximately 690 wires was made possible by applying MICS to the interface between the AN/SPG-55B Radar and the TERRIER Fire Control System. In both cases, additional benefits accrued through the MICS application in that the ship's system flexibility was greatly enhanced now by making redundant signal paths available between the forward fire control system and the aft fire control launcher. Furthermore, MICS made it possible to have a remote central location for monitoring, test, and checkout equipments. Other areas of study underway include the remoteing of internal equipment test points to a central monitoring point and the use of computer programs for shipboard performance analysis. The MICS approach to ships systems shows promise of realizing entirely new concepts in vessel construction, utility and employment, and opens up a brand new RCA product line.

Fig. 10—ICCS for shipboard use.

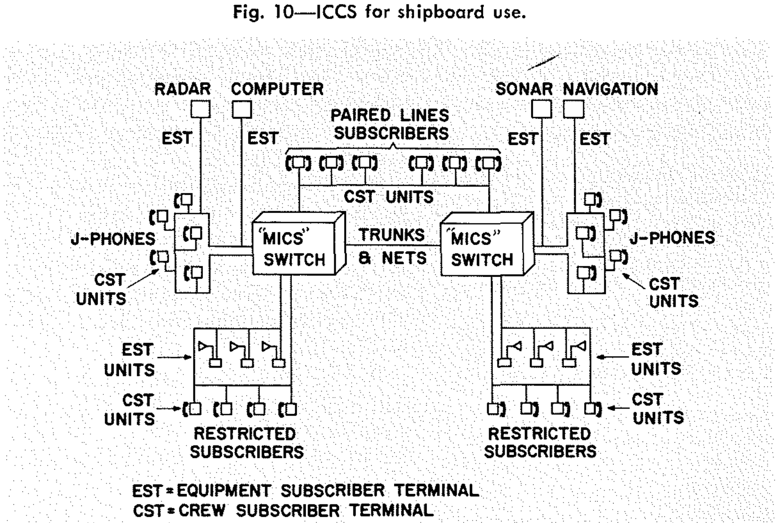
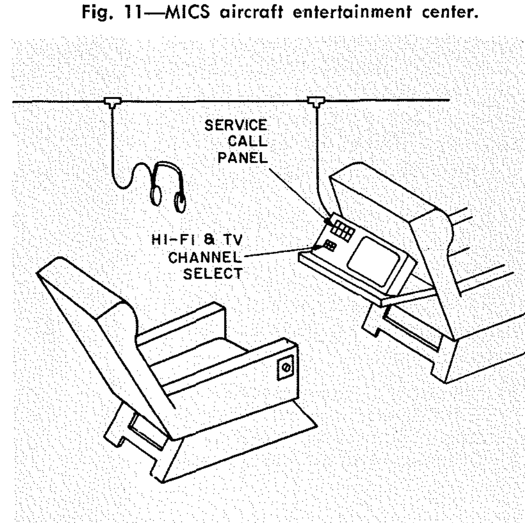


Fig. 11—MICS aircraft entertainment center.



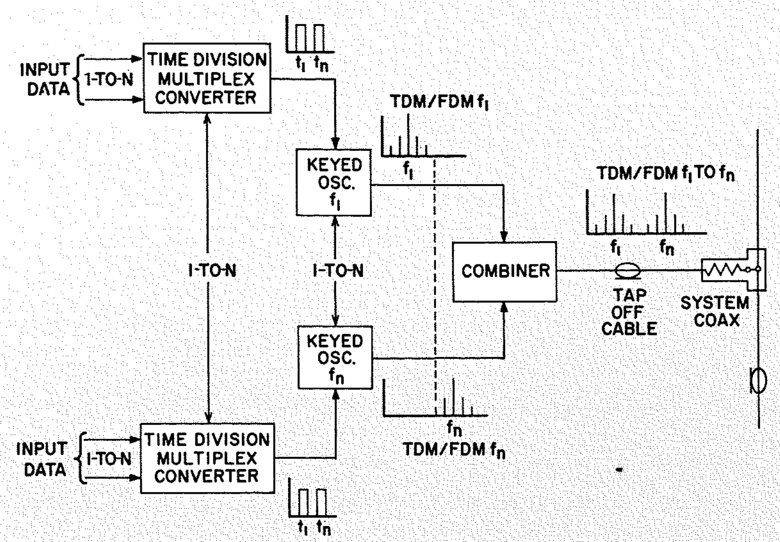


Fig. 12—MICS technique of multiplexing data over coaxial cable.

COOPERATION WITH OTHER RCA DIVISIONS

The work performed by the Air-Surface Systems group in the MICS area is being coordinated for timely application to products of other RCA Divisions. Some of these areas are the Variable Instruction Computer (VIC) of the Aerospace Systems Division in Burlington and checkout and monitoring products of the Aviation Equipment Department of the West Coast Division.

RCA is also looking into the possibility of using MICS in future 300- to 400-passenger commercial aircraft. Since it is the desire of the airlines to provide several types of aural and visual entertainment for their passengers during flight, this will lead to a severe cabling problem. However, the new approach will allow a complete entertainment center at each seat to be connected through one coaxial cable (Fig. 11). Furthermore, in the non-defense area, the MICS is being discussed with RCA Electronic Data Processing for possible use with data management systems and other commercial computer system applications, and with the RCA Home Instruments Division for eventual use in their products.

APPLICABILITY CONSIDERATIONS

The first step in determining whether the MICS approach is a proper vehicle for the implementation in a system is an investigation which is conducted in five basic steps:

1) Definition of all operational elements of the system, in terms of information inputs and outputs, and function.
2) Definition of all information paths and flows. By this is meant which system component communicates with what other units, and by what route.
3) Selection of multiplexing techniques, such as FDM, TDM, combination FDM-TDM and others.
4) Trade-off studies, in terms of cost and complexity between the various multiplexing techniques and the hardwire approach. The output of the trade-off studies is a system definition.
5) The final step in the MICS study cycle defines the system interface requirements, and provides the equipment procurement specifications.

The value of MICS is not always apparent at first. Furthermore, the application criteria differ from one application to the next. The most obvious advantage is the reduction in the amount of copper wire interconnecting the components of a system. The A-NEW study, which was discussed above, shows that 74 miles of wire weighing 1,000 lbs could be eliminated from the A-NEW system. However, for surface systems where weight reduction is less important, other advantages accrue. Currently, the U.S. Navy is examining MICS for use in the fleet.

The reason for this interest is obvious when the following problem areas which exist at present are considered:

1) *Complexity*—The complexity of the interconnecting wiring for a large system poses a serious problem to installation personnel. Furthermore, the design documentation, installation, and upkeep of complex wiring systems is costly.
2) *Flexibility*—Modifications to existing systems are often impossible because of the excessive time and cost of the retrofit.
3) *Maintenance* — Performance monitoring and checkout capabilities are required of nearly all existing or newly conceived weapon systems. Although monitoring and checkout philosophies vary from one system to the next, all have a common need for signal acquisition. The limiting factor is often the number of wires which can be tolerated.
4) *Reliability*—Reliable operation is always the paramount factor for any system and complexity often reduces this all-important requirement. The A-NEW study has shown that the new technique often replaces not only wires but the interface circuitry required to drive these wires. This, coupled with the fact that the reliability of integrated circuits used with MICS is improving faster than the reliability of wire connection, makes MICS a natural candidate for improving system reliability.

DESIGN TECHNIQUES

One of the basic design techniques employed in the application of MICS is to convert as many signals as possible to a single, high-speed time-division multiplexed (TDM) pulse train which in turn keys an oscillator. The outputs of several keyed oscillators (each operating at a discrete frequency) are then combined and inserted for transmission over the system coaxial cable. The net result is a combination TDM-FDM system. Fig. 12 illustrates this basic technique. Data extraction is the reverse process. Since all transferred information is present on a single coaxial cable, data extraction can be accomplished at any point(s) along the cable by any number of units.

Component size reduction has been one of several factors in making the new concept practical. Fig. 13 shows a breadboard of a 10-channel keyed oscillator group 6 x 9 x 1½ inches which was developed for the A-NEW feasibility model in 1965. This circuit has been breadboarded in 1966 in integrated circuit form and measures 3 x 2½ x ½ inches.

Latest developments in integrated circuits have indicated that significant size reductions will be made in future MICS systems. Fig. 14 shows the expected sizes of basic circuit components for the 1970 era.

All terminals of the MICS transmit and receive over a single coaxial distribution system. A combiner-divider within each terminal separates the carriers. An ideal distribution system would have equal loss between any two stations and can be represented by the diagram of Fig. 15.

An in-line configuration is more desirable and the non-equal cable loss between stations is not a problem for the MICS frequency spectrum for relative short coax runs of several hundred feet. The in-line configuration is illustrated in Fig. 16.

In Fig. 16, the loss from point *1* to *n* will be greater than the loss from *1* to *2*; however, considering the use of RC-59 for frequencies up to 200 MHz, the difference in loss will be approximately 6 dB. When cable runs become excessive the use of larger coax, variable attenuation taps and dynamic range of AGC can provide additional compensation. For long cable runs, the loss as a function of frequency must be considered. However, variable attenuation or AGC techniques can compensate for these unequal losses if they become critical.

Community Television Antenna Systems which employ a single coaxial cable to feed up to 1,000 television sets, maintain the power level between channels (channels 2 through 13, 54 to 213 MHz) to within 2 dB. Since the wavelength of the frequencies used are in the order of meters, the main cable is electrically "long," the VSWR must be considered from a transmission power viewpoint, and the main cable must be terminated in its characteristic impedance.

While the main coaxial line is electrically "long" the coax tap offs to the combiner-dividers are electrically "short." The bridging resistors of each terminal present a high impedance relative to the characteristic impedance of the main coaxial trunk. Energy propagated on the main coax is thus absorbed by the main cable terminations and the only reflections or echoes are produced by mismatch introduced by the bridging resistors.

Since the terminals are lightly coupled into the line to reduce the effects of reflections, the coupling will represent a relative high attenuation to the transmitter signal. This attenuation is adequately compensated for by the power of the transmitter and gain provided by amplification in the receiving terminals.

SUMMARY

The use of the Multiple Internal Communication System provides the following advantages:

1) It will permit the transmission of all types of information over a single transmission path.
2) It provides low installation and documentation cost because considerably fewer cables are required.
3) It permits flexibility for redesign or layout change.
4) It makes possible increases in system reliability, and reduces volume and weight in critical applications.

Although much remains to be accomplished to make the new concept applicable to future complex and sophisticated electronic systems, the groundwork has now been laid. Its advantages have been demonstrated in practice and thereby have made a very significant contribution to the use of total integrated circuit systems. In the future, multipin connectors with all their inherent weaknesses may become as outdated as vacuum tubes are today for general use in electronics.

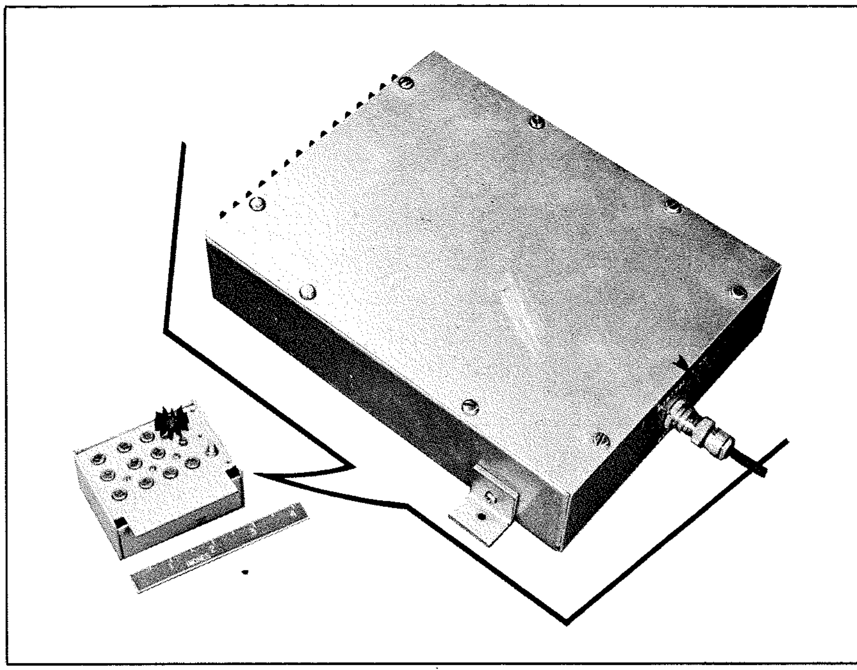


Fig. 13—Two breadboards of a 10-channel keyed oscillator group developed for A-NEW (note 3-to-1 size reduction of integrated circuit version at left.)

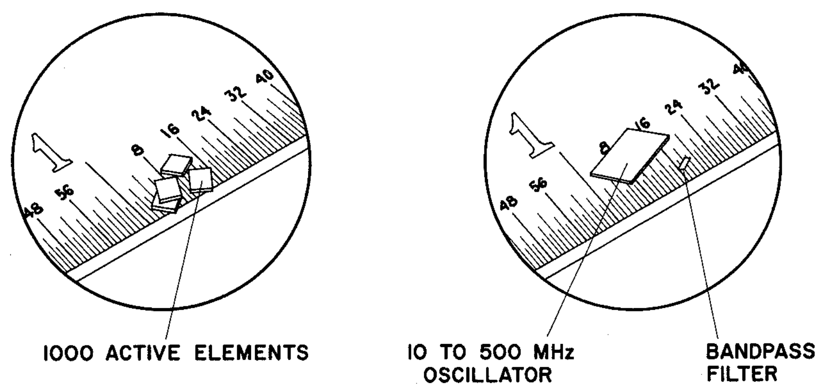


Fig. 14—Component circuit sizes (1970 era).

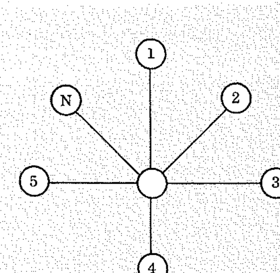


Fig. 15—Representation of ideal distribution system.

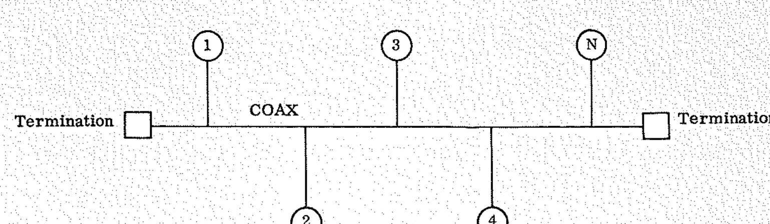


Fig. 16—In-line configuration.

TRENDS IN THE PACKAGING OF MILITARY ELECTRONICS EQUIPMENT

This paper examines the requirements and trends of packaging electronic equipment for military use. The role of integrated circuits, the mechanical design of module boards, interconnectors and packages, and a discussion of the problems of nuclear radiation effects and heat dissipation are treated. Advanced trends in power supply design and cooling techniques are described.

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WILLIAM BLACKMAN graduated from New York University with a BS in Industrial Engineering in 1956. He joined RCA that year and has participated in design and development assignments on the miniaturization and general packaging of combat radios, including production design studies. He designed the control panel and associated test equipment for the Air/Ground Automatic Control System (AGACS). For the past six years he has been working as a mechanical systems engineer on the mechanical systems design of the Minuteman missile ground support equipment. Mr. Blackman was discharged in 1953 from the U.S. Army Signal Corps after service as a radio communications crew chief and later as a commissioned officer in the U.S. and Korea.

No area of military equipment design has been left untouched in the scramble for increased packaging density by the application of integrated circuits. Major concentrations of effort have existed in airborne, missile, and man-pack communications equipment. Integrated circuit techniques have caused the present clamor for new design and packaging techniques. Engineering groups realize that microminiaturization is still in its relative infancy, and the present integrated circuits of two-, three-, or four-element functions will be giving way to complete arrays of registers, demodulators, encoders, and decoders. Relatively new devices such as metal oxide semiconductor (MOS) and evaporated thin-film circuits promise to produce a thousand active and passive components on a single chip.

The impact of these devices is of greatest concern to mechanical engineers, both directly and indirectly concerned with the development of packaging techniques for module, equipment, and system configurations. Military and space electronics make harsh demands on ingenuity in the application of both present and beyond-the-state-of-the-art designs. We must look forward to entirely new families of control and communications equipment.

A few of the areas of most critical interest to the mechanical engineer are: 1) interconnections, 2) module board design, 3) connector selection, and 4) environmental limits.

At the present time, the environmental constraints specified by shock, vibration, and moisture and thermal limits appear to be the areas that can be handled with relatively straightforward analysis. The only difficulty here is that meeting the first three requirements usually leaves little time for the latter, and only foresight and careful planning can produce the results necessary to meet the specifications.

Final manuscript received November 16, 1966.

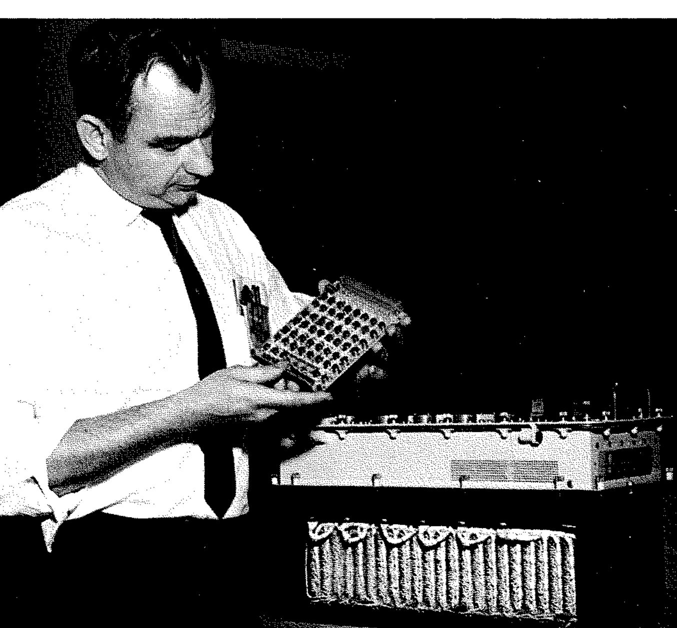


Fig. 1—C. R. Sieben, M.E. holds a printed-circuit board from the minuteman test set (data analysis central AN-GSM-147). This portable tester is used in the minuteman missile silos and replaces a four-unit tester now used. The board contains integrated circuits and uses the tuning fork and blade technique for connection to the chassis.

CONNECTOR SELECTION

Connector selection remains unique to the equipment design requirements and although standardization has been attempted (successfully in some cases) the final selection must be made by the design group. The types of connectors available fill the pages of technical papers and journals. Their general categories (see Figs. 1, 2, 3, and 4) are:

1) *Tuning Fork & Blade:* Applied by RCA in missile communication, control, and test equipment (ground based MINUTEMAN plus airborne data link systems e.g., (F104 Data Link). One of the most versatile of connector design techniques available due to the ease by which new design may be fabricated (by the use of metal header assemblies) for application in a quick reaction type facility with molded final form parts developed and supplied at a later date with no change in form factors or reliability predictions.
2) *Wirewrap:* Introduced to RCA by the efforts of the DEP Central Engineering and MINUTEMAN design engineers in 1959.
3) *Pin and Socket:* Probably the most extensively used and with the greatest amount of reliability and design documentation. Almost exclusively used in the combat radio design area with a high degree of confidence and success.
4) *Card Edge:* Still a controversial technique at RCA with little specific application to final design. It is, however, extensively used by U.S. Naval Air Development on anti-submarine-warfare systems.

MODULE AND WIRING TECHNIQUES

Closely allied with connector design are the module board and wiring techniques required to achieve the miniaturization goals of packaging. As long as there is more than one mechanical packaging design engineer, there will be a variety of approaches to the module and wiring techniques (Fig. 5) as follows:

1) Multilayer printed wiring with plated feed-through holes
2) Multilayer printed wiring with solid built-up layers
3) Welded wire matrix
4) Two-sided printed boards with plated feed-through holes
5) Wired boards (soldered or wire-wrapped)

Within RCA, the first of these techniques is by far the most popular. DEP Central Engineering has provided techniques and design guides[1,2,3] of careful study, test and analysis for the preparation of module configurations.

In prognosticating packaging techniques for the near future, we will find the same problems challenging the ingenuity of the mechanical design engineer: packaging density, heat transfer and environmental requirements. Reduced cost in the development of integrated circuit devices will lead to the procurement of functional arrays rather than specific circuits in integrated circuit configurations. This should alleviate the wiring density problems that generally restrict the multilayer board wiring requirements (Fig. 5). Since the practical limit of printed lines and spaces is presently being approached, we can expect to see wiring density increased by a judicious marriage of the best features of the *multilayer plated-through hole* technique with the *solid-layer built-up board* configuration.

RADIATION EFFECTS

Potentially, one of the most difficult problems to be faced is packaging for maximum resistance to nuclear radiation. This is particularly true of airborne equipment where space and weight are at a premium and where massive shielding which would be employed at a ground location would be intolerable.

The two classes of equipment in which this problem is of particular significance are satellites and space vehicles which will be exposed to space radiation over long periods of time, and military weapons systems which must survive exposure to radiation produced by a nuclear explosion. Of these, the weapon systems requirements are more severe. For example, it is highly conceivable that future guided missile electronic systems will be required to function properly after passing through the high radiation levels present in the fireball produced by an anti-missile nuclear weapon.

In designing for radiation resistance, four basic phenomena must be considered. These occur essentially simultaneously and tend to reinforce each other, making the total effect significantly worse than if the four occurred independently.

Gamma Radiation

Gamma Rays are extremely-short-wavelength, high-energy electromagnetic radiations originating from atomic nuclei. In general, the most serious gamma effects occur in semiconductor devices. Short gamma pulses result in ionization of the material and produce spurious currents which can temporarily disrupt circuit operations. Long-term, high-level exposure tends to produce permanent degradation of semiconductor devices.

Neutrons

Neutrons are heavy, electrically neutral atomic particles produced by fissioning of an atomic nucleus. These particles produce permanent degradation of semiconductor devices, by changing the characteristics of the material through atomic displacement.

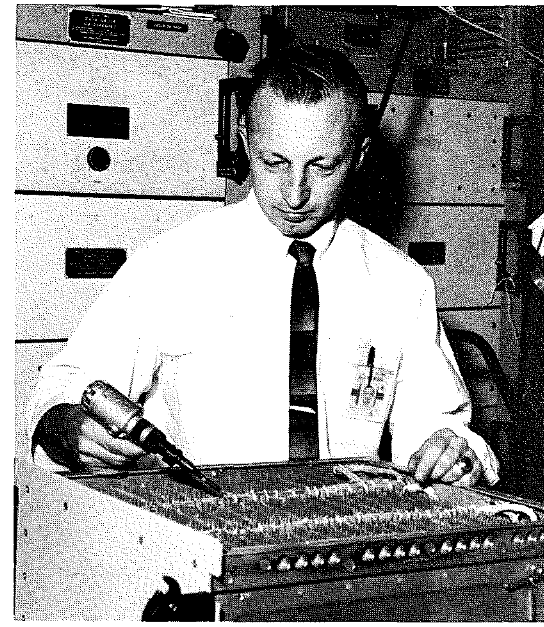


Fig. 2—R. L. Mangels, M.E. holds wire-wrap gun to a drawer from the minuteman digital data group. This minuteman ground support equipment uses the wire-wrap wiring technique extensively.

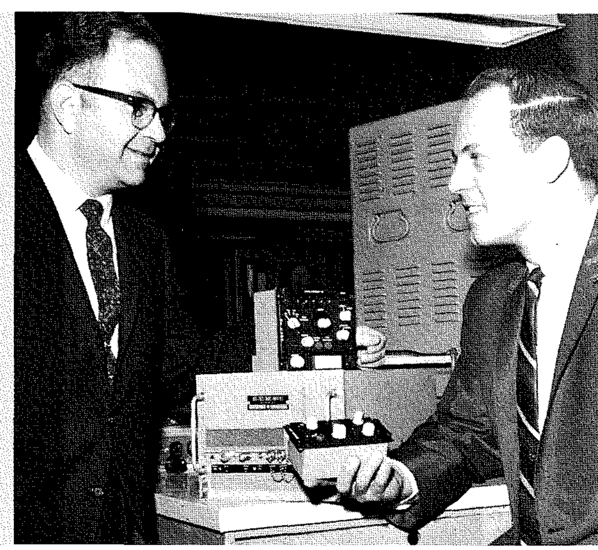


Fig. 3—With the ARC-104 are W. A. Brill, M.E. and Group Leader, (left) and E. M. Morse, Lead Mechanical Engineer. The ARC-104 is an advanced airborne transceiver built for the U.S. Navy. Other mechanical engineers on the ARC-104 project were M. R. Alexy, E. J. Burns, W. E. Comfort, M. Holbreich, L. H. Maher, E. Weber, I. Woolf and H. Sokolov. The tuning fork/blade and the pin/socket interconnection techniques were used.

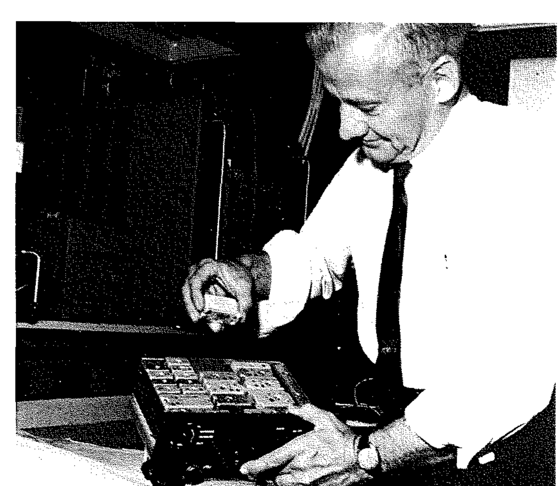
X-Rays

X-Rays are extremely short-wavelength, high-energy electromagnetic radiation originating from the orbital electrons adjacent to the atomic nucleus. X-Rays can produce ionization effects similar to those resulting from gamma radiation. In addition, the energy spectrum of certain X-Rays is such that the rays can be absorbed rapidly by materials having a high absorption coefficient and destruction of the material results.

Electromagnetic Pulses

Ionization of the atmosphere during a nuclear blast produces a high level electromagnetic pulse having characteristics

Fig. 4—J. L. Slivinski, M.E. pulls module from PRC-25 man-pack transceiver. PRC-25 is the "work horse" combat radio used by the army and marines in Viet Nam. The PRC-25 illustrates the pin-and-socket connection mode.



similar to a lightning discharge. This pulse can be picked up by an antenna system or directly penetrate an enclosure to cause burn-out of electronic circuits.

In designing for high-level radiation resistance, a combination of exotic shielding techniques and imaginative circuit design is required. Typically, a multilayer shield (i.e., tantalum, beryllium and aluminum) is used to provide partial attenuation of radiation, with the circuits designed to allow for anticipated degradation and transients resulting from radiation that will penetrate the shield. To achieve an optimum balance between packaging protection and circuit design, a close tie-in between the mechanical and electrical engineers must be maintained throughout all phases of the design program.

TRENDS IN THERMAL DESIGN

It is logical to expect that thermal requirements for densely packaged microminiature electronic equipment will result in the need for smaller heat exchangers and cooler units. Power amplifiers, particularly in the 1-to-20-kW CW range, usually employ tubes requiring liquid cooling. The cooling units may occupy as much as 25% of the total volume required for the entire power-amplifier package.

In the never-ending search for inventive means of thermal dissipation, an advanced state-of-the-art thermal conductor called a *heat pipe* has been developed by RCA Electronic Components and Devices at Lancaster, Pa. ECD is marketing such a heat pipe operating on an evaporation-condensation cycle which can transfer heat very efficiently for any one of many selected temperatures depending upon the working medium (Fig. 6). This heat pipe can concentrate or diffuse heat flux depending on the application. The thermal conductivity of this device may be as high as 10,000 times that of a comparable copper tube.

The heat pipe consists of an evacuated enclosure with a capillary structure, or wick, along its inside surfaces. It contains a small amount of fluid which saturates the capillaries and has a substantial vapor pressure at the desired operating temperature. The fluid is evaporated at the heat input end, condensed at the heat output end, and returned to the evaporator (through the wick) by capillary action.

An important consideration in forced-convection cooling is the pressure drop required to force the air or water through ducts or pipes and over the parts or cases to be cooled. A blower or pump must furnish this energy, which, in turn, is usually supplied by an electric motor. The power required by this motor to force the coolant through the passageways varies directly with friction or pressure drop. High rates of heat transfer are due to high velocities which require relatively large cooling power requirements.

A heat pipe may be used to good advantage to reduce the power requirements. In a klystron application, for example, the heat pipe can transfer the heat from the dense "collector slug" area to a less dense "low pressure radiator" heat sink. A smaller fan or pump, with lower pressure requirements can then be utilized to remove the required heat. It may also be possible, in many instances, to dissipate the heat by natural means without the need for any rotating machinery.

There is no doubt that the heat pipe will see usage, in the future, in many diversified thermal applications, especially in the electronics industry.

ADVANCED COOLING TECHNIQUES

Packaged electronic circuits, subassemblies and systems can now be completely immersed and efficiently cooled in high dielectric strength (40-kV) liquids that have recently been developed. These fluorocarbon liquids cool electronics by means of evaporation of the coolant at a constant temperature. The coolant vapor enters a condenser and returns in the liquid form to repeat the evaporative cooling process at the boiling point of the particular fluorocarbon chosen. Thus, unwanted heat emitted by packaged electronics is quickly and efficiently removed.

Fluorocarbon coolants are fully fluorinated, inert liquids with boiling points in the range of 122°F to 345°F. The electrical insulating and dielectric characteristics of these coolants are excellent. This is due to their completely fluorinated structure, which also accounts for their inertness and unmatched compatibility with other electronic materials and parts.

Use of the new fluorocarbon coolants[4] permits greater miniaturization of immersed electronic packages because these inert liquids have the capability to more efficiently remove heat and, at the same time, to supply the required electrical insulation. Efficient removal of heat permits the use of much higher component and circuit densities, thus leading to the design of lighter, smaller and more compact electronic packaging. Fluorocarbon coolants are now being used in many military, aerospace, industrial and commercial applications. Some applications are proprietary and/or classified.

Care should be taken in selecting the proper fluorocarbon coolant for a particular package cooling application. To obtain maximum reliability, one should try and select a fluorocarbon liquid with the lowest boiling point consistent with the outside temperature environment which, in turn, determines the condenser temperature.

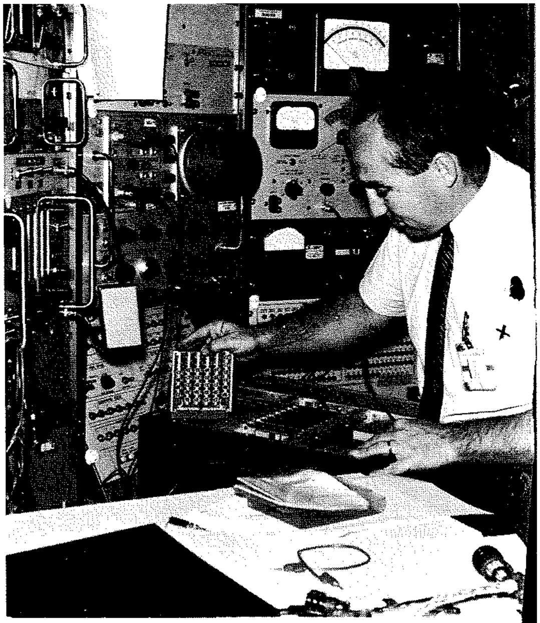


Fig. 5—M. W. Schmutz, M.E. holds integrated circuit board from the waveform inverter drawer, part of the Minuteman ground support equipment. The drawer is undergoing tests via a testing console. The self-test board shown embodies the multilayer board packaging technique.

POWER SUPPLY TRENDS

One of the major factors contributing to the size and weight of communications equipment is the power supplies required. However, recent advances in high-temperature core and insulating materials used in the iron core components, combined with fluorochemical cooling techniques, have greatly reduced the size and weight requirements of power supplies. The near future requirements will stress not only further reductions in size and weight, but also the ability to operate from any available primary power source.

Solid-State Inverters

The use of solid-state inverter power supplies promises solutions to the problems associated with these requirements. Advances in producing high-voltage transistors, capable of switching moderately high currents, have moved the inverter power supply out of the "hundreds of watts" region into the realm of "tens of kilowatts," with efficiencies in the order of 80%.

An inverter power supply capable of supplying 2.5 kW at a voltage of 4.4 kV-DC was recently breadboarded in CSD (Fig. 7). Preliminary tests indicate an efficiency of more than 80%. With convection cooling, this supply can be packaged within 0.5 cubic feet, at a weight of

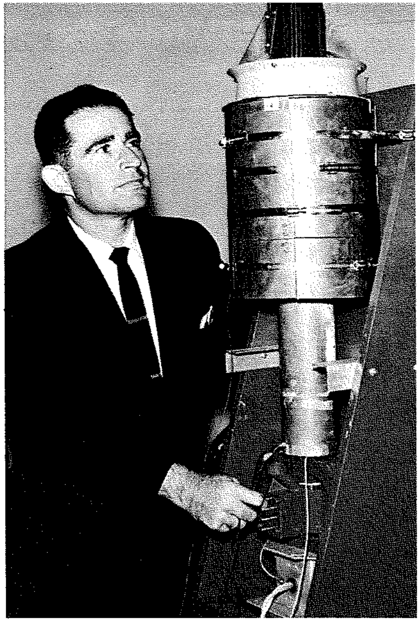


Fig. 6—G. Brauer, M.E. with heat pipe being used for high-power klystron cooling; his hand is on the klystron control, the rest of the upper portion is the heat pipe and its forced-air-cooled radiator.

less than 30 pounds. The primary power requirement for this supply is 208 volts-AC, three phase, at any commonly available line frequency, or 250 volts-DC. Using the same techniques, and currently available transistors, this power supply can be made to deliver in excess of 10 kW. It is expected that future transistor development will enable inverter power supplies of virtually unlimited power capacity to be built.

Space Equipment Power-Supply Packaging

In equipment which operates from limited energy sources such as batteries, a need has developed for highly efficient power supplies and regulators. Although, in most cases the primary interest is in conserving battery life, a further important advantage is gained in packaging.

Historically, power supplies have been inefficient. This was especially true in regulated supplies; as the source regulation became poorer, more power was dissipated in the form of heat.

The newer power circuits, however, particularly those of modest power employed in space equipment, and manpacks are very efficient. This means that in a power supply delivering 100 watts or so to a load, the power dissipated internally as heat might be only 5 or 10 watts. This eliminates in many cases, the need for elaborate heat exchangers—particularly blowers. Usually what heat is generated can be radiated or conducted away.

This reduction in heat allows such circuits to be packaged much more densely without high temperatures and the inherently low reliability.

High-Frequency Operation of Power Supplies

Operating the power supplies at higher frequencies also allows smaller packages. New devices and techniques are allowing significant increases in the frequency of operating DC-DC converters and of efficient switching of power regulators. Frequencies approaching 100 kHz are becoming more and more practical. The size and weight of magnetic components such as power transformers and chokes, as well as capacitors, etc., are dramatically reduced at these frequencies. An interesting result of this high-frequency operating technique is the elimination of the bulky 50-to-60-Hz transformers in small lightweight equipment which must operate from commercial and military power lines. This approach rectifies the 115V-AC input with a bridge rectifier. Typically this provides a 175V-DC which is fed to a DC-DC converter operating at a frequency much higher than the original input—for instance, 10 kHz. A transformer is still used to achieve the voltage changing, but in this case the transformer is a toroidal unit operating at 10 kHz instead of a "chunk" of iron at 50 Hz. This might result in a reduction in size and weight of 75 to 1.

These new developments have also had other more subtle effects. In some power supply devices overload protection is inherent, thus eliminating fuses and circuit breakers. Useful life of batteries is extended as better utilization of terminal voltage—both high and low, is feasible. Noise and ripple, although still present, are up only at inaudible frequencies, often beyond the response of the equipment in which they are employed (Fig. 8).

THE FUTURE

The future holds many new challenges for all concerned with military electronics. It behooves us all to keep abreast of the changing technology and maintain active internal communications between design groups; for, as our knowledge grows so will our participation in the military systems market.

ACKNOWLEDGEMENTS

The author is indebted to the following contributors to this paper: E. Van Keuren, G. Brauer, D. C. Bussard, R. P. Kisko, J. R. Hendrickson and C. W. Fields. Mr. Hendrickson also wrote the DEP Central Engineering Standardizing Notice No. 56-91-201 which describes the use of fluorochemical coolants for electronic equipment (RCA private communication).

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2. *I-C Packaging Evaluations' Format*, E. Szukalski (RCA); private communication.
3. *Multilayer Printed Wiring Study*, J. A. Dodd (RCA); private communication.
4. *Standardizing Notice No. 56-91-201*, DEP Central Engineering, RCA, Camden, N.J.

Fig. 7—D. C. Bussard with the inverter power supply for a 2-kW amplifier. The inverter shown is merely a breadboard; its ultimate size will be two-thirds smaller.

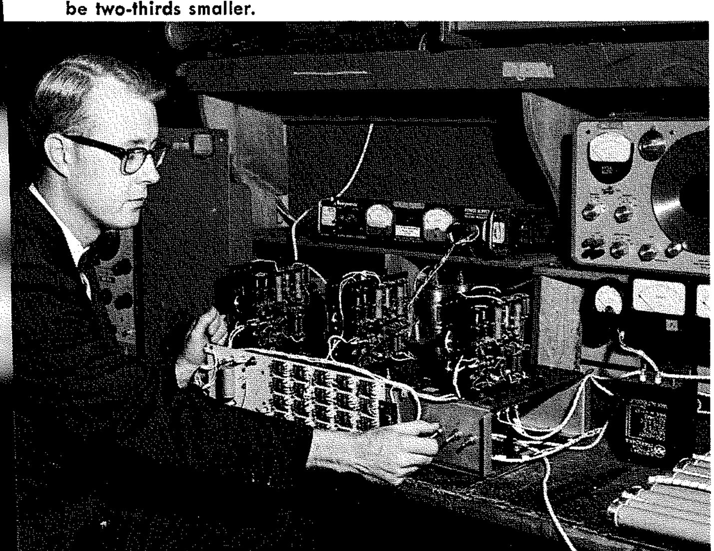

Fig. 8—CSD'S Power Supply Skill Center with some of the devices which they have supplied to both military and commercial design groups. Left to right are R. P. Kisko, Leader; N. Steinberg, M.E.; E. Lachocki, E.E.; J. E. Blume, E.E. Representative equipments are the PRC-62 Power Pack in front of Mr. Lachocki and the power regulator for the TR-70 Video Recorder, at Mr. Blume's left.

DIGITAL FREQUENCY SYNTHESIS

Particular tasks in the ADCOM-70 advanced technology program are directed to the use of digital techniques for transceiver functions heretofore implemented with analog circuits. Techniques considered impractical before must be re-examined in the light of recent and anticipated developments in digital integrated circuits. The synthesizer described here is a pertinent illustration of radical improvements achievable in transceiver design through the use of digital techniques. The unit is intended for use in UHF airborne transceivers. Only one reference crystal oscillator is required to provide 3,500 channels at 50 kHz increments to the requisite accuracy.

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FREQUENCY synthesis is a determining factor in the ultimate complexity and cost of communications equipment. Most frequency synthesizers have consisted of multiple crystal oscillators, harmonic generators, mixers, and filters utilized to combine several frequencies to provide the one desired output frequency. While the above techniques have been utilized to provide excellent performance in communications equipment, the increased emphasis on small size, high reliability, and maximum utilization of microcircuit techniques has spurred efforts to improve and simplify the frequency synthesis function. With the advent of low-cost, high-speed digital microcircuits, digital frequency synthesis has become practical, and promises order of magnitude improvements in size and reliability of the frequency synthesizer; the concept of digital frequency synthesis is, of course, not of recent origin and has been implemented with vacuum tube circuits as early as 1947.[1]

BASIC SYNTHESIZER COMPONENTS

Fig. 1 shows the functional block diagram of the digital frequency synthesizer. The VCO is maintained at the desired output frequency by means of a phase locked loop. The VCO output is divided by a divide-by-N counter. An integral number, $N_1, N_2 \cdots$ is associated with each desired VCO frequency. The division by N results in a frequency F_N which is compared with a reference frequency F_R derived from a precision crystal oscillator. If the VCO oscillates at the correct frequency, then $F_N = F_R$. If the VCO tends to drift, then $F_N \neq F_R$ and the phase comparator produces a correction voltage to pull in VCO to the correct frequency.

The attractive features of the basic digital frequency synthesis approach are therefore evident:

1) Only one stable crystal reference is required. The output frequency accuracy (long-term stability) is determined by the above crystal.
2) Minimal use of tuned circuits. The VCO is the only RF function in the system.
3) Maximum use of digital microcircuit techniques for reduced size and increased reliability.
4) The synthesizer may be remotely programmed by the application of DC logic levels.

UHF DIGITAL FREQUENCY SYNTHESIZER

A particular application will serve to illustrate the choice of basic parameters for a digital frequency synthesizer.

Fig. 2 shows the block diagram of a UHF digital frequency synthesizer.[2] The range of the divide-by-N counter is determined by the output frequencies and channel spacing. Thus:

$$N_{max} = \frac{399.95 \text{ MHz}}{50 \text{ kHz}} = 7{,}999$$

$$N_{min} = \frac{225.00 \text{ MHz}}{50 \text{ kHz}} = 4{,}500$$

The highest F_R should be utilized to obtain the highest correction rate and to reduce pull in time of the VCO. The upper limit of F_R is determined by the highest practical operating frequency of the divide-by-N counter. Since logical operations at UHF frequencies with monolithic circuits are at present not feasible, the variable counter must be preceded by a fixed divider (divide-by-K). From a circuit standpoint, it is convenient that K be a power of two. Thus, for $K = 32$, the maximum operating frequency of the divide-by-N is approximately 12.5 MHz.

Important considerations in the choice of the crystal oscillator frequency are oscillator stability requirements and the relationship of oscillator harmonics and the IF frequencies. A frequency of 5.3 MHz is near-optimum for long-term stability and ease of temperature compensation. The only significant harmonics close to a 60-MHz IF are the 11th 58.4 MHz) and the 12th (63.7 MHz). These are sufficiently attenuated and removed from the information pass-band.

Three VCO's are used to cover the output frequency range in 60-MHz increments. (Three VCO's are sufficient, since both high- and low-side injection are used in the *receive* mode.) Thus, the 60-MHz offset required when switching from the *transmit* to *receive* injection mode is conveniently provided. Fig. 3 shows an experimental UHF hybrid, thin-film VCO. The fixed-divider utilizes tunnel diode flip-flops for the first four stages and operates over the input octave bandwidth without requiring tuning.

Fig. 4 shows the developmental model of the synthesizer digital counters. The depicted model is essentially a breadboard. Point-to-point wiring was used for circuit interconnections. Therefore, the number of flatpacks accommodated by a board was limited.

Fig. 5 illustrates the basic operation of the digital phase comparator. A voltage ramp is generated at the reference frequency rate. When the system is phase locked, a fixed phase difference will exist between the divide-by-N and divide-by-R outputs. The divide-by-N pulse momen-

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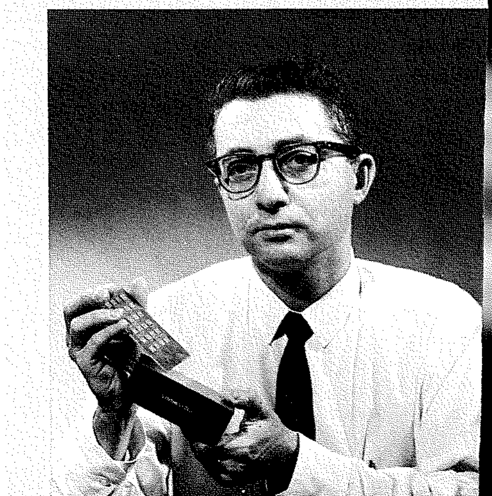

E. D. Menkes holds mock-ups of the digital synthesizer and a multilayer board which performs all the counter logic; modules in foreground are a frequency synthesizer used in current military aircraft.

Final manuscript received November 11, 1966.

tarily closes the sampling switch, charging C_H to the voltage level of the ramp existing at the sampling instant. Thus, the phase difference has been translated to a proportional DC level required to position the VCO. By the use of a field effect transistor for the sampling switch, an extremely high ratio of *off* to *on* impedance is obtained, resulting in minimal discharge of C_H when the sampling switch is open. Similarly, the input stage of the buffer is a field effect transistor (FET) providing the large discharge time constant. It is important that the capacitor voltage be maintained constant, since the sawtooth ripple present at the phase comparator output can cause incidental FM on the VCO output. The high-impedance properties of the FET stages minimize the above effect and further attenuation of the ripple components is provided by suitable filtering.

SYSTEM TRADEOFFS

Major considerations in frequency synthesizer design are power consumption and output spectral purity. It is easier to obtain high spectral purity with a high-phase comparison frequency F_R, since:

1) Higher ripple frequencies may be more easily filtered.
2) The higher modulating (ripple) frequencies result in smaller FM modulation indices producing relatively smaller FM sidebands.
3) A higher loop bandwidth would provide more effective degeneration of incidental FM caused by vibration.

However, in order to obtain the higher phase comparison frequencies, the divide-by-N counter must use high-speed digital circuits which necessarily consume more power.

In most airborne applications (which have the severest environmental specifications) power consumption is not a primary consideration. However, minimum power consumption is required for battery operated, tactical radio sets. Therefore, if low-speed counters are used for these applications, more-stringent filtering and shielding must be employed, and the VCO must be carefully packaged to minimize effects of vibration.

Alternate approaches offering both high phase comparison frequencies and low power consumption are being investigated.[2]

CONCLUSIONS

The practical implementation of analog functions with digital techniques is a relatively recent art. However, it is evident that these techniques will revolutionize the design of communications equipments. Digitizing of functions such as amplification, demodulation and filtering is being given strong emphasis in the ADCOM-70 program.

Improvements approaching an order of magnitude in size, reliability and power consumption are anticipated in the 1970's through the use of bipolar and MOS arrays. Batch processing manufacturing techniques will reduce systems costs. The entire control logic for a synthesizer such as described herein will be contained in a few flat-packs. The concept of a universal synthesizer with a high order of modular commonality for applications in all communications bands will be closer to realization.

ACKNOWLEDGEMENTS

The author wishes to acknowledge the contributions of O. J. Hanas, A. N. Kaschok, W. McNamara and E. A. Sherman in circuit design and test of the synthesizer.

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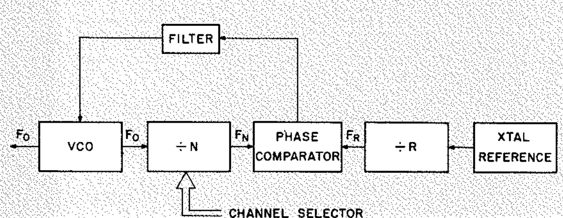

Fig. 1—Generalized block diagram—digital frequency synthesizer.

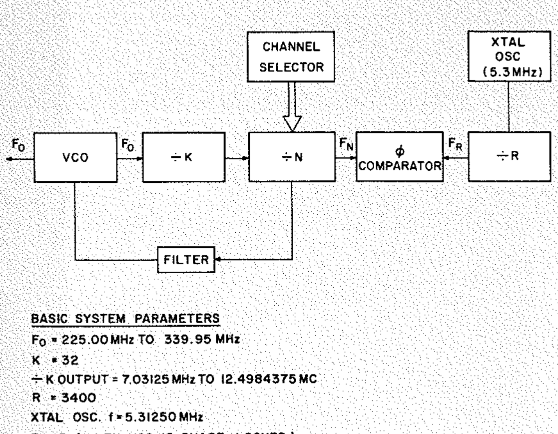

Fig. 2—Functional block diagram—UHF digital frequency synthesizer with parameters defined.

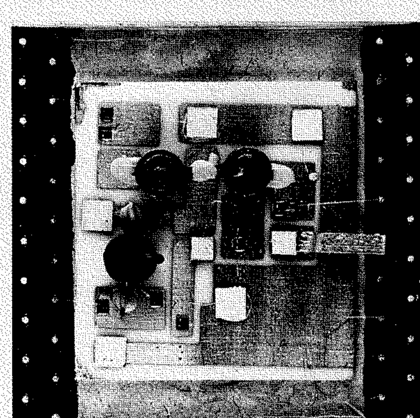

Fig. 3—UHF hybrid thin-film VCO; actual size is one square inch.

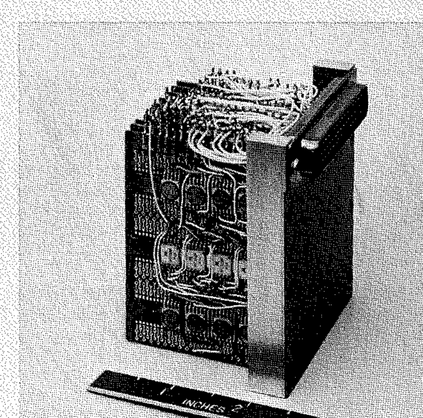

Fig. 4—Breadboard of 3500-channel UHF digital frequency synthesizer.

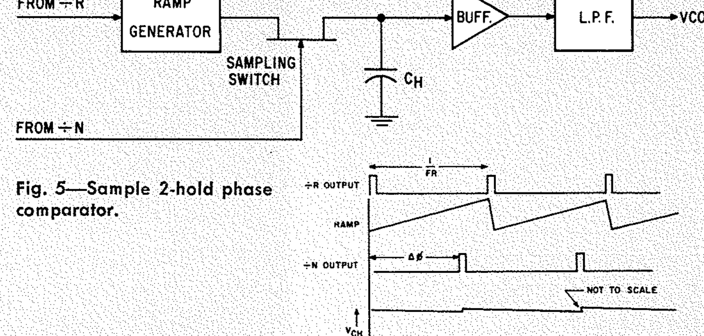

Fig. 5—Sample 2-hold phase comparator.

PRECISE ANALOG FREQUENCY SYNTHESIZER WITH OVENLESS STANDARD FOR MILITARY RADIOS

The development of an analog type frequency control or synthesizer provides high-order frequency stability for operation of SSB voice or CW telegraphic transceivers and narrow-band teletype communication equipment. The synthesizer design described herein is suitable for all frequency ranges, but was specifically designed for the AN/PRC-62 HF transceiver modified for use in the AN/ARC-104 transceiver and VHF AN/TRC-97 Marine Corps tropo communication equipments. The derivation of the synthesizer output frequencies is based on the control of low cost crystal oscillators (XLO's) by phase synchronization to a stable reference standard. Simplicity of operation is provided by the incremental or digitally selectable crystal banks calibrated directly in terms of the desired output frequency. This allows the synthesizer to be operated by personnel having a minimum of operator training and experience.

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ENGINEERS in the RCA Communications Systems Division have developed a precise frequency synthesizer with frequency stability of 1 part in 10 million *without* the traditional oven-controlled standard. The synthesizer offers a selection of 28,000 channels in 1-kHz increments by means of decade-set control knobs with direct frequency read-out. This analog synthesizer is based on increment selection of low-cost, crystal-locked reference oscillators (XLO's) combined by repetitive mixing and dividing. This new frequency synthesis technique is applicable to any transceivers where frequency control is critical; however, we will use the vehicle of the new CSD-designed AN/PRC-62 transceiver to describe the technique.

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Man-pack single-sideband radio sets in the HF and VHF ranges for the military have always posed the critical problems of frequency stability, size, weight and battery drain. The new synthesis technique plus unique packaging design have gone a long way toward solving these problems in the PRC-62. Both transmitter and receiver injection frequencies are generated by the synthesizer.

CSD has had extensive experience in synthesizer design over the past 10 years and in a more recent IR & D program achieved a significant improvement in performance and simplification of circuitry.

The development of the stable, miniature, ovenless frequency standard provides a breakthrough in improved stability, small size, light weight and low power drain. Previous designs using proportioned oven control require excessive battery drain because of heater needs and require insulating materials or a dewar flask to maintain elevated temperatures.

The ovenless standard described here uses a temperature compensating technique developed at RCA, rather than the brute-force methods of oven control. Knowing the crystal temperature characteristics, we developed an inverse temperature compensating network to counteract the frequency change of the crystal.

The new synthesizer compares favorably with competitive units. Problem areas resolved included spurious attenuation, phase modulation or jitter, signal-to-noise ratio, complexity, channel selection and cost. The capability of the basic approach has been proven and synthesizers for the AN/PRC-62 Army HF transceiver, the AN/ARC-104 Navy HF transceiver and the AN/TRC-97 Marine Corps tropospheric scatter system were supplied.

I. I. Grasheim (author at right) displays a PRC-62 combat radio module from the frequency synthesizer to S. V. Mormile, an electrical engineer on the PRC-62 project.

TABLE I—General Features of the AN/PRC-62 Solid-State Synthesizer.

Frequency Range	2.625 to 32.625 MHz
No. of Channels	28,000
Channel Spacing	1kHz
Stability	±0.75 ppm max. (Including ½-year aging allowance)
Temperature Range	—40 to 75°C
No Warm-up	Temperature compensated (no oven)
Primary Power Requirements	1.1 W, 9.5 V
Active Devices	Transistorized
Size	6 x 3 x 4½ inches
Weight	2 lbs

SYSTEM OF FOUR CRYSTAL-LOCKED OSCILLATORS

Briefly, the synthesizer system is based on frequency synthesis by incremental selection of low-cost crystal, phase-locked reference oscillators combined by repetitive mixing and dividing as shown in Fig. 1. Overall stability of the synthesizer injection signal is based on a miniature 3 MHz ovenless temperature-compensated, crystal-controlled frequency standard to which the above XLO's are phase-locked, and from which all additional reference frequencies are derived. Four crystal-locked oscillators provide the four variable frequency reference signals; the *MHz reference* (set by the megacycles control), the *100 kHz* (set by the 100-kHz control), the *10 kHz* (set by the 10 kHz control) and a *units-kHz* (set by the 1-kHz control). Each of the XLO's includes selectable crystals, which are phase-locked to a 100-kHz signal derived from the 3-MHz standard. The oscillators are designed with adequate pull-in to maintain lock over a $-40°C$ to $+90°C$ temperature range.

The four XLO's are combined in deriving the final output frequency as follows: The *units* XLO operates with 10 crystals selectable one at a time in the 18.5-to-19.4-MHz range, in 100-kHz steps. It is combined in the *units adder* with a 120-MHz reference frequency. The difference frequency is divided by ten to 10.15 MHz. The 10.06- to 10.15-MHz signal is fed to a *tens adder* unit where it is up-converted with a 105-MHz fixed frequency to a range of 115.06- to 115.15-MHz, and then mixed with crystals 13.7- to 14.6-MHz (10's select) to provide an output covering 128.76- to 129.75 MHz. This signal is again divided by ten to a range of 12.876- to 12.975-MHz. The *hundreds selector* XLO (16.5- to 17.4-MHz) is up-converted by mixing with a 90-MHz reference frequency and converted to a 119.376- to 120.375-MHz signal in the final mixer. The *megacycle* XLO operates at 12.3- to 15.2-MHz, utilizing 28 crystals, phase-locked, one at a time and selected by the MHz control. This unit feeds a ten-times multiplier providing a 123- to 152-MHz signal in megahertz steps to the final mixer circuit, where it is combined with the 119.376- to 120.375-MHz signal to produce output frequencies of 2.625- to 8.625-MHz for the three lower RF bands and 10.625- to 32.624-MHz for the three upper bands. The synthesizer output frequency is amplified, cleaned up by a tuned buffer amplifier, and injected through a receiver-transmit relay to the receiver mixer or to the transmitter balanced mixer.

The system (Fig. 1) incorporates the following advantages:

1) The stability of the complete system is dependent on one standard reference. This provides the flexibility of using a reference crystal with the desired accuracy or stability as required. The AN/PRC-62 frequency standard is contained in a 1-cubic-inch module and provides a long term stability of 1×10^{-7} ppm per month without the use of oven control.
2) Crystal locked oscillators provide a sinusoidal output inherently free from spurious frequencies. Sidebands are attenuated at least 75 dB by heavy filtering in the phase detector bias lines.
3) Spurious frequencies are controlled by proper frequency relationship and low level injection at the mixer inputs and effective filtering of the mixer output. Bandwidths of the filters are kept narrow and a further cleaning effect is obtained by the use of subsequent division.
4) Circuit similarity of the subassemblies used in generation of 1-, 10-, and 100-kHz digits permits the use of similar modules in those sections.
5) The system lends itself to 100-Hz step requirements by the addition of an additional decade unit and to rapid remote frequency selection by the use of diode switching techniques in the XLO's
6) Where further clean-up is required (100 dB spurious attenuation, or more) an output clean-up loop can be easily adapted.

STABILITY DETERMINING DEVICE

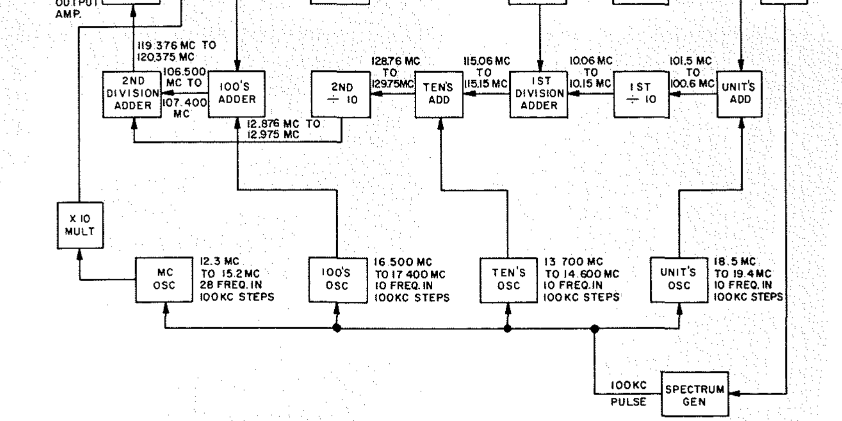
The 3-MHz frequency standard provides a stable, accurate source for reference signals used in the transceiver synthesizer. The basic frequency determining device is a 3-MHz, fundamental-frequency, AT-cut crystal, mounted in an HC-27/U type glass holder. The crystal is connected to the oscillator as well as to a temperature-compensating network that varies the bias voltage supplied to correction varactor diodes in accordance with temperature changes.

The oscillator output is taken from its collector and is capacitively-coupled to a buffer stage to isolate the oscillator from external circuitry, as well as to amplify the oscillator output. A small variable capacitor at the collector of the oscillator permits the oscillator to be tuned to its exact frequency to compensate for changes due to component aging.

The oscillator has been compensated to within 0.50 ppm over the temperature range of $-40°C$ to $80°C$. The short-

I. I. GRASHEIM received his BSEE from Drexel in 1940 and has been with RCA since this time. His primary responsibility is design and development of specialized miniature receivers and transmitters for military use. Mr. Grasheim has made valuable contributions to the development and design of a number of communication systems such as: MAR, a UHF communication receiver-transmitter, and the AN/URR-31, 32, 33 and 34 receivers. He was also the project engineer on the design of the SP-600 USAF Receiver single sideband system, AN/PRC-7, a high frequency pack set, and URR-7 and 8, high stability frequency shift teletype URR-33 and 34 Receiver. Mr. Grasheim's experience includes manpack and vehicular communication equipment. He has been involved in the continued development and analysis of advanced equipments in this field such as the AN/PRC-25, AN/PRC-34, and the AN/TRC-97 produced in large quantities. Mr. Grasheim specializes in the development of frequency control synthesizer tuning systems and was responsible for such system development in the AN/PRC-25, AN/PRC-35, AN/SRR-16, AN/URC-30, ARC-104, TRC-97 and the AN/PRC-62.

Fig. 1—AN/PRC-62 synthesizer block diagram.



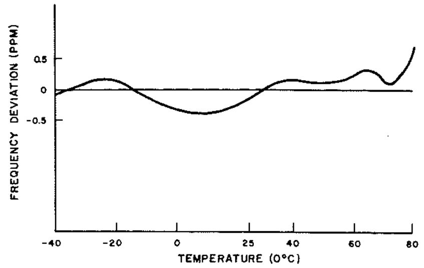


Fig. 2—Frequency standard deviation versus temperature.

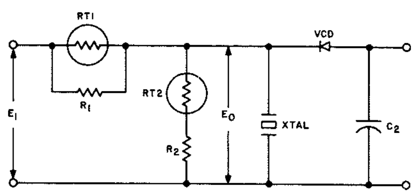


Fig. 3—Temperature compensating network.

term stability at the normal room temperature is better than 2×10^{-9} for the averaging time of 10 seconds. The average long-term stability based on a 49-day observation was better than 1×10^{-7}/month. Fig. 2 shows frequency vs temperature characteristic of a typical temperature compensated frequency standard.

The temperature compensating network (Fig. 3) consists of a voltage divider with a series-parallel combination of temperature sensitive fixed resistors. This combination is selected to provide an inverse bias for the temperature compensating variable-capacity diode which is in series with the total load capacity presented by the oscillator circuitry. The temperature versus bias characteristic of the inverse bias is shaped to have an equal and opposite effect on the oscillator frequency variations relative to temperature.

COMPENSATING NETWORK DESIGN

Assuming the type of compensating network as indicated in Fig. 3, and negligible loading by the variable capacity diode, we arrive at the following expressions:

$$E_o = \frac{E_1 Z_2}{Z_1 + Z_2}$$

$$\frac{E_1}{E_o} = N$$

Then:

$$\frac{Z_1 + Z_2}{Z_2} = N$$

$$\frac{Z_1}{Z_2} + 1 = N$$

And:

$$\frac{Z_1}{Z_2} = N - 1 = K$$

$$= \frac{\left(\dfrac{R_{T_1} R_1}{R_{T_1} + R_1}\right)}{R_{T_2} + R_2}$$

where: E_1 = available compensating voltage; and E_o = desired diode bias. The total resistance in the parallel leg of the divider is:

$$Z_1 = \frac{R_{T_1} R_1}{R_{T_1} + R_1}$$

And, the total resistance of the series leg is:

$$Z_2 = R_{T_2} + R_2$$

If this network could approximate the required K-versus-temperature characteristic close enough, then by taking experimentally four measurements of required compensatory voltage within the operating temperature range, the exact values of R_{T_1}, R_1, R_{T_2}, and R_2 could be calculated. In practice, due to the nonlinear (exponential) characteristic of R_{T_1} and R_{T_2}, calculations become involved. Some simplification can be achieved provided the exact resistance versus temperature characteristic of temperature sensitive resistors is known and can be directly substituted in the above equation for each test temperature with the resulting effect of having obtained the correct compensating bias for the specified test points. Normally in order to approximate the required compensating bias characteristic within the specific tolerances of ±0.5 ppm more than four elements are required with the resulting increase in complexity of the calculations to be performed. It would appear that a computer approach to the network calculations would be a natural solution to a very large scale production which would pay for computer use.

In small-to-moderate compensation processes a combination of empirical and simplified calculation approach is used. The steps that are usually followed are these:

1) With an external compensating bias applied to the variable capacitance diode a number of measurements are made at specified temperatures.
2) From the obtained values K, and the simplified four-element network is calculated and installed in the unit.
3) A second temperature run is taken and deviations of frequency from the specified limits noted.
4) Corrections and additions to the simplified network are introduced using either further calculations or compiled charts indicating the effect of various network elements on frequency, presuming a reasonable amount of experience and facilities for handling a larger number of units simultaneously, this approach can be quite economical.

DERIVATION OF FIXED FREQUENCY REFERENCE SIGNALS

A variable capacity divider circuit is used to divide the 3-MHz signal from the frequency standard to obtain a 100-kHz reference signal. Two divider circuits are used, a divide-by-six and a divide-by-five. Further amplification and pulse

Fig. 4—Divide-by-ten circuit.

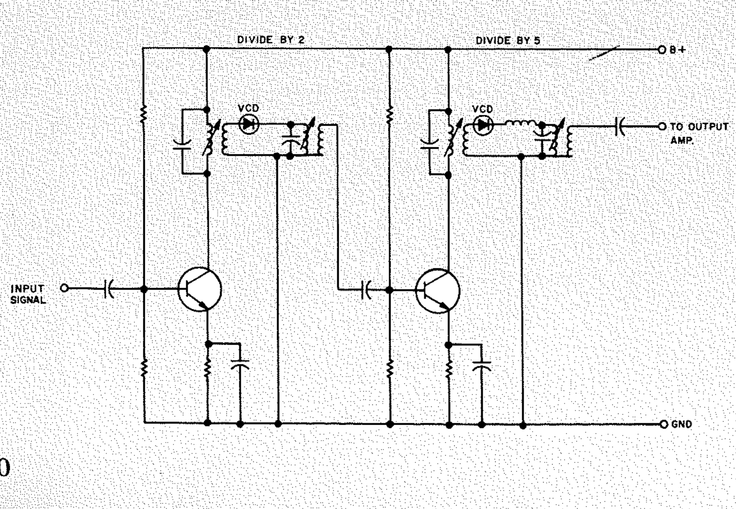
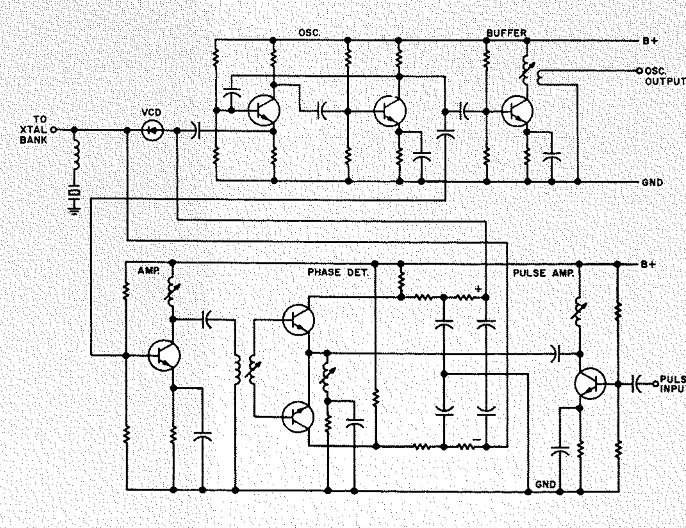


Fig. 5—Crystal locked oscillator typical circuit diagram.



shaping is obtained in each of the crystal-locked oscillator units. Problem areas are similar to those discussed in the divide-by-ten units and the circuitry is also identical to that shown on Fig. 4.

Three generators produce signals of 90, 105, and 120 MHz specifically for use in up-converting in the synthesizer mixers to minimize generation of spurious responses in the mixer circuitry. These signals are derived from the 3-MHz standard, and are of the same stability as the source signal.

All three generators use the same basic circuitry and transistor complement. Spurious measurements indicate the 3-MHz sideband are attenuated 65 dB and all others are greater than 75 dB. Temperature variations of −40 to +85°C result in ±1.5 dB change in output level.

DERIVATION OF STEP VARIABLE REFERENCE SIGNALS

A phase-locked, crystal-controlled oscillator is used to obtain the reference increments. The crystals, which are higher order multiples of 100 kHz, are selected one at a time, by the front panel control. A phase discriminator circuit compares and maintains the phase reference to the 100-kHz signal derived from the frequency standard. Crystal frequencies through a variable capacitance diode connected in series with the crystal can be trimmed by use of a small variable inductor in series with the crystal, to bring the crystal free running frequency to within its pull-in range.

The oscillator is a modified Butler circuit used so that one side of the crystal can be grounded. No tank circuit is required. The crystal is connected from the emitter to ground, acting as a bypass capacitor or low impedance at the crystal frequency and is connected in series with the varactor diode reactance element and the crystal-resonating trimmer inductor. Amplification takes place in the second transistor collector circuit and feedback is applied from the collector to the base of the oscillator transistor. Output from a buffer amplifier is link-coupled to a phase-detector driver stage and then transformer-coupled to the bases of an active phase discriminator. A train of 100-kHz pulses supplied from the 100-kHz generator is amplified and capacity-coupled to the emitter of the phase discriminator. Sampling of the oscillator output takes place in accordance with the phase of the 100-kHz pulses. Any phase error causes the discriminator transistor to conduct and supply a correction bias to the varactor reactance. A negative or positive bias causes a capacitive change in the voltage-sensitive varactor to oppose the change in oscillator frequency and maintain synchronization.

Various phase detector circuits were investigated in the design phase of the program. The passive type using two diodes was considered but because of its low impedance, it is necessary to provide a high-power, sharp pulse in order to obtain adequate DC swing of approximately 2 volts. This presents the problem of containing or filtering this high level pulse from the other circuits. The active differential type, although more critical to level changes, was chosen. This permits the use of a low-level pulse to obtain the same DC bias swing. The differential connection offers better stability under environmental changes.

ADDERS AND DIVIDERS

The adder receives the up-converted signals and mixes the signal with the increment oscillators. The collector is tuned to the difference frequency as indicated in Fig. 1. The signal is capacitively coupled in the emitter circuit. Output is obtained through a double-tuned circuit in the collector. An additional amplifier and tuned circuit permit rejection of the undesired input signals or higher order combinations of the input signals. The signal from the adder unit is divided by ten. Two variable capacity divider circuits are used: a divide-by-two, and a divide-by-five. Each is preceded by driver amplifier circuits and the resulting signal is amplified and fed through a low-pass filter to the following adder circuit.

Problems of design consisted mainly of obtaining adequate drive levels particularly at the low temperature end of the range and stabilization for variation in gain over the complete temperature range. Low impedance input and sufficient drive to cause limiting are used to prevent jitter at points within the passband. Both dividers are similar, with the varicap diodes self-biased.

The circuit performed best with silicon variable-capacity diodes having about 6.5-pF capacity at 4-volt reference. These diodes were readily interchangeable with a minimum of retuning. Stability was determined in terms of the extent to which a parameter could be varied while the output remained locked at one-tenth the input frequency. With input voltage variations of ±10% the input frequency can be varied ±3 MHz without loss of lock.

FORWARD LOOK

Man-pack transceivers such as the PRC-25 VHF-FM 920-channel set have proven themselves in the field under extreme jungle warfare conditions. The new PRC-62 SSB, HF 28000-channel transceiver will provide for an ever increasing load and longer range communications. Packaging efficiency improvement permits use of four times the number of components in approximately the same volume.

Fig. 6—Morris Lysobey, EE (left) and Roy A. Beers, Leader, pose with the AN/PRC-62 manpack military radio. Mr. Lysobey holds the new miniaturized frequency standard used in the PRC-62 in his right hand and compares it with the model used in previous radios.

Since the advent of integrated circuits, a forward look in combat radio can anticipate further microminiaturization including weight and size reduction. Frequency coverage, number of channels, and power output can be increased. Combat radios in the next 15 years will be further improved by the use of new components and new circuitry. New single wafer filters are available, packing high amounts of selectivity in small volumes. The varactor diodes provide parametric amplification and electric-tuning in the HF and VHF range for improved front-end performance. High speed diode switching makes remote control capability more flexible and does away with electro-mechanical switching. New balanced mixer and phase detector circuitry require no transformers or inductors, and direct coupling in the integrated-circuit packages eliminates large coupling capacitors. Hybrid adders simplify the solid-state, high-power requirement for transmitters.

The new components and circuit designs have set the stage for high performance and high reliability communications of the future.

ACKNOWLEDGEMENT

The author acknowledges the contribution of the members of the AN/PRC-62 design team in the preparation of this article: Engineering Leader, R. A. Beers, and the following design engineers on the AN/PRC-62 Synthesizer: M. Lysobey, R. DeCredico, R. J. Carignan, S. V. Mormile, W. R. Thompson, and J. L. Slivinski.

CADRE
Computer Aided Design and Reliability Engineering

ADCOM-70 effort in advanced technology has extended its scope of interest beyond the development of new circuits and equipment techniques into the area applying digital computers to relieve the design engineer of much of his clerical burden. Computer Aided Design and Reliability Engineering — CADRE — is a tool for the engineer and first levels of management to more effectively manage the data associated with components, subassemblies and equipment in the development and design phases.

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The digital computer can serve the design engineer in two separate and somewhat distinct roles. One is that of a computational tool for circuit analyses, filter design, thermal analyses, and similar design tasks. As such its utility is principally that of performing numerous complex computations in a highly efficient and rapid manner. Increasing numbers of engineers are learning that the cost of calculation is often less than that of experimentation in arriving at a final design.

The second role of the computer is that of an instrument for storage and rapid retrieval of engineering data. The design engineer and first levels of engineering management are feeling an ever increasing burden of equipment design data management in their day-to-day design tasks. The capability of the computer to rapidly retrieve facts, place them in a specified order, and perform necessary computation as directed, is that attribute of the computer with which CADRE is concerned. This data management function (CADRE) is principally concerned with providing a framework for:

1) Organizing and storing of equipment, circuit and component data, such as cost, reliability, size, weight, etc.
2) Processing and computation of stored data, for example, computing MTBF, summing up weight, cost, etc.
3) Retrieval, organization and concise presentation of both stored data and computed results.

The significance of the CADRE program can be appreciated only if the problem which it relieves is understood. The design engineer, responsible for an equipment design, reviews the specification and prepares a detailed preliminary design or technical proposal. He must identify the various functions in the equipment and describe in detail the circuitry which is required to perform these functions. He often has imposed upon him a number of conflicting requirements, among which are the following:

1) Meet all electrical performance specifications.
2) Achieve a specified reliability level
3) Maintain a competitive equipment cost
4) Meet weight and size specifications
5) Take maximum advantage of existing developments
6) Summarize and tabulate data on each component, module and subassembly sufficient for design review purposes
7) Identify each component for factory costing

To meet all of the above requirements and document the proposed design it is necessary for the engineer to have access to a wide range of parts data and to be able to process this data efficiently to obtain the answers required. This is a data filing and handling problem which when performed by hand soon overburdens the engineer and his clerical assistants. The CADRE system is based on selection and storing by experienced engineers of components and circuit stages which they know will provide certain performance and which in general have been proven. The novel feature in CADRE is the employment of a digital computer to rapidly process thousands of bits of information on any combination of stages which the engineer has established by drawing a block diagram. CADRE does not relieve the engineer from doing engineering nor does it supply judgment not already there.

The principles employed in the CADRE organization are presented in the following paragraphs.

A SYSTEMATIC FRAMEWORK

The accumulation, storing and calling out of facts pertaining to electronic systems equipments, circuits and parts has been established. The data selected for storage is readily obtainable from existing functions within RCA. For example,

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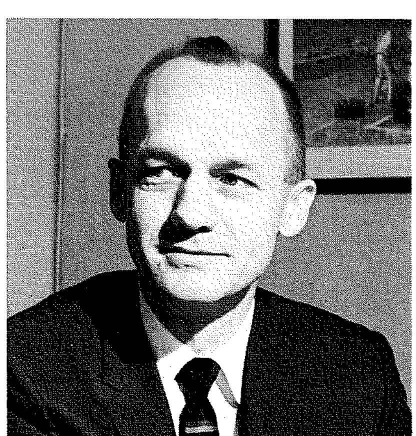


components selected for storage are those which the Design Group is using every day, those which appear in *Defense Standards*, and/or those new components which appear to be promising. This system does not duplicate or replace such functions as Central Engineering, Reliability Engineering, Drafting, or Cost Estimating. However, it puts at the engineers' and managers' fingertips basic data originated from these groups in a form which enables rapid decisions to be made and trade-offs to be well documented.

The framework established consists of data formats and symbols which are recognizable by the computer together with a formalized breakdown of an equipment similar to the drawing system. The finest breakdown level is that of the component and a complete computer tape is used to store electrical-mechanical characteristics, reliability numbers, vendors, prices, etc., on each component selected to be placed in the computer file. A second computer tape stores stages, modules, assemblies, equipments and systems or equipment grouping. This tape identifies those components and certain other stage data unique to a particular stage. In all, there are *five levels* in the CADRE processing system. This enables breakdown of any equipment or system into succeedingly finer groupings established as follows:

1) Equipment
2) Assembly
3) Subassembly
4) Stage

The equipment consists of one or more assemblies. The assembly is made up of one or more subassemblies, each of which consists of one or more stages. The stage is the basic building block and consists of one or more components, which are the raw material. Each component is given an identifying number and a data sheet, illustrated in Fig. 1, is filled out and key punched preparatory to reading in all component data into the component master file tape. Note that the data stored is both alphabetic and numeric. The numeric data is restricted to the center one-third of the form. Pertinent facts, such as cost in various quantities, failure rate, weight, size, and number of leads, as shown are stored for each component.

The stage is a functional circuit such as an oscillator, an amplifier, a phase detector, etc., or it can also be a group of mechanical components. The input data format for the stage level (Level 4) is shown in Fig. 2. The format is very similar to the component input data sheet, in that the upper six lines are allocated to alphabetic data. The next eight lines contain only numeric data, and the lower

Fig. 1—Component input sheet.

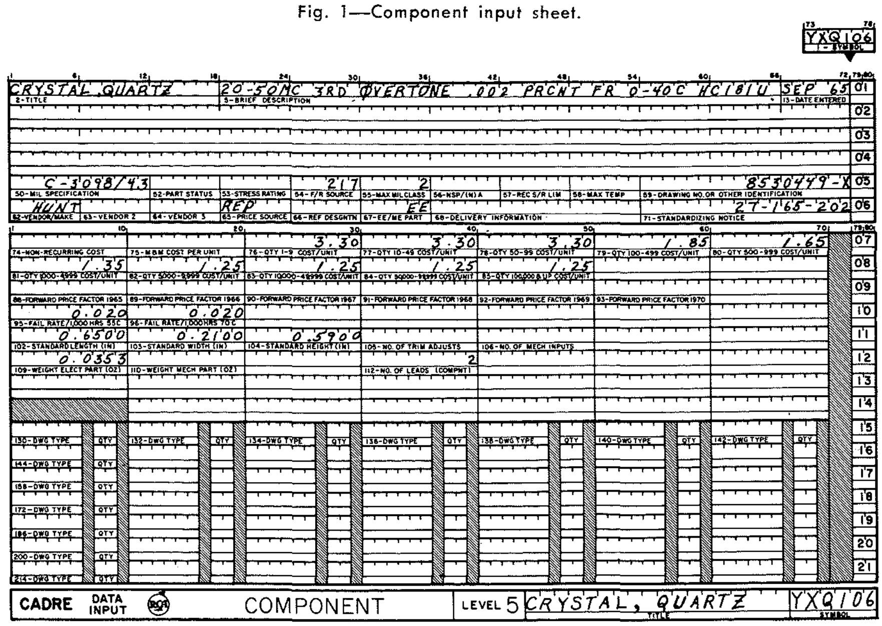
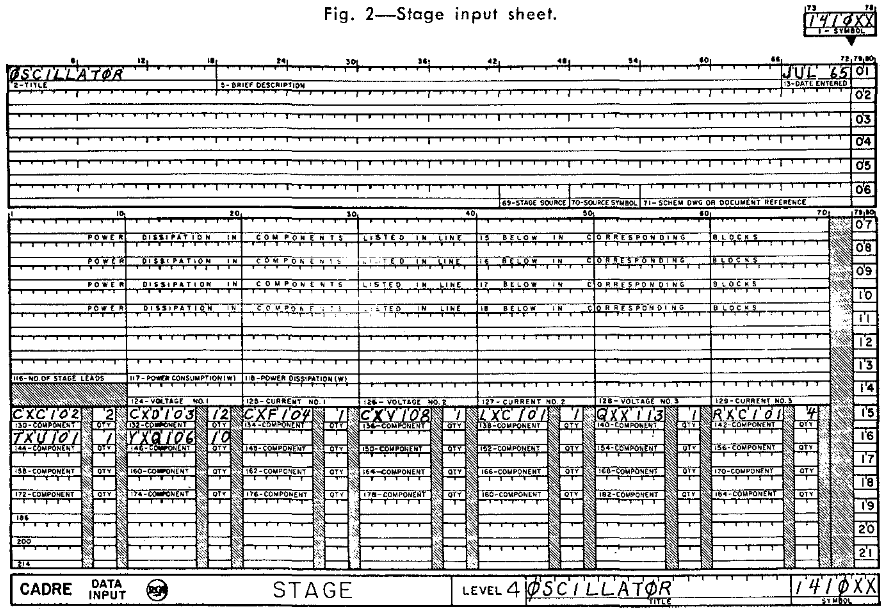
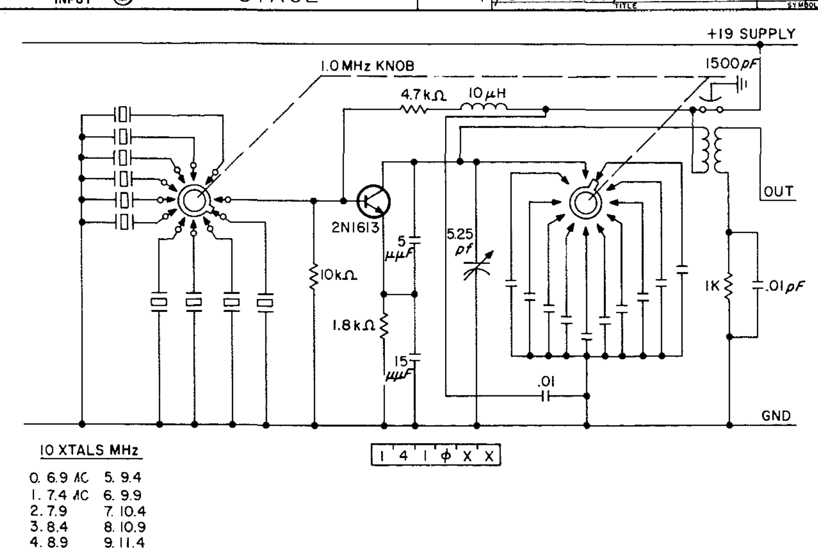


Fig. 2—Stage input sheet.





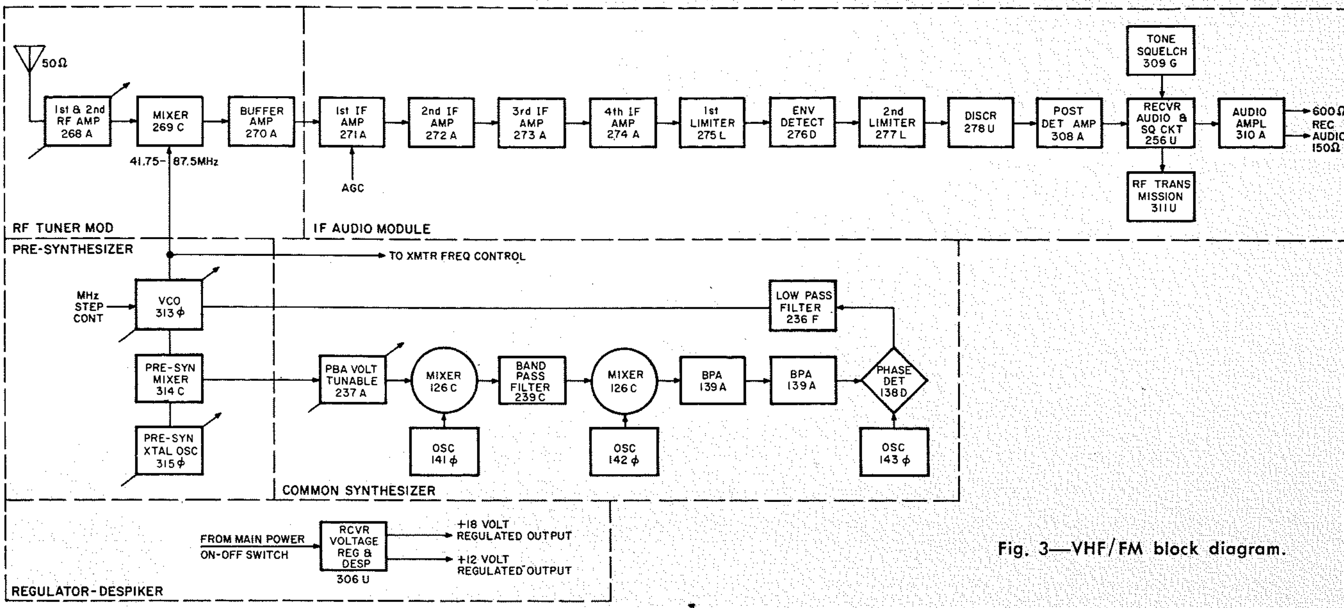


Fig. 3—VHF/FM block diagram.

seven lines contain space for the identification and quantity of each of the components which go to make up the stage. The *stage* building blocks were chosen since these are the functional elements which the design engineer considers in synthesizing a new equipment, or in describing a preliminary design. The data stored for each stage is limited to that which is unique to the group of components which make up that stage and give it its functional identity. There is no attempt to summarize component data and incorporate it on the input form for a stage inasmuch as the computer will accomplish this task. The numeric data stored by stage is limited to the power dissipation by component where this data is significant. It is evident from the description thus far that power dissipation cannot be stored in the component file since a particular component will have an operation unique to the particular stage in which it is being used. The stage data sheets are converted to punched cards and read into the stage master file tape.

VERSATILE PROGRAMMING CAPABILITY

It is possible to rapidly and with minimum programming skill establish new requirements for extracting and processing data which is required. This has been incorporated to meet the variety of requirements and the fast turn around time required. The need to continuously update equipments and monitor decisions is further reason for establishing the versatile programming capability.

COMPREHENSIVE AND VERSATILE DATA PRINTOUTS

The systematic framework above is of value in itself, however, the real purpose of CADRE is to provide the capability to produce comprehensive output data. The data obtainable, for example, can be a complete block diagram breakdown of an equipment to the component level with any combination of cost, size, weight, reliability, etc., data accumulated and extended at the levels of component, module, subassembly, assembly, and/or the equipment. A second format for data is the sorting and accumulating of like parts in an equipment without regard to their location by stage or module. Such a sort, for example, will present the number of carbon resistors, tantalum capacitors, power transistors, etc., together with their cumulative bit failure rate, weight, size, and the like. Sorting may be alphanumeric, increasing/decreasing order of magnitude on any parameter, or selecting of components on the basis of any parameter exceeding a certain threshold (for example, printout of all components costing $50 or more).

EXAMPLE OF USE

To illustrate the operation of CADRE, a specific equipment is cited and typical elements of the input and output are described below:

A communications group responsible for a VHF-FM receiver design receives an equipment specification calling for an advanced design. Engineering discussions and decisions proceed to the point where a block diagram is drawn which appears to accomplish all electrical performance requirements imposed by the specification. This preliminary design is shown in the block diagram of Fig. 3. The separate functional blocks are enclosed within dashed lines to indicate a "first cut" on the packaging of the stages into the subassemblies. The complement of subassemblies comprises the complete equipment (Level 1). Where possible, the engineer has chosen the functional stages from his CADRE catalog of stages which have been accumulated over a period of time from existing designs, advanced development projects, or circuits obtained from sources outside of his immediate group. Those stages which remain unidentified are synthesized as a paper design using the engineer's best judgment. He will choose components from the CADRE catalog of stored components wherever possible. Where components do not exist in the file he must take steps to store the new components to be used. He then proceeds also to store the new stage data by identifying the components chosen. Once he has drawn his block diagram and identified the blocks as stages which are stored, he has only to call for the type of data processing and printout to be performed.

Figs. 4 through 7 contain computer printouts which show representative output data capabilities.

CADRE INPUT DATA LIST

VHF/FM	VHF/FM RECEIVER
0112RA	VHF/FM R-F TUNER MOD.
268A	1ST AND 2ND RF AMPLIFIERS
269C	MIXER
270A	BUFFER AMPLIFIER
0114NN	VHF/FM PRE-SYNTHESIZER
315O	PRE-SYNTH. XTAL OSC.
314C	PRE-SYNTH. MIXER
313O	VCO
0106NN	VHF COMMON SYNTHESIZER
237A	VOLTAGE TUNABLE BPA
126C	MIXER (2)
141O	OSCILLATOR
239F	BANDPASS FILTER
142O	OSCILLATOR
139A	BANDPASS AMPL. (2)
143O	OSCILLATOR
138D	PHASE DETECTOR
236F	LOWPASS FILTER
0115RA	VHF/FM IF AUDIO MODULE
271A	1ST IF AMPLIFIER
272A	2ND IF AMPLIFIER
273A	3RD IF AMPLIFIER
274A	4TH IF AMPLIFIER
275L	1ST LIMITER
276D	ENVELOPE DETECTOR
277L	2ND LIMITER
278U	DISCRIMINATOR
308A	POST DET. AMPL.
309G	TONE SQUELCH
256U	RECVR AUDIO AND SQUELCH CKT.
311U	RETRANSMISSION CKT.
310A	AUDIO AMPL.
0119RD	REGULATOR-DESPIKER
306U	RECVR VOLT. REG. AND DESPIKER

Fig. 4—Tabulation of input data for VHF/FM receiver.

Fig. 4 is the tabulation of the input stages and subassemblies comprising the equipment and identified by the CADRE symbol.

Fig. 5 is a representative page of the first breakdown requested showing costs by component and stage for various quantities of equipments. As an example, the *oscillator stage 1410* is comprised of components in the quantities shown in the tabulation. Component unit costs and component costs are accumulated in columns for each equipment quantity requested. The illustrative component input sheet of Fig. 1 for *quartz crystal YXQ106* appears near the bottom of this tabulation. A total of 10 of these units are used in this function, however the prices used in the compilation take into account, if necessary, the fact that more of these crystals may be used elsewhere in the receiver. The components for this stage are totaled and printed as stage totals for each quantity. The engineer may examine these prices and readily identify those stages which are contributing most to the equipment cost. CADRE catalogs can be consulted for potentially lower cost alternatives where it might be beneficial to the equipment design.

Fig. 6 shows the final page of the reliability calculations. The failure rates are summed for each stage, subassembly and overall equipment and the corresponding MTBF's at each of these levels are calculated. The results show that at 55°C., the MTBF for the overall equipment is 15,500 hours. This drops to 8,601 hours at 70°C.

Fig. 7 contains a tabulation of several of the physical constants of weight and size by component and a calculation of area, volume and density. These figures are not intended to be final design goals since the squared off area and volume of the components are used as the basis for calculation. However, the numbers appearing will be related to the final volume by a factor which will be established as experience accumulates in comparing computed results with actual design results. The chassis weight was computed from a simple relationship involving component areas and is intended to be only a first approximation. Storage and processing of the mechanical components and stages were omitted from this example, hence they do not appear in any of the tabulations.

SYMBOL		TITLE	VENDOR	NON-RECUR. COST	COST IN QTY 1-9	COST PER/EQUIP QTY (1)	COST PER/EQUIP QTY (5)	COST PER/EQUIP QTY (8)	COST PER/EQUIP QTY (15)	COST PER/EQUIP QTY (500)
RXC101		RESISTOR, CARBON		-0.	0.05	0.05	0.05	0.05	0.05	0.05
TOTAL	3			-0.		0.15	0.15	0.15	0.15	0.15
TTH102		TRANSFORMER TOROID	RCA	-0.	2.00	2.00	2.00	2.00	2.00	2.00
TOTAL	1			-0.		2.00	2.00	2.00	2.00	2.00
•TOTAL	1			-0.		11.75	11.75	11.35	10.33	8.96
126C		MIXER (2)								
CXD103		CAPACITOR,DURMICA	ELMNCO	-0.	0.25	0.25	0.25	0.25	0.25	0.25
TOTAL	1			-0.		0.25	0.25	0.25	0.25	0.25
LXU104		COIL	RCA	-0.	3.00	3.00	3.00	3.00	3.00	2.78
TOTAL	1			-0.		3.00	3.00	3.00	3.00	2.78
QXX113		TRANSISTOR 2N1613	RCA	-0.	2.15	2.15	2.15	2.15	1.43	1.20
TOTAL	1			-0.		2.15	2.15	2.15	1.43	1.20
RXC101		RESISTOR, CARBON		-0.	0.05	0.05	0.05	0.05	0.05	0.05
TOTAL	2			-0.		0.10	0.10	0.10	0.10	0.10
EACH	1			-0.		5.50	5.50	5.50	4.78	4.33
•TOTAL	2			-0.		11.00	11.00	11.00	9.56	8.66
1410		OSCILLATOR								
CXC102		CAPACITOR,CERAMIC	VITR	-0.	1.00	1.00	1.00	0.80	0.65	0.65
TOTAL	2			-0.		2.00	2.00	1.60	1.30	1.30
CXD103		CAPACITOR,DURMICA	ELMNCO	-0.	0.25	0.25	0.25	0.25	0.25	0.25
TOTAL	12			-0.		3.00	3.00	3.00	3.00	3.00
CXF104		CAPACITOR,FEEDTHRU	ERIE	-0.	0.30	0.30	0.30	0.30	0.30	0.30
TOTAL	1			-0.		0.30	0.30	0.30	0.30	0.30
CXV108		CAPACITOR,VAR CRMC	ERIE	-0.	1.70	1.70	1.70	1.70	1.70	0.56
TOTAL	1			-0.		1.70	1.70	1.70	1.70	0.56
LXC101		COIL, CHOKE	RCA	-0.	0.40	0.40	0.40	0.40	0.40	0.40
TOTAL	1			-0.		0.40	0.40	0.40	0.40	0.40
QXX113		TRANSISTOR 2N1613	RCA	-0.	2.15	2.15	2.15	2.15	1.43	1.20
TOTAL	1			-0.		2.15	2.15	2.15	1.43	1.20
RXC101		RESISTOR, CARBON		-0.	0.05	0.05	0.05	0.05	0.05	0.05
TOTAL	4			-0.		0.20	0.20	0.20	0.20	0.20
TXU101		TRANSFORMER	RCA	-0.	3.09	3.09	3.09	3.09	3.09	2.78
TOTAL	1			-0.		3.09	3.09	3.09	3.09	2.78
YXQ106		CRYSTAL,QUARTZ	HUNT	-0.	3.30	3.30	3.30	1.85	1.85	1.25
TOTAL	10			-0.		33.00	33.00	18.50	18.50	12.50
•TOTAL	1			-0.		45.84	45.84	30.94	29.92	22.24

Fig. 5—Representative portion of pricing breakdown.

Fig. 6—MTBF calculations for VHF/FM receiver.

SYMBOL		TITLE	PART STATUS	FAIL RATE AT 55 DEG.	FAIL RATE AT 70 DEG.	NO. OF COMP. USED	MTBF AT 55 DEG.	MTBF AT 70 DEG.
RXC101		RESISTOR, CARBON	QPL	0.0020	0.0060			
TOTAL	5			0.0100	0.0300	5.		
•TOTAL	1			0.4011	0.8621	14.	2.493E 05	1.160E 05
••TOTAL	1			0.4011	0.8621	14.	2.493E 05	1.160E 05
••••TOTAL	1			6.4500	11.6270	473.	1.550E 04	8.601E 03

SYMBOL		TITLE	WEIGHT	LENGTH	WIDTH	HEIGHT	AREA	VOLUME	NO. OF COMP.	COMP. PER SQ. INCH	COMP. PER CU. INCH
CXC101		CAPACITOR,CERAMIC	0.0210	0.5400	0.1300	0.1300	0.0702	0.0091		14.2450	109.5770
TOTAL	2		0.0420				0.1404	0.0183	2.		
CXN105		CAPACITOR,TANTALUM	0.0530	0.8600	0.2500	0.2500	0.2150	0.0537		4.6512	18.6047
TOTAL	6		0.3180				1.2900	0.3225	6.		
QXX143		TRANSISTOR 2N3565	0.0020	0.3500	0.3500	0.3600	0.1225	0.0441		8.1633	22.6757
TOTAL	3		0.0060				0.3675	0.1323	3.		
RXC101		RESISTOR, CARBON	0.0101	0.4000	0.1500	0.1500	0.0600	0.0090		16.6667	111.1111
TOTAL	14		0.1414				0.8400	0.1260	14.		
RXV104		RESISTOR,POT	0.3220	0.5600	0.5600	0.3200	0.3136	0.1004		3.1888	9.9649
TOTAL	1		0.3220				0.3136	0.1004	1.		
TXU107		TRANSFORMER OUTPUT	1.0000	0.6000	0.8300	1.0000	0.4980	0.4980		2.0080	2.0080
TOTAL	1		1.0000				0.4980	0.4980	1.		
•TOTAL	1		1.8294				3.4495	1.1974	27.	7.8272	22.5488
••TOTAL	1		5.7833				28.8526	8.4420	207.	7.1744	24.5201
			CHASSIS WEIGHT=	2.9529							
0119RD		REGULATOR-DESPIKER									
306U		RECVR VOLT. REG. AND DESPIKER									
CRG144		DIODE 1N458	0.0070	0.5150	0.1650	0.1650	0.0850	0.0140		11.7682	71.3222
TOTAL	1		0.0070				0.0850	0.0140	1.		
CRZ143		DIODE 1N753A	0.0070	0.5500	0.1900	0.1900	0.1045	0.0199		9.5694	50.3651
TOTAL	1		0.0070				0.1045	0.0199	1.		
CRZ145		DIODE 1N3031	0.0490	0.8100	0.3250	0.3250	0.2632	0.0856		3.7987	11.6882
TOTAL	1		0.0490				0.2632	0.0856	1.		
CXC110		CAPACITOR,CERAMIC	0.0177	0.5400	0.1300	0.1300	0.0702	0.0091		14.2450	109.5770
TOTAL	1		0.0177				0.0702	0.0091	1.		
CXN105		CAPACITOR,TANTALUM	0.0530	0.8600	0.2500	0.2500	0.2150	0.0537		4.6512	18.6047
TOTAL	1		0.0530				0.2150	0.0537	1.		
QXX151		TRANSISTOR 2N3133	0.0170	0.3950	0.3950	0.3200	0.1560	0.0499		6.4092	20.0288
TOTAL	2		0.0340				0.3120	0.0999	2.		
QXX152		TRANSISTOR 2N3567	0.0200	0.4300	0.4300	0.3600	0.1849	0.0666		5.4083	15.0231
TOTAL	2		0.0400				0.3698	0.1331	2.		
RXC101		RESISTOR, CARBON	0.0101	0.4000	0.1500	0.1500	0.0600	0.0090		16.6667	111.1111
TOTAL	5		0.0505				0.3000	0.0450	5.		
•TOTAL	1		0.2582				1.7198	0.4603	14.	8.1406	30.4155
••TOTAL	1		0.2582				1.7198	0.4603	14.	8.1406	30.4155
			CHASSIS WEIGHT=	0.3376							
••••TOTAL	1		67.2812				89.8450	47.5845	473.	5.2646	9.9402
			CHASSIS WEIGHT=	9.6994							

Fig. 7—Computer calculation and tabulation of physical constants.

PRESENT STATUS

The CADRE program is currently in the stage of debugging and is not in full operational use. The primary application to date has been in documenting experimental work and furnishing data for technical proposals. It is planned to implement a more complete components file and continue to accumulate preferred circuits in the form of stage and subassembly inputs prior to wider application.

The principal benefits of CADRE for the near future appear to be in the synthesizing of optimum designs, documentation of equipment in the development stages and control of the cost, size, weight, reliability, etc., of a design up to the time it is fully documented in formal drawings and data packages. The type of detailed data accumulated in the CADRE files is identical to that which is required in the more formal packages and wherever possible the CADRE data sheets contain drawing numbers of components or subassemblies where they apply. In operation a smooth transistion is predicted in phasing a design from the flexible CADRE system into a formal computer controlled drawing configuration and control system.

ACKNOWLEDGEMENTS

The CADRE program was developed in close cooperation with M. S. Corrington of Applied Research. Credit is due T. C. Hilinski and W. B. Schaming for the programming and debugging of CADRE.

NEW TECHNIQUES IN MICROWAVE DESIGN

The most significant new technique in the development of microwave circuits is the utilization of coupled lines in the transverse electro-magnetic field. This technique has the advantage of compressing the circuitry volume due to the absence of physical coupling elements. The bases for this technique are the even- and odd-mode characteristic impedances of the transmission line. A method is presented for calculating these impedances for microstrip lines. The microstrip line is a single ground-plane line with widespread application in microelectronics at microwaves. The method employed is one of dividing the cross-section of the transmission line into regions with respect to the uniformity of the field distribution. This reduces the problem to conductor configurations for which closed form solutions exist. The analytical approach is supplemented by experimental techniques to determine the characteristics of the microstrip line for a selected substrate material. Curves and tables are presented.

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SEMICONDUCTOR devices operating at microwaves are the major contributors to the trend of microwave miniaturization. Microwave systems in waveguide already outsize and outweigh all the supporting lower frequency components of integrated circuits.

The microstrip transmission line on a high-dielectric-constant material is the way to microwave miniaturization. The high dielectric constant of the transmission line reduces the size of microwave circuit components. To utilize the microstrip, the characteristics of this transmission line must be expressed and verified. These characteristics are the even- and odd-mode impedances and the propagation constant. The odd-mode characteristic impedance is important for miniaturization of coupled circuits to eliminate the physically mutual component.

CHARACTERISTICS OF THE DIELECTRIC SUBSTRATE MATERIAL

The criteria for selecting the material for the substrate is minimum size and lowest dielectric loss. The size of circuitry is directly related to the wave length λ, and losses are characterized by the dissipation factor which is proportional to the attenuation constant of the dielectric α_d. The best material for miniaturization is the one with the lowest ($\alpha_d\lambda$) product even though for some circuit configuration the conductor losses become dominant.

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Relative $\alpha_d\lambda$ products are tabulated below:

	Dielectric	$\alpha_d\lambda$
Rexolite or Teflon		3
Fused Quartz	(SiO_2)	15
Beryllia	(BeO)	8
Alpha Alumina	(Al_2O_3)-alpha	0.85
Rutile	(TiO_2)	60

The alpha-alumina of 99.9% purity was selected. It is a polycrystalline ceramic with a dielectric constant of 9.9 and a dissipation factor of 0.000025 at 8 GHz. This material is an excellent substrate for miniature microwave circuits.

CHARACTERISTICS OF THE MICROSTRIP LINE

Propagation Constant

The propagation constant of a transmission line is defined by the two terms:

$$\gamma = \alpha + j\beta \tag{1}$$

where α is the attenuation constant and β is the phase constant. The attenuation constant is the sum of the dielectric losses and conductor losses. Having knowledge of the characteristic impedance of the microstrip line, the expression for α and β can be derived in terms of the velocity of propagation v. The expression for both terms of the propagation constant have been derived but the work is too long to be included in this exposition. The velocity of propagation is for a TEM mode equal to the speed of light c over the square root of the dielectric constant ε. The change in the velocity of propagation, which takes place in the microstrip line, is related to the amount of field in the air as compared to homogeneous dielectric.

The velocity of propagation in a microstrip, v_m is:

$$v_m = ck$$

$$k = \left[1 + l^2(\varepsilon - 1)\right]^{-1/2} \tag{2}$$

$$l = \frac{\text{homogeneous dielectric impedance}}{\text{microstrip impedance}}$$

Characteristic Impedance

Like the coax transmission line, the parallel plane transmission line propagates energy in the TEM mode, first of all. Of course, a small field vector exists in the direction of propagation for the microstrip line. For the following graphical solution to the characteristic impedance, this fact may be ignored.

$$Z_o = \frac{1}{vC} \tag{3}$$

The geometry of the transmission line may serve as the basis for mapping the field. To determine the capacitance per unit length one could attempt a solution of the Laplace equation by the conformal-transformation method. However, there is no known closed-form

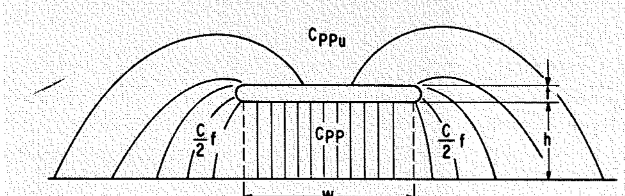

Fig. 1—Simplified map of even-mode transmission (dielectric not shown).

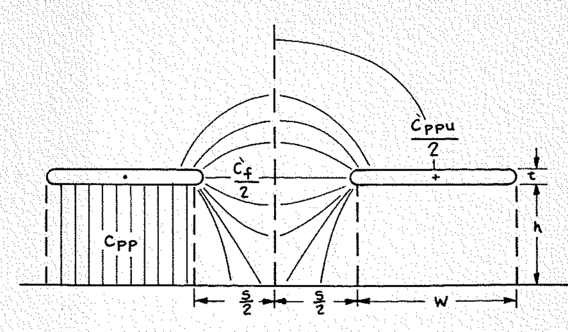

Fig. 2—Simplified map of odd-mode transmission.

solution for this configuration other than a zero-thickness approximation.

A practical approach is to break the field distribution of the transmission line into regions for which the capacitances are known or easy to calculate. The capacitance per unit length of Fig. 1 consists of three terms:

$$C = C_{pp} + C_{ppu} + C_f \qquad (4)$$

where C_{pp} is the parallel plate capacitance, C_{ppu} is the capacitance due to the upper plane and C_f is the fringing capacitance.

$$C_{pp} = \frac{\varepsilon}{c\eta}\left[\frac{w}{h}\right]$$

where η is the free space impedance

$$C_f = \frac{\varepsilon}{c\eta}\left[\frac{2.7}{\log\left(\frac{4h}{t}\right)}\right]$$

C_f is the round-wire equivalent as $W \rightarrow 0$. The following is determined by graphical methods from the map of the field in the upper plane:

$$C_{ppu} = \frac{\varepsilon}{c\eta}\left[\frac{2W}{3h\sqrt{\varepsilon}}\right]$$

The characteristic impedance of the microstrip line, that is, the even-mode characteristic impedance of the line becomes:

$$Z_o = Z_{oe} = \frac{h}{k\varepsilon}\,\frac{1}{\frac{W}{h} + \frac{2W}{3h\sqrt{\varepsilon}} + \frac{2.7}{\log\left(\frac{4h}{t}\right)}} \qquad (5)$$

For uncoupled lines Z_o equals Z_{oe}. The nonhomogenuity of the dielectric constant has been accounted for directly in the upper plane capacitances. Therefore, in the impedance expressions Z_{oe} and Z_{oo}, the value of the $k\varepsilon = \sqrt{\varepsilon}$.

The results obtained from this equation are in close agreement with actual

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measured values. There is also good correlation between the results from this expression and Wheeler.[3] Expressions by Assadourian and Rimai,[1] and Dukes[2] agree only for low impedances.

In a similar way, the odd-mode characteristic impedance is derived. The difference is in the fringing capacitance, C_f and C_{ppu}. Here, C_f is the round-wire equivalent as $W \rightarrow 0$ in an E-field boundary of a distance h above ground and a distance $S/2$ from perpendicular sides. The upper plane contribution C_{ppu} for the odd mode is represented graphically[4] by the simplified map of Fig. 3. The equipotential lines of this field are almost radial. The solution to the capacitance for this configuration is the summation of the capacitances between a series of inclined planes defined by the equipotential lines.

In terms of the inclined plane capacitance:

$$\frac{C'_{ppu}}{2} = \frac{\varepsilon}{c\eta}\,\frac{W}{h+a}\left(\frac{1}{3} + \frac{1}{9}\,\frac{a^2}{(h+a)^2}\right)$$

where:

$$h + a = \left(\frac{S}{2} + \frac{W}{2}\right)\sin\frac{\pi}{6} = \frac{S+W}{4}$$

The second term being negligible,

$$\frac{C'_{ppu}}{2} = \frac{\varepsilon}{c\eta}\,\frac{W}{S+W}\left(\frac{4}{3}\right)$$

$$\rightarrow C'_{ppu} = \frac{\varepsilon}{c\eta}\,\frac{1}{\frac{S}{W}+1}\,\frac{8}{3}$$

The odd-mode characteristic impedance becomes:

$$Z_{oo} = \frac{h}{k\varepsilon}\,\frac{1}{\frac{W}{h} + \frac{8}{3\sqrt{\varepsilon}}\,\frac{1}{\frac{S}{W}+1} + \frac{2.7}{\log\frac{4S\tanh 4h/S}{\pi t}}} \qquad (6)$$

where the last term in the denominator is C'_f.

This expression is in agreement with the odd-mode impedance measurements made in an experiment described later. The odd-mode impedance expression for single edge coupling, Z'_{oo}, can be derived from the above. The capacitance per unit length here consists of five terms:

$$C = C_{pp} + \frac{C_{pp}}{2} + \frac{C_f}{2} + \frac{C'_{ppu}}{2} + \frac{C'_f}{2} \qquad (7)$$

The family of curves in Fig. 4 are even- and odd-mode characteristic impedances for microstrip line. They are of a specific substrate and for equal line widths only.

The accuracy for the impedance expressions is highest for low impedances and low frequencies. Lower accuracy can be expected at higher frequencies and impedances. In general, for impedances less than 75 ohms and the line-width not greater than $0.1\lambda_g$, the graphical solutions are representative of the actual. These restrictions offer no great difficulty since most networks require no higher impedances with extreme accuracy and the line-width may be reduced by reducing the height of the substrate.

Experimental Techniques

The selection of alpha-alumina only as

Fig. 3—Simplified map of the odd-mode upper plane contribution.

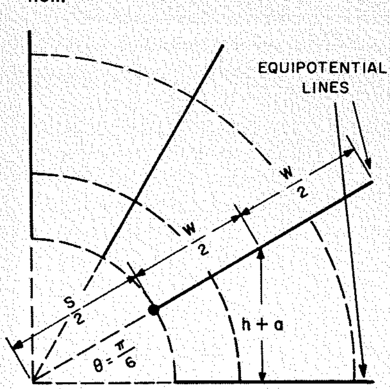
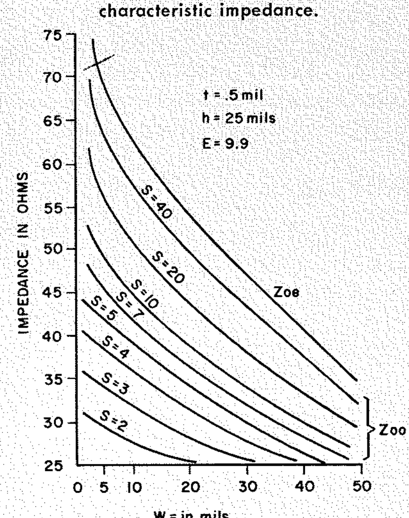


Fig. 4—Even and odd-mode characteristic impedance.



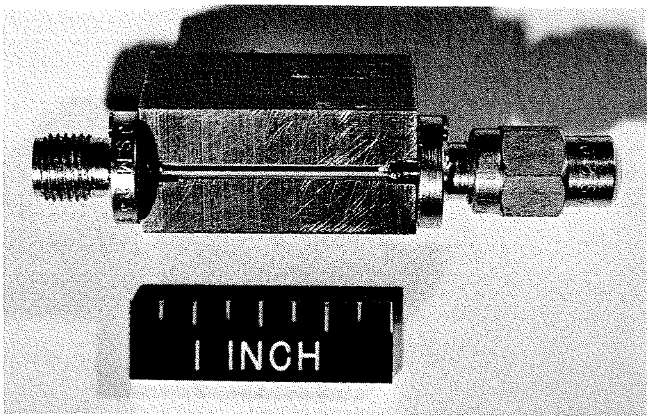


Fig. 5—Odd-mode impedance test fixture.

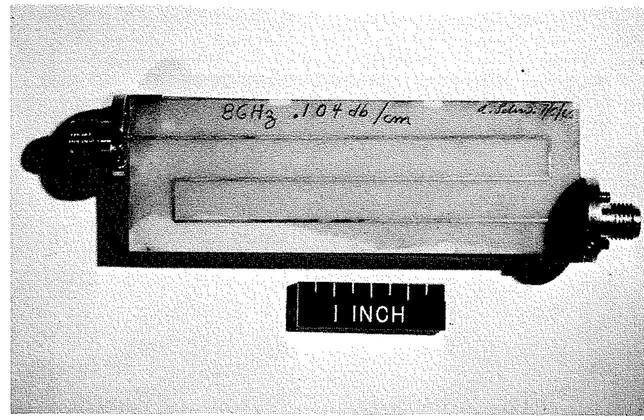


Fig. 6—Attenuation constant test set.

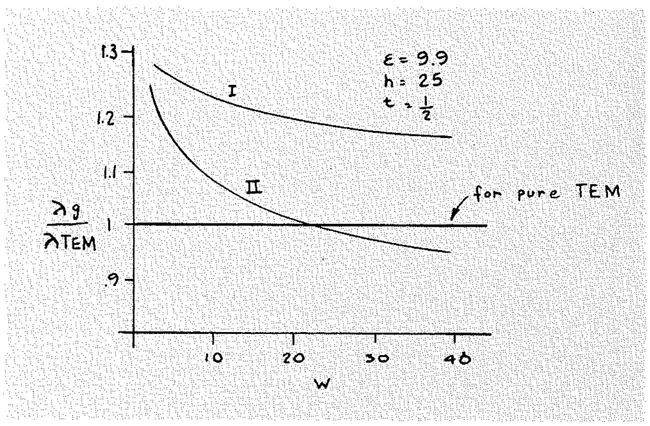


Fig. 7—Guide wavelength of microstrip line with and without end capacitance effects.

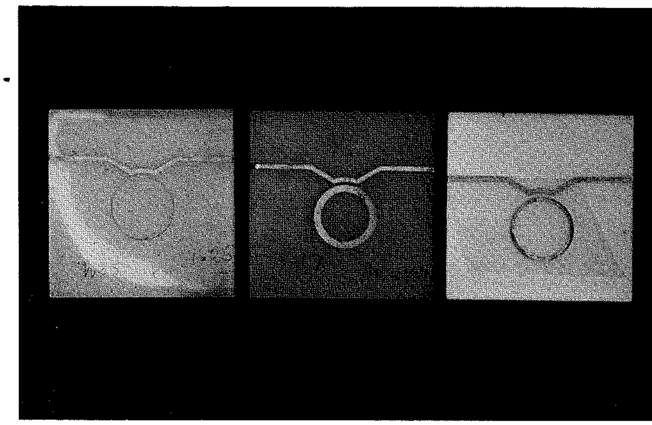


Fig. 8—Guide wavelength test sets.

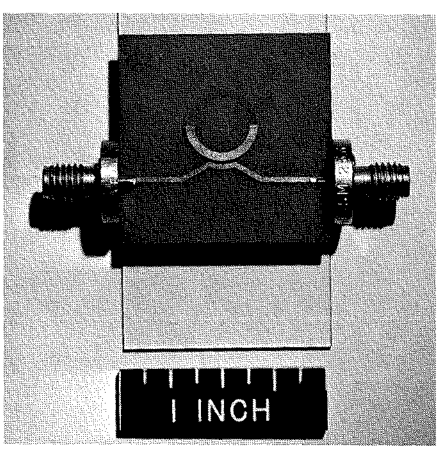


Fig. 9—Capacitance end-effect test set.

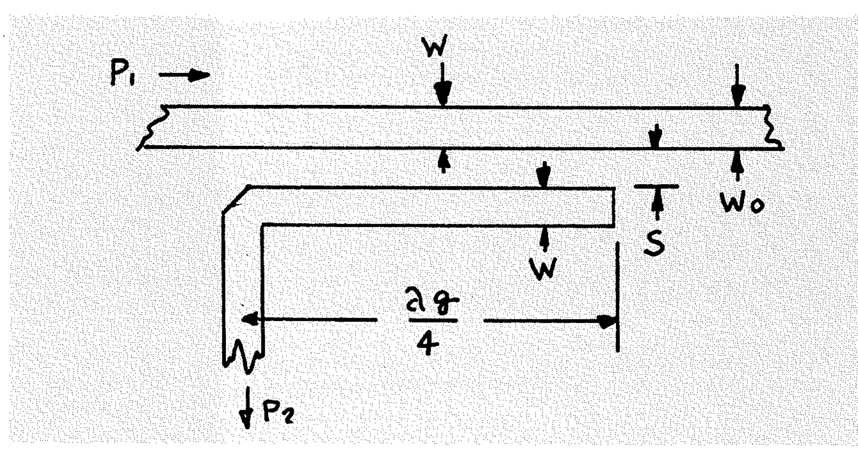


Fig. 10—Schematic of quarter-wave directional coupler.

the substrate material simplified the experimental verification of the microstrip properties.

The even-mode characteristic impedance determination consisted of measuring the surge impedance of various line-widths for a 1-inch length at X-band. The odd-mode characteristic impedance can be experimentally represented by placing a physical ground plane at a distance $S/2$ from the strip where the E-field boundary would occur. For the odd-mode experiment one such ground plane to either side of the microstrip is the equivalent representation of this mode. The odd-mode test set is shown in Fig. 5. Tight-fitting slices of various line-widths in the trough-type fixture were measured in the same manner as the even-mode line above a wide ground plane.

The attenuation constant was measured on a 50-ohm microstrip line by simple insertion method. The line length was 22 cm long and meandered on a 3-inch-long substrate as shown in Fig. 6. The attenuation constant was essentially uniform from 8 to 10 GHz and about 0.1 dB/cm. The substrate height was 25 mils from the ground plane and line thickness was ½ mil of copper. The thickness of the conductors and height of the substrate were kept the same for all of these experiments.

The conductor thickness was chosen for six-skin depth at 2 GHz, which is approximately 0.5 mil. The height of the substrate is a minimum for $h = 25$ to still obtain reasonably low-loss microwave circuits.

From a simplified expression of the attenuation constant, in which the losses are inversely proportional to the height of the substrate, a height of 50 mils should give proportional improvement in the quality of microwave circuits. At X-band, most linewidths are still much less than one-tenth of a wavelength.

The phase constant is more commonly expressed in terms of the guide wavelength, λ_g. To determine the guide wavelength, loosely coupled resonant rings of various linewidth were tested for the location of their absorption notch occurring at $n\lambda_g/2$. This method of finding the guide wavelength of the microstrip eliminates the end capacitance effect or the short circuit effect, and at the same time is applicable directly for any linewidth. The results are plotted in Fig. 7, curve I. The test sets are shown in Fig. 8. Since capacitance end-effect of the microstrip line is of great importance in matching network, filter, and resonator designs, the above experiment was repeated for semi-rings. The results are plotted in Fig. 7, curve II. A sample test fixture is shown in Fig. 9.

The latter curve may be used for half-wave resonators with two end-effects directly or for quarter resonators with one capacitive end-effect making separate adjustments of the resonator for the shorted end.

APPLICATIONS

The coupled line components are a direct application of the even and odd mode characteristic impedances.

As in plane waves, the coefficient of coupling for quarter wave resonators is equal to the difference over sum of these impedances. Now for any desired coefficient of coupling K, the separation may be found. The application of this is in filters and directional couplers primarily. The advantage of the coupled-lines directional coupler is that it requires very little space.

For input matching the following condition must be met:

$$Z_{oe} \cdot Z_{oo} = Z_o^2 \qquad (8)$$

The forward coupling coefficient for quarter wave coupled lines is:

$$K = \frac{(Z_{oe} - Z_{oo})}{(Z_{oe} + Z_{oo})}$$

Then:

$$Z_{oo} = \frac{Z_o}{\sqrt{\dfrac{1+K}{1-K}}}$$

$$Z_{oe} = \left(\frac{1+K}{1-K}\right) Z_{oo} \qquad (9)$$

For the condition that W is equal, the separation S is found from Fig. 4.

The photograph of Fig. 11 shows 3, 6, 10, and 20-dB directional couplers. The test results are in agreement with the theoretical values of the even and odd mode impedances. The couplers operate at X-band.

The design of odd-mode coupled band pass filters involves reducing the desired response shape to the coefficients of coupling and then to the even and odd mode impedances per coupled section. In this form the filter is represented in the universal set for odd mode coupled structures. By Fig. 4, these impedances are reduced to a line width and separation per coupled section for the specific microstrip structure. An example of this is shown in the photograph of Fig. 12.

Two coupled sections of this form of structure are one-half of a guide wavelength long. With the help of Fig. 7, the proper length including end-capacitance effect is obtained.

An application to a major component is illustrated in the photograph of Fig. 13. The device is a wideband, times-four frequency multiplier with output in the X-band. The test results on the prototype version are 10-dB insertion loss for 10% BW. Power level saturation is 13 mW of output.

CONCLUSION

The techniques described in determining the characteristics of the miniature microstrip transmission line have been found in good agreement with the experiments. The technique may be applied to other complex odd-mode coupled structures for which closed form solutions are not directly attainable.

The use of the expressions for the even- and odd-mode impedances are not restricted to the substrate used in these experiments. They should apply as well to thin-film monolithic construction of the microstrip line on silicon as many future microwave circuits will be constructed. The applications demonstrate that the odd-mode coupled structure for the high dielectric substrate combines amazing compactness with still remarkable performance. This new microwave technique will influence designs of communications equipment for the 1970's due to the enormous size reduction, potentially higher reliability and economy.

ACKNOWLEDGEMENT

The author wishes to thank D. S. Hall for working out the numerical example of Fig. 4. Special mention should be made of Drs. Nergaard and Pan for making their Research Laboratories in Princeton and New York available for the preparation of the experiments.

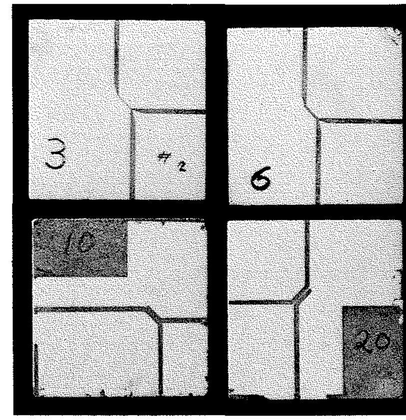


Fig. 11—Set of directional couplers at X-band.

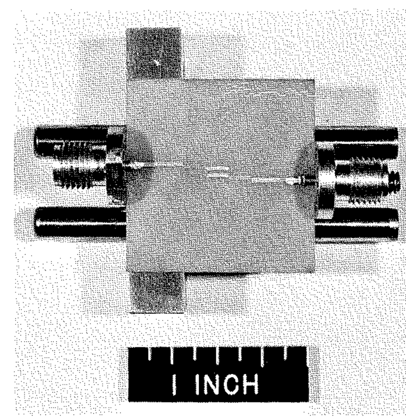


Fig. 12—Bandpass filter at X-band.

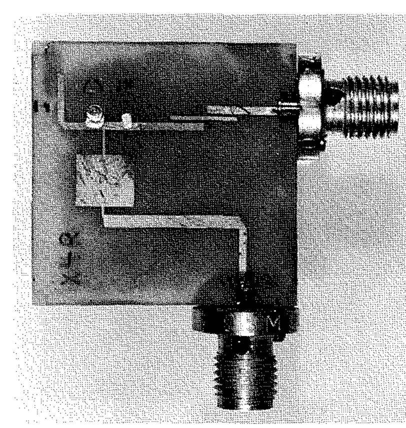


Fig. 13—Wideband frequency multiplier with output at X-band.

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MICROWAVE ADVANCED PRODUCT DEVELOPMENT

MICROWAVE DEVICES—ELECTRON TUBES VS SOLID-STATE

This paper discusses the relative attributes of solid-state and electron-tube devices and presents arguments to illustrate the continuing and growing requirement for electron tubes in spite of the inroads being made in microwave equipment applications by commercially available solid-state devices.

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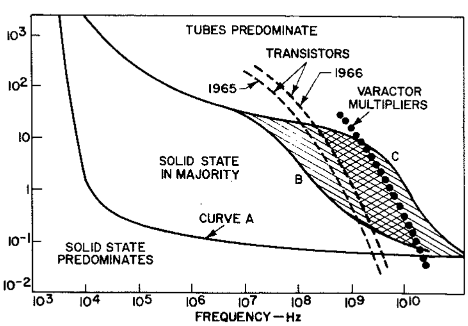


Fig. 1—Comparison of power-frequency capability of electron-tube and solid-state devices.

The progress made in the development of solid-state microwave devices during the past two years seriously challenges the supremacy of the electron tube in a number of areas where the capabilities of the two types of devices overlap. Most of the solid-state equipments operating today in the microwave region (above 500 MHz) owe their performance capabilities to either new transistor technology or the use of internal or external varactor-diode frequency multipliers. There is strong evidence, however, that suggests the continuing need for the electron tube in many applications both in the overlapping area and beyond this region.

AREAS OF ELECTRON-TUBE AND SOLID-STATE PREDOMINANCE

A chart of power as a function of frequency which specifically indicates the areas of contention is shown in Fig. 1. As indicated on the chart (taken from E. W. Herold[1]), solid-state devices have established marketing predominance in the region below curve *A*, a region formerly dominated by the older gridded electron tube. In this region, experience shows that there is little future need for electron tubes except for special applications.

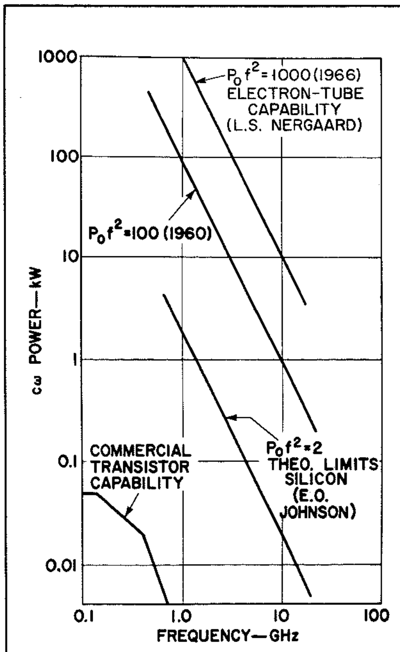
In the area bounded by curves *A* and *B*, solid-state devices are finding increased acceptability and are in majority use in the active-devices field. However, this situation does not imply the demise of the electron tube in this region. On the contrary, tubes are enjoying renewed usage in this area because the phenomenal and continued growth of color television has led to the establishment of reliable and economical circuits requiring the unique characteristics of the gridded tube.

Above curve *C*, electron tubes predominate because they can be designed to dissipate heat at the anode and other elements without the large thermal resistance experienced by solid-state devices. Solid-state devices suffer in this respect because of their crystalline-lattice interaction structure. The fact that the physical size of a device designed for handling high frequencies must be very small creates enormous heat-conduction problems for solid-state devices in the microwave region.

The region between curves *B* and *C* (specifically above 500 MHz) is of particular interest to the microwave engineer. Although tubes still predominate in this region, solid-state devices are continually making progress. In this region the functional requirements of the system and the relative capability of solid-state and electron-tube devices must be critically evaluated to determine the best component from a practical and economical viewpoint.

The two dashed lines within the area bounded by curves *B* and *C* indicate the limits of performance of a single commercially available transistor at the end of 1965 and as projected for the end of 1966. Continued effort in the development of transistors has led to commercially available types that provide an output of 10 watts at 400 MHz and 1 watt at 1 kHz when operated as straight-through amplifiers, and types that provide 3 watts output at 1 kHz when operated as amplifier triplers. Power beyond present transistor capability can be achieved by use of driven varactor multiplier chains. Chain performance limits are shown by a dotted line in Fig. 1.

Fig. 2—Theoretical power-frequency limits for tubes and transistors.



Final manuscript received January 9, 1967.

COMPARISON OF DEVICE CAPABILITIES AND LIMITATIONS

The transistor, as it is now conceived, has a very short interaction region that tends to prevent efficient energy transfer from the DC source to the RF output over wide bandwidths at high frequency. There are, on the other hand, many electron-tube devices that have long interaction regions (circuits), are not inhibited by transit-time difficulties and/or incorporate long beam drift-bunching paths. These characteristics virtually guarantee the superiority of such electron tubes over transistors in most applications where they apply. The following summary statements on various microwave electron tubes that meet the above criteria emphasize the outstanding performance and, perhaps, irreplaceability of these devices:

1) *The traveling-wave tube* performs outstandingly because of a long, nondispersive interaction circuit (30 to 40 wavelengths) that permits the realization of high gain (40 to 50 dB), extremely wide bandwidth, and very-high-frequency capabilities (to 100 GHz).
2) *The klystron amplifier*, which utilizes transit time, provides high gain with the use of multiple cavities (to 50 dB), high cw power (100 kilowatts at X band), high pulsed power, and good efficiencies.
3) *The reflex klystron*, an extremely versatile device, provides very-high-frequency performance (100 GHz) and good mechanically tuned bandwidth.
4) *The backward-wave oscillator* is unsurpassed for electron tuning over wide bandwidths (half an octave at 40 GHz, 20% at 75 GHz).
5) *The magnetron oscillator* is capable of 1-megawatt pulsed performance at X-band frequencies, inverted coaxial magnetrons are capable of 100 kilowatts at 30 GHz. Both magnetrons operate at high efficiency.
6) *Crossed-field amplifiers* are capable of 1 megawatt pulsed performance at 10 GHz (low gain, 8 to 16 dB) with high efficiency (approximately 50%).
7) *Pencil tubes,* which represent perhaps the lowest-cost microwave active devices, provide relatively high efficiency in cw amplifier and oscillator operation, and can be housed in low-cost cavities. Their excellent performance in pulsed power operation provides low-cost, small-size units that have good stability.

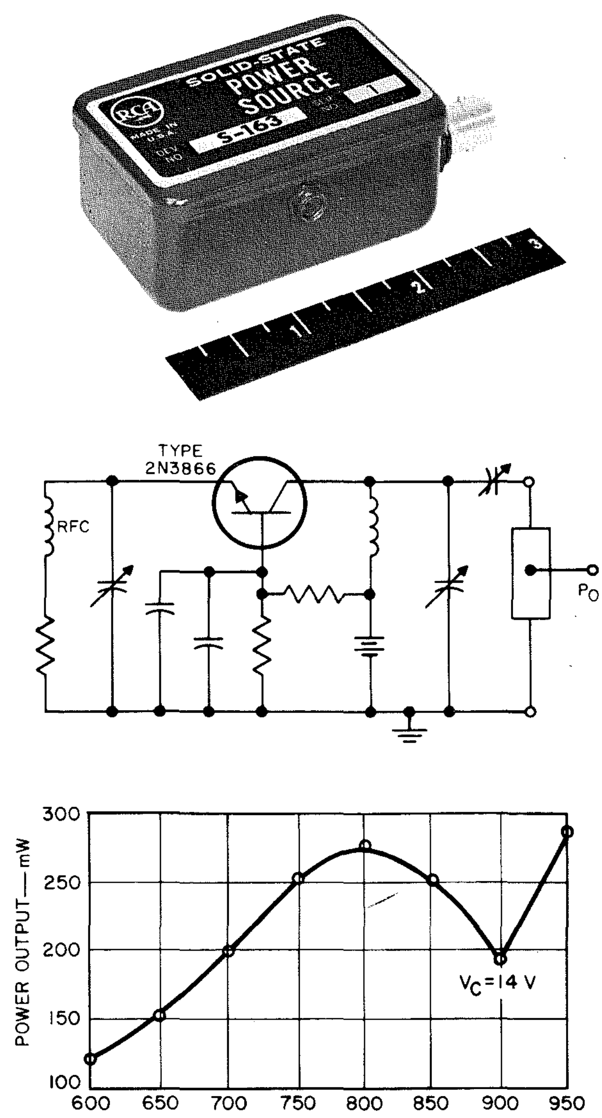


Fig. 3—Photograph (a) schematic diagram (b) and performance curve (c) for RCA S-163 solid-state power source.

H. J. WOLKSTEIN received his BSEE in 1953, and has completed his work toward an MSEE at the Newark College of Engineering. From 1948 to 1955, he worked in the Research Laboratories of National Union Electric Corporation. He joined the RCA Microwave Tube Operations Department in 1955. In 1958, he became Engineering Leader in charge of the development of low-noise traveling-wave tubes. He was promoted to Manager, Traveling-Wave-Tube Design and Development, in 1961. In July 1964 he was made Manager, Microwave Advanced Product Development. He now directs product development of traveling-wave products. Mr. Wolkstein has made many outstanding contributions to the design of traveling-wave tubes, periodic-permanent-magnet focusing structures, slow wave structures, and electron guns. He was particularly instrumental in the development of the RCA family of ultra-low-noise traveling-wave tubes, memory-storage tubes, and the RCA tricoupler and multi-coupler switching tubes. He has been awarded several patents in the electron-tube field and has written numerous papers on traveling-wave-tube design and application. He is a member of IEEE and the IEEE Groups on Electron Devices, and on Microwave Theory and Techniques.

TABLE I—Comparison of Commercially Available Power Sources (Frequency Range 0.5—2 GHz)

	Transistor Oscillator	Solid-State Crystal Oscillator Controlled Chain	Reflex Klystron	Pencil Tube Triode	Backward-Wave Oscillator
Temperature stability (PPM/°C):					
Uncompensated	100	1	5		10
Compensated	10	0.1 Oven controlled	0.5	10	
Voltage stability:					
Plate or collector ±5% Δ V	±1.2-1.5 MHz	Includes 5% Δ V	±7.5 MHz	±0.4 MHz	±35 MHz
Filament ±3% Δ V				±0.2 MHz	
Pulling figure (1/5/1 VSWR)	±2 MHz		±50 kGz	±2 MHz	± 1 MHz
Tuning:					
Mechanical: 0.5—1 GHz	Octave		30%		
1—2 GHz	30-50%				
Electrical: 0.5—1 GHz	12%				Octave
1—2 GHz	12%				
AM Sideband noise	Fair	Fair (Harmonic generator by itself—good)	Good (2-cavity klystron best)	Fair; comparable to uncompensated reflex klystron	Fair
FM noise (relative)	Poorest	Good (Lowest near carrier)	Fair—when unstabilized Good—when stabilized Best—2-cavity Klystron		Poor

The comparative differences in capabilities between semiconductor devices and electron tubes is shown more clearly by the chart of frequency as a function of power in Fig. 2. The RF power limitation for electron tubes, as developed by L. S. Nergaard,[2] can be approximately related to tube constants as follows:

$$P_{rf} \leq j^2 \frac{1}{\Delta\omega C} \lambda^2$$

where j is the RF emission-current-density limit of the cathode, ω is a term equal to 2π times the bandwidth BW, λ is the wavelength, and C is the shunt output capacitance per unit area. The limiting equation given by Nergaard, when compared to more recent actual attainments of electron tubes, yields the following limit:

$$Pf \leq 1{,}000$$

where the power P is in kilowatts and the frequency f is in GHz.

E. O. Johnson[3] has shown that the power/frequency capability of solid-state devices is limited by two important parameters that are characteristic of the semiconductor material, as follows:

$$(P_m X)^{1/2} f_T \leq \frac{E_{vs}}{2\pi}$$

where P_m is the power output, X is the reactive impedance at frequency f_T of the collector base impedance, and f_T is the transit-time cutoff frequency. This equation shows that the power/frequency limit of a semiconductor material is dependent on two factors:

1) E, the maximum electric field that the material can sustain without breakdown,
2) v_s, the maximum saturated carrier drift velocity.

As an example, for silicon $E = 2 \times 10^5$ volt/cm and $v_s = 6 \times 10^6$ cm/s; the equation then yields a material constant as follows:

$$(P_m X)^{1/2} f_T \leq 2 \times 10^{11} \text{ volts/s}$$

Sterzer[4] determined the limiting power capability for silicon devices using a 20-ohm base-to-collector reactance, as follows:

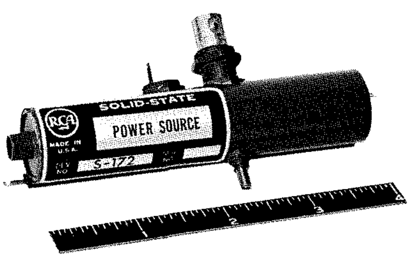
$$P_m f^2 \cong \frac{2 \times 10^{21}}{10^{21}} = 2$$

where the maximum power P_m is in kilowatts and the frequency f is in GHz.

This theoretical limiting relationship between power and frequency for silicon devices is shown in Fig. 2 for comparison with the actual attainment for both electron tubes and silicon transistor devices. Electron tubes are capable of providing performance almost three orders of magnitude better than the theoretical limit for silicon devices and five orders of magnitude better than the actual performance of any commercial silicon transistor device.

The survey of electron-tube and solid-state-device performance in the following

Fig. 4—Photograph (a) and diagram (b) of RCA S-172 coaxial-cavity power source.



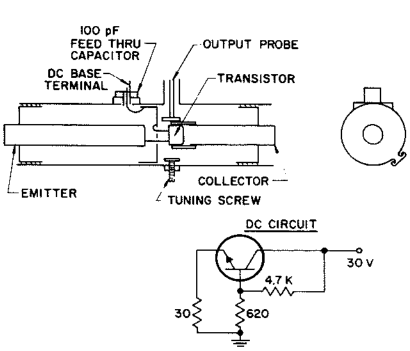


TABLE II—RCA Rocketsonde Transmitter Characteristics

	Solid-State Sonde	Pencil Tube and Cavity
Type	S-170V3 (Transistor Osc.-Mult.)	Dev. No. A-15497
Frequency	1680 MHz ±10 MHz	1680 MHz ±2 MHz
Power Output	200 mW (min.)	575 mW
Heater power	None	1 W
Operating voltage	12 V	100 V
B+ power	2.1 W	3 W
Efficiency, (η) typ.	10%	14.3% (typical, incl. heater power)
Pulse modulation (100%)	—3 V	20 V
Frequency modulation (0.5 MHz)	1 V p/p	3 V p/p, 200 kHz
Typical Environmental Requirements:		
Acceleration	200 g's, 15 s	
Shock	200 g's, 11 ms	
Temperature	—55°C to 70°C (S-170V3: 0-70°C)	
Altitude	300,000 feet	

section is confined to examination of some examples of each type of device in the region of overlapping capability pointed out in Fig. 1.

ELECTRON-TUBE AND SOLID-STATE-DEVICE PERFORMANCE IN THE MICROWAVE REGION

Various solid-state and electron-tube oscillators are available for use as power sources in the microwave region. These power sources can be divided into several different categories as a function of their frequency stability, bandwidth, tuning, and power-output level.

Table I indicates the comparative capability of some of the available power sources covering the UHF and L-band regions. As indicated, free-running solid-state CW sources are available with frequency stability and pulling figure comparable to those of pencil-tube triode oscillators provided adequate temperature compensation is used. The electron-beam reflex klystron, however, provides somewhat better performance if greater temperature stability, higher modulation linearity, and wide mechanical tuning are required.

When low AM and FM noise and good overall stability are required, the crystal-controlled solid-state multiplier chain with temperature compensation is unsurpassed. The backward-wave oscillator has little competition at frequencies above 2 GHz when extremely wideband (octave or half octave) electronic tuning is required.

RCA has produced a number of different solid-state power sources of the free-running variety, varactor chains for frequency multiplication, crystal-controlled oscillator chains, and various combinations of the devices. Some typical examples of these solid-state devices and, in special cases, their electron-tube counterparts are described below.

Tunable Solid-State Sources.

Several different versions of simple mechanically tuned fundamental oscillators have been used as signal sources for test equipment, multiplier chains, and telemetry applications. These devices are inherently compact and rugged, require no heater power, and operate at low voltage. These tunable solid-state signal sources, which take advantage of the performance features of the overlay transistor, operate in the P-band region (0.5 to 1 GHz) and provide more than 100 milliwatts of output power.

Fig. 3 shows a photograph, schematic, and performance curve for a lumped-circuit oscillator, RCA S-163. The circuit can be described as a Colpitts oscillator with feedback through the collector-to-emitter capacitance of the overlay transistor. The device is tuned by a single screw that varies the shunt capacitance in the emitter-to-ground circuit. The unit can be adjusted to provide minimum power of 0.5 watt over any 100-MHz portion of the P band or 100 milliwatts over any 300-MHz portion of the same band.[5]

RCA types S-171 and S-172 use coaxial cavities to supply the necessary resonance for collector and emitter tuning, and incorporate a DC resistance biasing network and a feedthrough capacitor (their only passive components) to enable operation. Fig. 4 shows a cross section and a circuit diagram of the S-172 signal source.

Provision for electronic tuning can be accomplished by variation of the bias applied to a varactor diode in the collector cavity of the device. The unit provides a power output of 200 milliwatts and has an FM capability of 16 MHz/volt. The S-171 power source is a similarly housed coaxial-cavity oscillator in which octave mechanical tuning is achieved by variation of the depth of a shorting plunger in the collector cavity. This device provides 100% tuning bandwidth and more than 100 milliwatts of output power.

Rocketsonde Transmitter Application.

The meteorological field is currently dominated by a low-cost high-volume RCA pencil-triode tube-and-coaxial-cavity design which has many outstanding features. Nevertheless, for certain applications requiring the use of sophisticated and novel circuitry and low operating voltages, a solid-state device is preferred even if it represents initially higher unit costs. Solid-state transmitters exhibiting the preferred characteristics and capable of multiple functions have been introduced and used recently in a meteorological weather sonde application.

Fig. 5 shows solid-state and pencil-tube transmitters[6] designed especially for rocket-launch applications. These rugged devices are capable of withstanding extremes in shock, vibration, and acceleration during operation, and incur negligible frequency and power changes. Table II compares the characteristics of these devices; the lower-cost pencil-tube

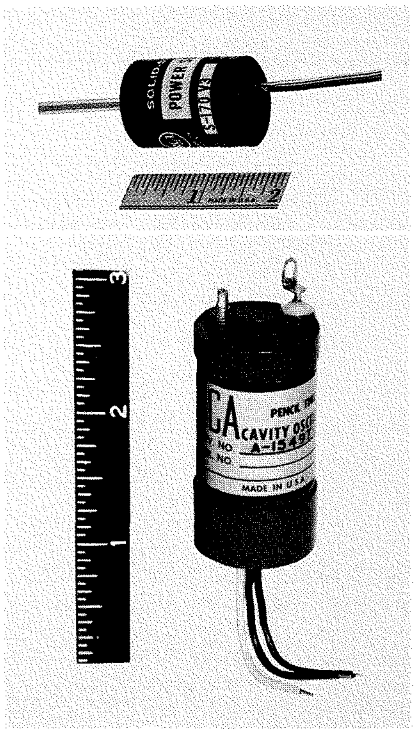


Fig. 5—Photographs of solid-state (a) and pencil-tube (b) rocketsonde transmitters.

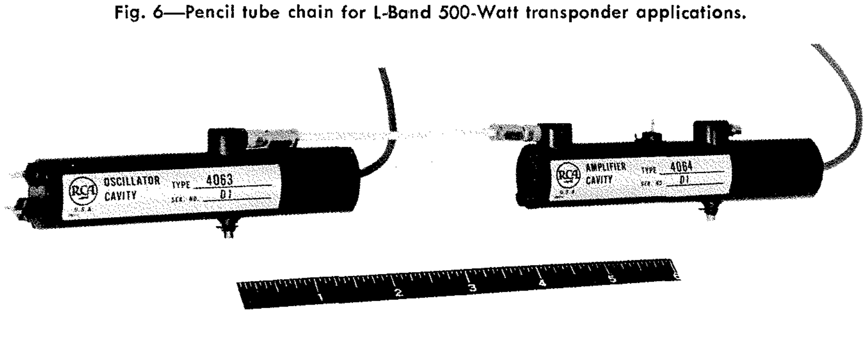
and cavity combination has a higher power output and transmitter efficiency in spite of the heater-power requirements. On the other hand, the higher voltage required for the pencil tube necessitates a larger battery supply.

Pencil-Tube Transponder Chain

Although solid-state devices are capable of providing cw performance comparable to that of pencil-tube-and-cavity combinations in most applications, at this time pencil tubes and cavities can be produced at a far lower cost than their solid-state counterparts. In addition, pencil tubes can greatly out-perform solid-state components in applications which require high peak pulsed power.

The low-cost coaxial pencil-triode tube is capable of reliable pulsed performance at peak power levels of 500 watts to 1 kilowatt for low-duty-cycle (less than 1%) applications. This capability in the L band is far in excess of the present state-of-the-art for transistors. Although

Fig. 6—Pencil tube chain for L-Band 500-Watt transponder applications.



bulk-effect devices, such as the Gunn transit-time diode, hold promise of achieving this performance, additional applied research will be required before a reliable solid-state device capable of operation at this power level and duty cycle is realized.

RCA has developed a two-tube ruggedized oscillator-amplifier chain for a new universal airborne transponder application (Air Traffic Control Radar Beacon System, IFF) that takes advantage of the pencil-tube peak power capability. This two-tube coaxial-cavity chain, shown in Fig. 6, consists of a grid-pulsed oscillator and a grounded-grid power amplifier operating at 1,090 MHz. The two-tube light-weight system, which operates at 500 watts with a duty cycle of 1%, provides good frequency stability ($< \pm 2\frac{1}{2}$-MHz deviation) with wide variations in load mismatch and temperature. The outstanding performance of the chain results from the output buffer-amplifier stage which decouples the tube load and its variations from the grid-pulsed oscillator. The chain operates over a temperature range from −54 to +95°C and can withstand shock in excess of 15 g's.

Varactor Multiplier Chains

Varactor frequency multipliers permit the extension of the operation of solid-state power sources well into the microwave region, as indicated in Fig. 1. These devices operate by virtue of a nonlinear capacitance variation as a function of RF drive level when they are operated in the reverse-bias region. Such capacitance variations create harmonic currents of the input frequency. Operation of the varactor as a frequency multiplier in conjunction with a suitable microwave driving source has resulted in the generation of harmonics well into the millimeter region. Varactor-driven sources challenge the use of klystron sources for low-power devices in local-oscillator, telemetry, and transmitter applications below the Ka band.

The cutoff frequency (f_c) of varactor diode may be used to determine the conversion efficiency (η) of the device as a function of the operating frequency (f_{in}), as follows:

$$f_c = \frac{1}{2\pi\, r_s\, C_{min}}$$

$$\eta = 1 - D\left(\frac{fin}{fc}\right)$$

where r_s is the equivalent diode series resistance, C_{min} is the diode junction capacitance at −6 volts, and D is a constant equal to 20.8 for a doubler and 35 for a tripler.

The maximum input power to a varactor diode (P_{in}) is limited by the breakdown voltage of the device (V_b) and/or the maximum tolerable temperature rise due to dissipative losses, as follows:

$$P_{in\ (max)} = k\, V_b^2\, f_{in}\, C_{min}$$

where k is a constant.

In general, the varactor diode is used at multiple frequencies no greater than the fourth harmonic if power, efficiency, and bandwidth are important parameters. This practical limitation is caused in part by the necessity for providing idler resonant circuits at most included harmonic frequencies as well as resonant circuits at the input and output frequencies (a rather complex requirement).

Harmonic multipliers have been built and produced in single varactor multiplier units as well as in chain multiplier configurations with several cascaded varactors housed in the same chassis. These devices are completely passive in that no DC voltage is supplied to the device. Generally, self bias, derived from the operating RF signal, is obtained by each varactor diode. This biasing stabilizes the stage as a function of drive power and temperature and helps to reduce the generation of parasitics.

Varactor multiplier chains have also been designed with various driver-input power sources. These power sources are normally of the transistor-oscillator, free-running, or crystal-controlled variety. Some of the driver circuits include transistor amplifiers that increase power at the low-frequency drive input level required by the multiplier chain.

The TOVAR (transistor oscillator-varactor) shown in Fig. 7 is an interesting compact self-contained power source for airborne telemetry applications in fixed-frequency service over the 1.4-to-2.3-GHz frequency bands. This device makes use of a transistor oscillator, a self-multiplier stage, and a varactor-diode frequency-doubler stage. The unit shown, which has provisions for FM modulation, has a nominal operating efficiency of 12% with power output ranging from 1 watt at 1.4 GHz to 0.5 watt at 2.3 GHz[5]. The overlay transistor is employed as a free-running lumped-circuit oscillator tuned to f_{osc}, one-fourth the final output frequency. The second harmonic of this frequency ($2f_{osc}$), generated by the internal varactor capacitance of the transistor, is taken from the collector to excite the foreshortened half-wave bar-coupler. The bar-coupler is used to excite the varactor for frequency doubling so that the required output frequency of $4f_{osc}$ results. The bar-coupler also provides some filtering of the fundamental and second harmonics of the oscillator frequency. The FM modulation is obtained by injection of a control voltage into the base circuit of the oscillator. This voltage changes the base-collector capacitance and thereby varies the operating frequency of the device. The circuit also contains provisions for making the FM characteristics linear and for temperature stabilization.

Harmonic-Generator Chain

Although electron-beam devices such as the klystron are already in use as the microwave transmitter sources in airborne and satellite applications, the need for efficient, light-weight, reliable devices with low FM and AM noise levels and good temperature stability has encouraged the development of varactor or harmonic-generator chains for these applications. One of the more interesting transmitter varactor chains was designed to meet extremely severe environmental conditions of shock and vibration with a resultant minimum production of microphonics and AM and FM noise.

A block diagram of a transmitter varactor chain is shown in Fig. 8. The same varactor chain (with variations) is used for velocity sensing, altimeter, and radar transmitter functions; the crystal-oscillator portion of the chain, however, is used only for velocity sensing. Because the input signal to the

TABLE III—Comparison of Low-Noise Amplifier Devices

	Noise Figure	Gain (dB)	Saturated Power Output dBm	Bandwidth	Magnetic Field
Traveling-Wave-Tube Amplifier					
0.5- 8 GHz	12	35	10-20	Octave	PPM
	10	30	10	Octave	Reversed
	6	25	0	Octave	Uniform
0.7- 4 GHz	4.5	25	0	500 MHz	Uniform
8-12 GHz	8	25	0	½ octave	Uniform
*Tunnel-Diode Amplifier (Ge)**					
1- 2 GHz	4.5	10	—25#	500 MHz	Circulator
2- 8 GHz	4.0	15	—25	250 MHz	Circulator
8-12 GHz	5.5	10	—25	1 GHz	Circulator
Transistor Amplifier					
0.5- 1 GHz	5	7-5/stage	0	Octave	None
1- 2 GHz	7	5/stage	0	Octave	None
1.7-2.7 GHz	9	3/stage	0	1 GHz	None

*GaSb tunnel-diode amplifier has ≈ 1 dB lower noise figure than indicated for Ge diode.
#This value is the power output for 1-dB compression.

transmitter is of such low frequency (100 MHz) and small amplitude (15 milliwatts), several stages of power amplification are required to obtain adequate drive power for the varactor chain. The transmitter output power is approximately 350 milliwatts at 9.6 GHz with an input frequency of 100 MHz.

The outstanding performance characteristic of the Fig. 8 varactor-multiplier chain is its low AM and FM noise spectrum close to the carrier. The performance obtained can be matched only with a stabilized two-cavity klystron.

Low-Noise Amplifier Devices

Table III compares the performance of low-noise receiver amplifier devices. The following general conclusions can be drawn from the data shown.

1) *Transistors*—Transistor amplifiers are capable of replacing tunnel-diode amplifiers and traveling wave tubes below 2 GHz provided power levels of less than 1 milliwatt are required for octave bandwidth performance. Moreover, the transistor amplifier requires no circulator and provides a noise figure comparable (within 1 or 2 dB) to that of the tunnel-diode amplifier with 20 dB more dynamic range.
2) *Tunnel-Diode Amplifiers.* Tunnel-diode amplifiers continue to be extremely attractive for application above 2 GHz when bandwidth of 10 to 20% is needed and power output of —25 dBm with limited linear drive range can be tolerated. However, the tunnel diode must be operated in a limited temperature environment because of its temperature sensitivity, and is susceptible to burnout in the presence of transient signals unless overload protection is provided.
3) *Traveling-Wave Tubes.* The traveling-wave tube with uniform field focusing capabilities provides a noise figure better than that of the transistor amplifier and almost comparable to that of the germanium tunnel diode over a 500-MHz bandwidth below 4 GHz. In addition, the traveling-wave tube operates with a gain considerably greater than that of either tunnel diodes or transistor amplifiers. When wideband performance up to 12 GHz is required, the traveling-wave-tube gain-bandwidth product and saturation power output provide considerable advantage over the tunnel-diode amplifier. However, the traveling-wave tube suffers because of its weight, size, and power-supply requirements as compared to the requirements of a tunnel-diode amplifier. Although transistorized amplifiers are still being improved, neither these devices nor tunnel-diode amplifiers can compete with the traveling-wave tube for applications requiring power levels over 10 milliwatts above 1 GHz.

Pulsed Operation

Solid-state devices generally do not have the capability to produce pulsed peak power greatly in excess of their CW power performance capability, primarily because of the limitations associated with breakdown voltage. For pulsed operation, therefore, there is a significant and glaring difference between the performance of electron tubes and solid-state devices. As an indication of the capability of some electron-beam devices in regard to peak power capability and bandwidth, Yocom[9] has presented a detailed comparison. As shown in Fig. 9, electron-beam amplifiers such as traveling-wave tubes, klystrons, and crossed-field amplifiers are capable of operating to 100 megawatts in the microwave region. This capability is an additional two orders of magnitude greater than the CW performance obtainable from similar electron tubes at S band.

THE CONTINUING NEED FOR THE ELECTRON TUBE

The theoretical and practical limitations cited for solid-state devices and the superiority of electron tubes in certain areas insure that the electron tube is here to stay! Even within the practical limits of frequency and power set for solid-state devices, there are many specific and unique applications now using electron-tube devices for which no solid-state replacement readily exists. For example, the cathode-ray picture tube, a relatively low-frequency device, still has no practical solid-state counterpart in spite of its high volume demand.

There are, therefore, innumerable and compelling reasons for believing that electron tubes will not only remain in extensive use, but continue to find new and expanded usage as electronic systems become more sophisticated and as tube device utility is further enhanced through advanced design.

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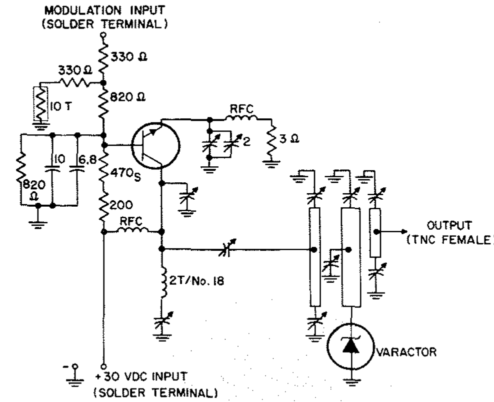

Fig. 7—Schematic diagram of RCA-S127V2 solid-state power source containing a transistor oscillator and a varactor.

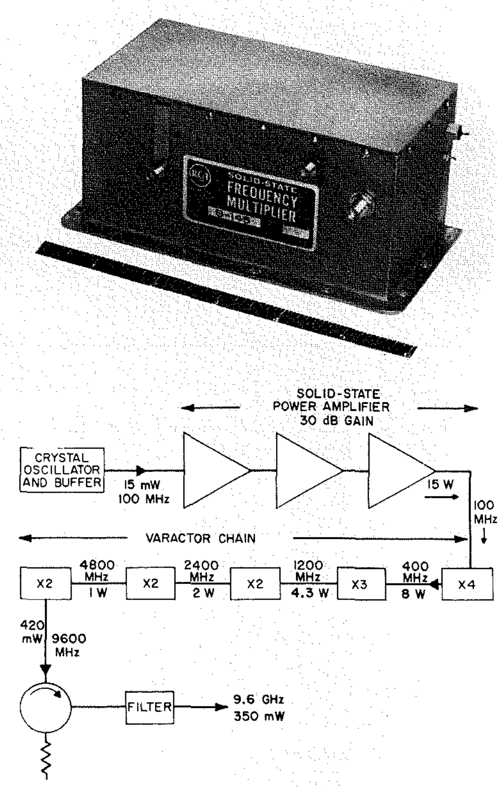

Fig. 8—Photograph (a) and block diagram (b) of a solid-state spacecraft transmitter.

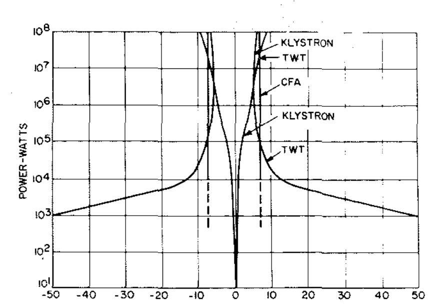

Fig. 9—Relative pulsed-power capability of various electron-beam amplifiers as a function of bandwidth and power level.

ALL-PASS NETWORKS FOR COMMUNICATIONS SYSTEMS

All-pass networks are often used to compensate for delay distortion in communications systems. Existing methods for synthesizing all-pass networks have important practical limitations. A new approach to this design problem is described here. The algorithm which results is applicable to a broad class of practical applications. Results for several problems which were solved by use of an RCA 601 computer program based on the new method are included.

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THE ability to design a network with a prespecified delay characteristic and constant amplitude response is necessary for success in many engineering applications. Such networks, known as *all-pass networks*, are often required to compensate for phase distortion introduced by filters used to provide a particular amplitude response.

Fig. 1 illustrates such a situation which occurred in the development of playback circuitry for a digital tape recording system. The desired pulse shaping network was to have an amplitude response which emphasized the high-frequency components and a constant delay characteristic from 0 to 750 kHz. As seen in Fig. 1a, the filter used to achieve this amplitude response introduced considerable delay distortion. The effect of this distortion can be seen in Fig. 1b by comparing the pulse before and after delay equalization. These output waveforms were obtained by using a digital computer to solve the network equations for the filter and delay equalizer.

A more striking example of the effect of delay equalization is shown in Fig. 2. These oscilloscope patterns were recorded during testing of a biphase signal channel.[1] The waveforms result from applying a periodic stream of binary information to the input of the signal channel. Without equalization, it is obvious that the information is hopelessly scrambled. With equalization, however, there exists a practical sampling interval throughout which *1*'s and *0*'s have distinct polarities.

For both these applications, all-pass networks capable of achieving the necessary equalization were determined by a computer program based upon the method described later in this paper. This method can be used to design both low-pass and band-pass equalizers and has been used in a variety of applications within RCA.

Final manuscript received November 18, 1966.

ALL-PASS NETWORKS

The transfer function for a first degree all-pass section has the form:

$$T_1(s) \triangleq \frac{V_I(s)}{V_O(s)} = \frac{a+s}{a-s} \qquad (1)$$

where V_I and V_O denote input and output voltages, respectively. For a second degree all-pass section, the transfer function is:

$$T_2(s) = \frac{s^2+bs+c}{s^2-bs+c} \qquad (2)$$

Fig. 3 shows lattice networks corresponding to these two cases. Unbalanced equivalent circuits are also available[2] and are much used in practice. No more complicated networks than those of Fig. 3 need be considered directly since any all-pass network can be represented by networks of these two types connected in tandem.[3]

The parameters a, b, and c of Equations 1 and 2 are directly related to the network poles. A first-degree section has a single real pole, $\sigma = a > 0$. A second-degree section may have two real poles $p_1, p_2 > 0$, or a pair of conjugate complex poles $p_1, p_2 = \bar{\sigma} \pm i\,\bar{\omega}, \bar{\sigma}, \bar{\omega} > 0$. In either case, $b = p_1 + p_2$ and $c = p_1 p_2$.

Phase functions, $\beta_1(\omega)$ and $\beta_2(\omega)$, are defined for first and second degree sections, respectively, by use of the expression:

$$T(i\omega) = |\,T(i\omega)\,| \exp[i\beta(\omega)] = \exp[i\beta(\omega)]$$

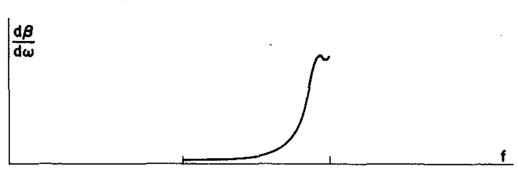
It follows that:

$$\beta_1(\omega) = 2\tan^{-1}\left(\frac{\omega}{a}\right) \qquad (3)$$

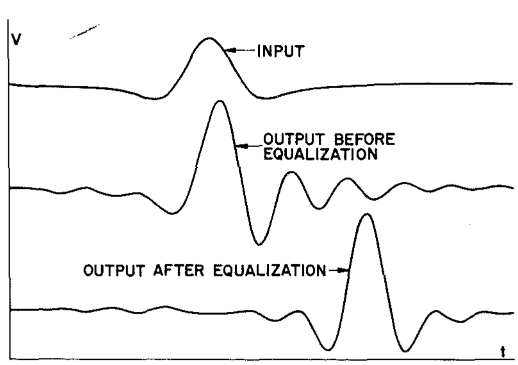
$$\beta_2(\omega) = 2\tan^{-1}\left(\frac{b\omega}{c-\omega^2}\right) \qquad (4)$$

There is considerable disagreement among practicing engineers as to whether specifications for circuit performance should be phrased in terms of phase delay (β/ω) or envelope delay

Fig. 1—Delay equalization for digital tape recording.

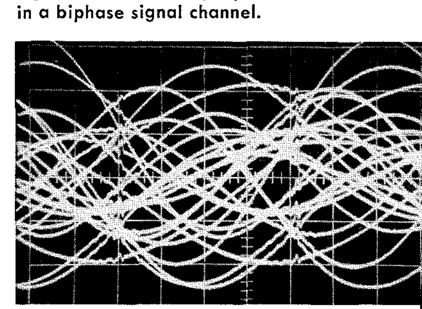


(a) ENVELOPE DELAY BEFORE EQUALIZATION

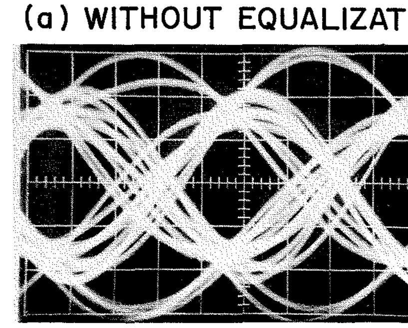


(b) EFFECT OF DELAY DISTORTION AND EQUALIZATION

Fig. 2—Effect of delay equalization in a biphase signal channel.



(a) WITHOUT EQUALIZATI



(b) WITH EQUALIZATIO

DR. ROGER L. CRANE received his BSEE from Iowa State University in 1956. From 1956-1959 he worked for the Univac division of the Sperry Rand Corporation in St. Paul, Minnesota as a digital computer circuit and logic design engineer. He returned to Iowa State University and received his MS and PhD degrees in mathematics in 1961 and 1962, respectively. He was a Member of the Technical Staff of RCA Laboratories at the David Sarnoff Research Center during the summers of 1960 and 1961. Dr. Crane joined the Applied Mathematics Research Group at RCA Laboratories in 1963. He is a member of SIAM, ACM, Phi Kappa Phi, Tau Beta Pi, Eta Kappa Nu, Pi Mu Epsilon, and Sigma Xi.

$(d\beta/d\omega)$. Interest at RCA Laboratories in developing effective means for designing delay equalizers was spurred, initially, by a color television application.[4] In this case, a specification in terms of envelope delay was given by the NTSC. The program discussed here provides for equalization in terms of envelope delay. It has, in addition, provided excellent results in several applications where a given tolerance on phase delay was to be achieved.

By differentiation of Equation 3, the envelope delay corresponding to a first degree section with real pole σ is:

$$\frac{d\beta_1}{d\omega} = \frac{2\sigma}{\sigma^2 + \omega^2} \qquad (5)$$

For a second-degree section with two real poles, differentiation of Equation 4 yields a sum of two terms of form shown in Equation 5. When the second degree section has a pair of complex conjugate poles, $\bar{\sigma} \pm i\bar{\omega}$, it follows after differentiation and some manipulation that:

$$\frac{d\beta_2}{d\omega} = 2\left[\frac{\bar{\sigma}}{\bar{\sigma}^2 + (\omega - \bar{\omega})^2} + \frac{\bar{\sigma}}{\bar{\sigma}^2 + (\omega + \bar{\omega})^2}\right]. \qquad (6)$$

The delays of cascaded first and second degree all-pass sections add linearly and thus, the envelope delay of an all-pass network with M real and N pairs of conjugate complex poles can be written as:

$$D(f) = \frac{1}{\pi}\sum_{i=1}^{M}\left[\frac{\nu_i}{\nu_i^2 + f^2}\right] + \frac{1}{\pi}\sum_{i=1}^{N}\left[\frac{\alpha_i}{\alpha_i^2 + (f - f_i)^2} + \frac{\alpha_i}{\alpha_i^2 + (f + f_i)^2}\right] \qquad (7)$$

where ν_i, $\alpha_i = \sigma_i/2\pi$ and $f_i = \omega_i/2\pi$. The problem of equalizer design considered here is that of choosing, for fixed M and N, the poles, ν_i, (α_i, f_i) so that Equation 7 approximates a desired delay.

Existing design procedures have important limitations. Some are restricted to particular applications. For instance, if the envelope delay to be achieved is approximately parabolic and if one or two all-pass sections will give sufficient accuracy, then tables and graphs can be used to obtain a design.[5,6] Some results are also tabulated for linear group delays.[7,8] An algorithm developed earlier at RCA Laboratories[9] is limited to equalization of low-pass filters only. Potential analog techniques[10,11] are applicable to general problems, but they require special equipment, the use of trial and error techniques, and may not provide adequate accuracy. (Ref. 11, p. 280). The method described below was developed to utilize a digital computer in providing a more general, accurate, and practical means for equalizer design.

MINIMAX APPROXIMATION

Let $\bar{D}(f)$ denote the envelope delay function to be approximated and define the relative delay function:

$$R(f) = D(f) + K \qquad (8)$$

where $D(f)$ is given by Equation 7 and K is constant. The approximation of $\bar{D}(f)$ will be made by use of Equation 8 rather than Equation 7, since addition of a constant delay is not critical in most applications.

Some criterion of approximation must be selected. The particular choice made here is motivated by the following results that are available from the theory of minimax approximation by means of rational functions, i.e. quotients of polynomials. Let $R(x) = P(x)/Q(x)$, where $P(x)$ is of degree p and $Q(x)$ is of degree q, and assume that $f(x)$ is to be approximated for $a \leqq x \leqq b$. It is known[12] that there exists a unique choice of P and Q which minimizes the maximum absolute error for $a \leqq x \leqq b$. Furthermore, the error function $E(x) = R(x) - f(x)$ corresponding to this choice is an equal ripple function. More precisely, there exists a set of at least $(p + q + 2)$ points $a \leqq x_1 < x_2 < \cdots < x_{p+q+2} \leqq b$ and a number ε such that $(-1)^i E(x_i) = \varepsilon$, and for $x \neq x_i$, $|E(x)| \leqq \varepsilon$. In the design of algorithms for determining the coefficients of the desired P and Q, it is commonly assumed that there exist exactly $(p + q + 2)$ extreme points on the error curve. Some algorithms work with the $(p + q + 2)$ extremes of the error function and others with its $(p + q + 1)$ zeros. Working with the zeros implies fitting $f(x)$ exactly on a set of points whose number equals the number of coefficients to be determined in $P(x)$ and $Q(x)$, since the specification of one coefficient corresponds to an arbitrary normalization.

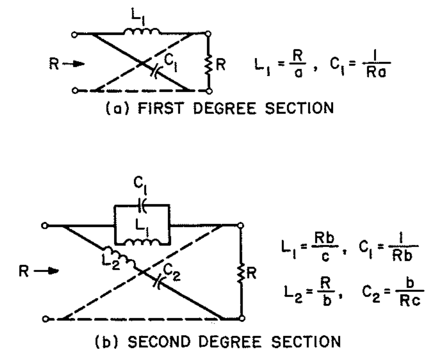


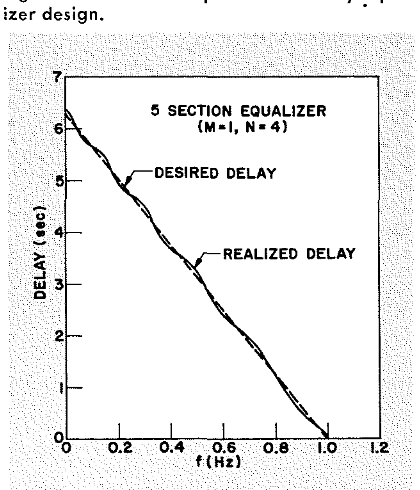
Fig. 3—All-pass lattice networks.

By obtaining a common denominator, Equation 8 can be seen to be a rational function of frequency with non-linear constraints among the coefficients. No theoretical results analogous to those stated above have been proved for this case of constrained rational approximation. However, the known results about the unconstrained approximation problem suggest that the equalizer design problem might be approached in two steps. The first of these steps consists of determining an approximation of Equation 8 which coincides with the desired delay on a specified set of points. The second step consists of perturbing these points in an attempt to find an equal ripple solution.

ALGORITHM FOR EQUALIZER DESIGN

There are $(M + 2N + 1)$ parameters, ν_i, α_i, f_i, and K to be determined in Equation 8. Hence, assume that a set of $(M + 2N + 1)$ frequencies $f_{min} < f_1 < f_2 <$

Fig. 4—Results of low-pass linear delay equalizer design.



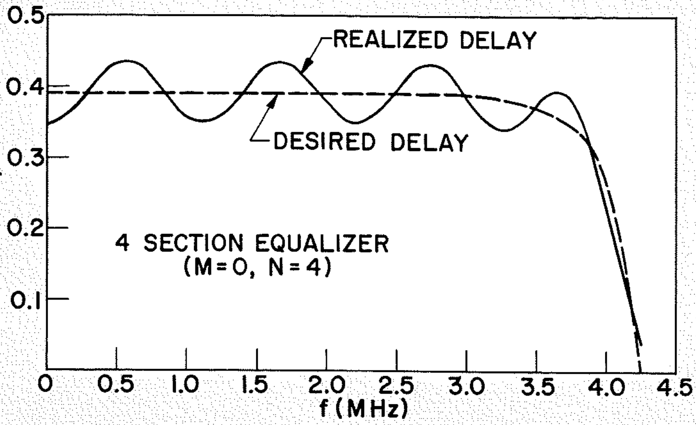


Fig. 5—Results of low-pass delay equalizer design.

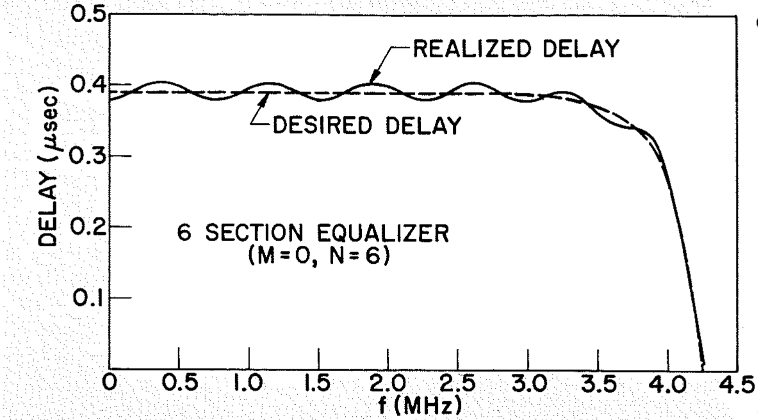


Fig. 6—Results of low-pass delay equalizer design.

$\cdots < f_{M+2N+1} < f_{max}$ are given and let $D_k \triangleq D(f_k)$ and $\bar{D}_k \triangleq \bar{D}(f_k)$. Thus, $D(f)$ must be determined so that:

$$R(f_k) = D_k + K = \bar{D}_k \qquad k = 1, 2, \cdots, M + 2N + 1 \tag{9}$$

Since K enters linearly, it can easily be eliminated, for instance by defining:

$$G_k \triangleq D_k - D_{M+2N+1} + \bar{D}_{M+2N+1} - \bar{D}_k = 0, \qquad k = 1, 2, \cdots, M + 2N \tag{10}$$

A solution of this set of nonlinear equations determines a delay equalizer which matches the desired envelope delay exactly at $(M + 2N + 1)$ points. Newton's method for the solution of nonlinear equations is known to converge quadratically in the neighborhood of a solution.[13] However, finding a solution "in the neighborhood" with which to start the iterative process is often difficult and sometimes impossible. A means for alleviating this problem was devised for the system of Equations 10.

Choose M values of ν_i and N pairs (α_i, f_i), and let $D^{(0)}(f)$ denote the envelope delay corresponding to an all-pass network with these poles. Choose an integer n and define:

$$\bar{D}^{(i)}(f) = D^{(0)}(f) + \frac{i}{n}[\bar{D}(f) - D^{(0)}(f)], \qquad i = 1, 2, \cdots, n \tag{11}$$

Now, consider a sequence of n problems:

$$G_k^{(i)} \triangleq D_k^{(i)} - D_{M+2N+1}^{(i)} + \bar{D}_{M+2N+1}^{(i)} - \bar{D}_k^{(i)} = 0, \qquad i = 1, 2, \cdots, n \qquad k = 1, 2, \cdots, M + 2N \tag{12}$$

One feels, intuitively, that a solution of the system $G_k^{(i)} = 0$ lies in a neighborhood of a solution of $G_k^{(i+1)} = 0$ and that successive application of Newton's method to the sequence of Equations 12 will lead to a solution of Equations 10. This has proved to be true in practice. The computer program that was coded provides for automatically modifying the number of steps in the sequence of Equations 12, increasing the number when necessary and decreasing the number if possible. At the option of the circuit designer who uses the program, $D^{(0)}(f)$ is provided by the program. The function provided, which depends only on the number of sections and the frequency range of interest, is satisfactory in most cases. Provision is also made for the designer to supply his own initial choice of the network poles from which $D^{(0)}(f)$ is computed. This option is often useful for designs which require a large number of sections. In these cases, good initial approximations can usually be easily inferred from the results obtained for fewer sections. Fewer steps are then required in the sequence of Equations 12.

The algorithm just described produces an equalizer design which matches the desired delay at certain frequencies. However, large deviations may occur between fitting points and the next step is to design a procedure for adjusting these fitting points so that the maximum absolute errors between fitting points are equal.

Again, a sequence of problems like those defined by Equations 12 is introduced. This time, however, the fitting frequencies are changed at each step in the sequence and the function to be fitted, $\bar{D}(f)$, remains fixed. The changes from step-to-step are small, and Newton's method is again used for the solution of the nonlinear equations.

Let $f_k^{(i)}$, $k = 1, 2, \cdots, M + 2N + 1$ denote the fitting frequencies and $R^{(i)}(f)$ the solution for the ith iteration. Define:

$$E^{(i)}(f) = |R^{(i)}(f) - \bar{D}(f)| \tag{13}$$

and choose $\bar{f}_k^{(i)}$ so that:

$$\varepsilon_k^{(i)} \triangleq E^{(i)}(\bar{f}_k^{(i)}) = \max_{f_{k-1}^{(i)} \leq f \leq f_k^{(i)}} [E^{(i)}(f)]$$

$$k = 1, 2, \cdots, M + 2N + 2$$

where $f_0^{(i)} = f_{min}$ and $f_{M+2N+2}^{(i)} = f_{max}$. Fitting frequencies for the next iteration are then determined by:

$$f_k^{(i+1)} = f_k^{(i)} + \alpha^{(i)} \cdot \left[\frac{(\bar{f}_{k+1}^{(i)} - \bar{f}_k^{(i)})\,(\varepsilon_{k+1}^{(i)} - \varepsilon_k^{(i)})}{(\varepsilon_{k+1}^{(i)} + \varepsilon_k^{(i)})}\right] \tag{14}$$

where the formula for $\alpha^{(i)}$ is given below. Equation 14, with $\alpha^{(i)} = 1/2$, can be obtained by considering only two adjacent extremes, assuming them to be of opposite sign and joined by a straight line which crosses zero at $f_k^{(i)}$, and then determining $f_k^{(i+1)}$ so that $\varepsilon_{k+1}^{(i+1)} = \varepsilon_k^{(i+1)}$. This vastly oversimplified model was chosen because it is easy to implement and has the desirable property of expressing the new fitting points as perturbations of the old. The factor $\alpha^{(i)}$ was introduced to control the magnitude of these perturbations, allowing larger changes when the approximation is far from equal ripple and smaller changes as the desired solution is approached. Specifically, $\alpha^{(i)}$ is determined by:

$$\alpha^{(i)} = \alpha_2 + (\alpha_1 - \alpha_2)\,\varepsilon^{(i)} \tag{15}$$

where:

$$\varepsilon^{(i)} = \frac{\min_k [\varepsilon_k^{(i)}]}{\max_k [\varepsilon_k^{(i)}]}, \tag{16}$$

and α_1 and α_2 are experimentally determined constants currently chosen as 0.4 and 0.8, respectively. When Equation 16 becomes sufficiently close to one, the iterative adjustment of the fitting frequencies is terminated.

In the preceding paragraphs, it was assumed that the sequence of problems defined by Equations 11 and 12 is to be solved before any adjustment of the frequencies is made. The initial fitting points are taken to be the zeros of the Chebyshev polynomial of degree $(M + 2N + 1)$, adjusted to the frequency range of interest. This choice is somewhat arbitrary and, hence, one adjustment of the fitting frequencies is made at each step of the sequence defined by Equations 11 and 12. Upon completion of this sequence, the frequencies are adjusted until Equation 16 is sufficiently close to one. The solution thus obtained corresponds to an all-pass network whose delay coincides with the desired curve on at least $(M + 2N + 1)$ points. The maximum deviations which occur between these fitting points are the same, within the tolerance specified in terminating the adjustment of the fitting frequencies. If less error is required, the number of all-pass sections may be increased.

EXAMPLES

The effectiveness of the algorithm outlined above can be seen by considering some typical practical applications. Ward[8] used both an electrolytic tank and a digital computer to produce an equalizer to approximate the function $\bar{D}(f) = 6.3\ (1 - f)$ for $0 \leqq f \leqq 1$. The error function for his six-section design (1 first- and 5 second-degree sections) is not equal ripple and according to the graph presented, has peak deviations of ± 0.15 second. A five-section design (1 first- and 4 second-degree sections) determined by the new algorithm had an equal ripple error of ± 0.1 second, as shown in Fig. 4. Thus, a smaller error was achieved with one less section. Running time for the RCA 601 FORTRAN program was about two minutes for this problem. When six sections were used, the equal ripple error of the network determined by the new algorithm was ± 0.03 second, down by a factor of about five from Ward's design. Poles for the two six-section designs are listed in Table I.

The new method has found its widest use, thus far, in the design of equalizers for color TV broadcast transmitters. Figs. 5 and 6 show results for a typical problem encountered. Fig. 5 is included to illustrate the value of knowing, under the assumptions which underlie the algorithm, whether a given error tolerance can be achieved with a certain number of sections. Previous work[4,9] resulted in four-section designs for equalizing a specified delay curve for a broadcast transmitter. Recent measurements on a particular transmitter yielded a delay curve similar to that used earlier,[4,9] but with a somewhat greater total excursion and a more rapid rate of change near 4 MHz. After laboratory experimentation had failed to produce a satisfactory four-section design, the new algorithm was employed. The results of Fig. 5 quickly showed that a four-section design was not adequate and the six-section design of Fig. 6 was then produced.

TABLE I—Comparison of Poles for Two All-Pass Equalizers

Design by Ward [8]		Design by new algorithm	
α	f	α	f
0.119	0	0.104	0
0.155	0.132	0.123	0.121
0.200	0.219	0.138	0.254
0.181	0.371	0.157	0.406
0.200	0.505	0.186	0.584
0.290	0.774	0.233	0.808

Band-pass equalizers can also be designed with the new algorithm and Figs. 7 and 8 show results for one problem of this type. The delay curve to be approximated is considerably more complicated than those of the previous examples. Fig. 7 illustrates an interesting property of the algorithm as currently programmed. As explained previously, the fitting frequencies are perturbed until the maximum absolute errors between fitting points are nearly equal. Although the perturbation method was derived under the assumption that adjacent peak errors are of opposite sign, no such check is made in the program and, indeed, the error shown in Fig. 7 does not have this property. The solution is still of practical value however, but one should expect more accumulated phase error across the band than would result in the case of alternating peak errors as illustrated in Fig. 8 for a four-section equalizer.

CONCLUSIONS

A new method for designing all-pass networks has been developed and programmed for the RCA 601 computer at RCA Laboratories. This program provides a much more effective tool for delay equalizer synthesis than was previously available to RCA circuit designers. Engineers from several divisions of RCA have used the program effectively to obtain equalizer designs for a variety of applications.

ACKNOWLEDGMENTS

The author wishes to thank Dr. R. W. Klopfenstein for many helpful suggestions during the development of these new techniques, Dr. G. L. Fredendall for his continued interest in the application of these methods to problems throughout RCA, and N. L. Gordon for supporting and encouraging this type of work within the Applied Mathematics Research Group.

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Fig. 7—Results of band-pass delay equalizer design.

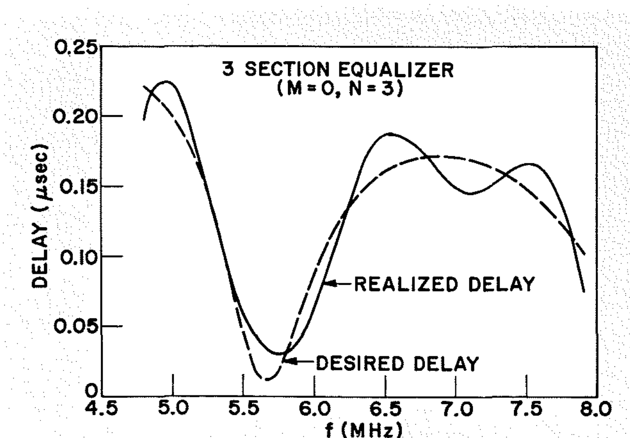
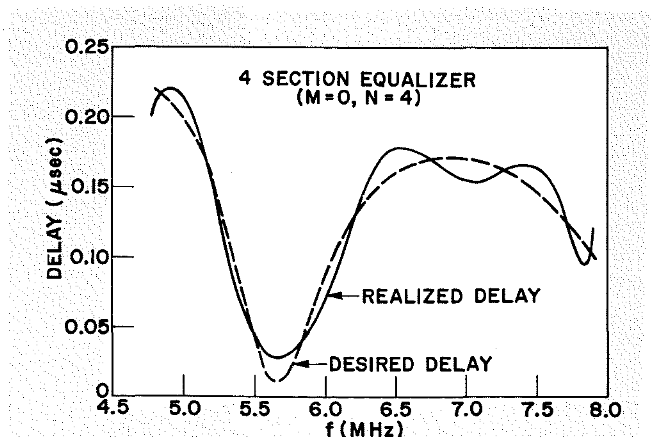


Fig. 8—Results of band-pass delay equalizer design.



DATA COMMUNICATION IN THE WORLD-WIDE NETWORK

This paper reviews some of the characteristics of HF radio and cable transmission of data in the world network. Briefly discussed is the impact that satellite communications systems may have on data communications channels. The role of the international working parties of the CCITT* in assuring operating standards for unhindered data flow between countries is explained. RCA Communications, Inc., is vitally concerned with and active in all these efforts.

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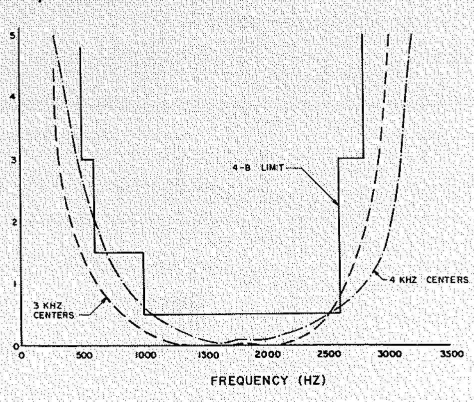
Electrical communication between the continents had its beginning in the work of Cyrus Field. One hundred years ago he laid the first submarine cable between the United Kingdom and Newfoundland. The network which grew from this beginning carried much of the world's commerce until World War I when wireless made its debut. Concurrently, the telegraph line of Samuel Morse and the telephone line of Alexander Bell developed into national networks embodying wire lines, coaxial cables, and microwave radio links. In this decade, a new tool became available, the satellite repeater. Field's first cable could handle only tens of words per minute. Today's cables handle many thousand words per minute, and tomorrow's satellites will handle millions.

Final manuscript received October 7, 1966.

* In this paper CCITT denotes International Consultative Committee for Telephone and Telegraph—and CCIR denotes International Consultative Committee for Radio.

Fig. 1—Group delay.



HF RADIO CHANNELS

The scope of radio, which embraces the lower end of the electromagnetic spectrum, has broadened out from the early wireless days, but for discussion we will separate high frequency and microwave and consider the latter along with cables and satellites.

High-frequency (HF) radio, that is frequencies between 3 and 30 MHz, have been used for intercontinental communication for almost four decades. These waves are propagated along a path around the earth by multiple reflection between the ground and the ionized layers which surround the earth from about 100 to 300 km above its surface. This path is subject to severe fluctuations. It is strongly influenced by the sun and follows a diurnal cycle. The ion density in the reflecting layers builds up rapidly under the ultra violet and other ionizing radiation from the sun at sunrise, staying in equilibrium during the daylight hours with a broad peak around noon. When the sun sets, the ion density drops rapidly in layers closest to the earth (*D* and *E*) and more slowly in the outer layers (*F*) due to the lower particle density and longer mean free path. The ion density determines the maximum frequency which can be propagated by this means. The amount of attenuation suffered by a signal is also a function of the ion density; the lower the frequency, the greater the attenuation for a given ion density. Thus, below the maximum usable frequency, the attenuation of a signal will change slowly with time as the ion density changes. This phenomenon is *flat fading*.

The number of paths through the ionosphere between two points on the earth's surface is potentially large. The ionized regions appear more as turbulent clouds than firmly defined shells, and while the reflection is not specular, it follows Snell's law to a first approximation. The difference in propagation time between signals following different paths varies up to several milliseconds. When the difference between the principal rays is equivalent to an odd number of half cycles of the carrier frequency, sharp fading occurs due to phase cancellation. In addition to this fading, there is an uncertainty in the timing of the mark-space transitions of the signal caused by the presence in the detector of an overlap of mark and space signals from different paths. The time spread between the first arriving signal and the last can be many milliseconds; however, we seldom find signals more than 3 ms apart whose amplitudes are so nearly equal that they seriously affect the mark-space transitions.

With a possibility of a 3-ms uncertainty in the timing of signal transitions, short signaling elements are impractical. Element lengths equal to at least twice the expected uncertainty are essential, and for a reasonably good operating margin, elements at least 10 ms long are required.

When data signaling rates much in excess of 100 baud are encountered, the bit stream must be commutated into a number of channels not in excess of 100 baud and reassembled at the receiver. Parallel channels capable of 100-baud signaling can be derived from HF single-sideband (SSB) transmissions by using audio tones appropriately modulated and separated by bandpass filters. The CCIR* has recommended standards for such transmissions which are in universal use on the worldwide network. The tones, 170 Hz apart, are frequency modulated with a modulation index of 0.8.

	Schedule 4 data	Schedule 4A data	Schedule 4B data
I Attenuation Characteristics			
Meas. between 600-ohm impedances at line-up	8 dB ± 1 @ 1000 Hz	8 dB ± 1 @ 1000 Hz	8 dB ± 1dB @ 1000 Hz
Expected max. var. of (L)[a]	Short term ± 3 dB Long term ± 4 dB	Short term ± 3 dB Long term ± dB	Short term ± 3 dB Long term ± 4 dB
Frequency response	Freq. range Var.-dB 350-2000 Hz —2 to +6 2000-2500 Hz —3 to +12	Freq. range Var.-dB 300-1000 Hz —2 to +6 1000-2400 Hz —1 to +3 2400-2700 Hz —2 to +6	Freq. range Var.-dB 300-500 Hz —2 to +6 500-2800 Hz —1 to +3 2800-3000 Hz —2 to +6
II Delay Characteristics			
Circuit delay	Not specified	Not specified	Not specified
Envelope delay	Less than 1000/ms over band from 1000 to 2400 Hz	Less than 1000/ms over band from 1000 to 2400 Hz	Less than 500/ms @ 1000-2600 Hz; less than 1500 ms @ 600-2600 Hz; less than 3000/ms @ 500-2800 Hz
III Noise Characteristics			
Circuit noise (S/N ratio)	48 dBa FIA (26 dB)	48 dBa FIA (26 dB)	48 dBa FIA (26 dB)
Impulse noise	Not specified	Not more than 70 noise peaks/hr will exceed —18 dBm (long term average)	Not more than 70 noise peaks/hr will exceed —18 dBm (long term average)

Fig. 2—Amplitude and delay limits established for data transmission by CCITT.

At the receiver the tones are separated with bandpass filters and demodulated. Dual-space diversity is used in the HF receivers to control fast fading. The signals are finally retimed, serialized, and delivered to the user by land line.

On the *transmit* side, the serial bit stream can be commutated into a number of parallel channels with little more than a clock and a shift register. At the *receive* end of a noisy, perturbed channel, the story is different: The timing information must be recovered from the signal transitions, and the bits from the parallel channels must be lined up and serialized. The bandpass filters in the parallel channels have different delays which must be equalized.

Telegraph distortion, a measure of circuit and equipment performance, is the amount by which the interval between the bit transition and a sampling pulse which marks the center of the bit interval is greater or less than one half bit. It is expressed as a percentage of the bit, thus 50% is the maximum distortion possible.

It is the nature of the bit edges or signal transitions to jitter giving rise to distortion which can be measured in two ways. Average distortion is measured by producing pulses whose width is equal to the amount of distortion and applying them to a d'Arsonval meter. The time constant of the meter circuit provides a running average. Peak distortion is measured by a latching type meter circuit which displays the greatest deviation during the observation interval.

Undistorted signals are obtained by sampling the detector output near the center of each bit and releasing a perfectly timed bit of the proper sense. The amount of distortion that can be corrected by this sampling technique is theoretically half the bit length less half the sampling pulse length. In the HF medium, the peak distortion is typically 30% to 40% with occasional peaks over 50% in the 100-baud parallel channels. With the sampling technique it is possible to correct a single channel with up to about 48% distortion. The propagation delay in the 3-kHz audio path between the channel modulators feeding the radio transmitter and the demodulators following the radio receivers is not uniform with frequency, and in addition, jumps with diversity switching. If the sampling pulse timing is derived from a single channel and applied to all channels, the correction capability is reduced to about 40%. Furthermore, it is virtually impossible to adjust the channel delay equalizers on signals which are bouncing around.

To overcome this effect, each 100-baud channel must recover its own timing and deliver a regenerated signal to the parallel to serial converter. This multiple regeneration is accomplished by driving a bank of regenerators from a stable crystal oscillator whose frequency is 127 times the baud rate. In each channel regenerator a 2^7 countdown chain which has one pulse per cycle normally inhibited provides the sampling pulses at the center of each bit. A detection circuit compares each incoming transition with the corrected clock and either inhibits a second pulse in the countdown or cancels the inhibit on the first one to slow down or speed up the sampling pulses. In this manner a correction rate of slightly less than 1% per bit is obtained.

The outputs of the channel regenerators are fed to the parallel-serial register through delay equalizers which now can easily be adjusted to bring the transitions from the various channels into alignment within about 1%. The high speed serial clock in the parallel-serial register derives its timing from one or more of the parallel channels in a manner similar to the channel clocks.

Bit error rates between 10^{-4} and 10^{-5} can be obtained on all but the most difficult radio circuits. This is reasonable performance in itself, and where higher orders of accuracy are required, the circuit can be treated with standard error control schemes. Interference, particularly on night frequencies, when propagation conditions are poor, is a major problem. Born of congestion, it is not likely to diminish in the future.

CABLE CHANNELS

Channels carried strictly on metallic links are rather few in the worldwide network, however, the term *cable channel* has become the designation for those channels which are carried between continents and islands by an underwater coaxial cable, and the terrestrial tail circuits associated with them. These tails may, themselves, be coaxial cable, wire line, line-of-sight microwave, or in a few extremes tropospheric scatter links. These latter links introduce problems similar to HF links, but not as severe.

The attenuation in a cable channel is fairly stable over long periods of time. The submarine environment has a fairly constant temperature and the voltages applied to the cable are well regulated. The terrestrial ends show small slow changes due to environment but these can be controlled by a pilot-operated AGC system. Transient effects such as short dropouts, noise bursts, and phase

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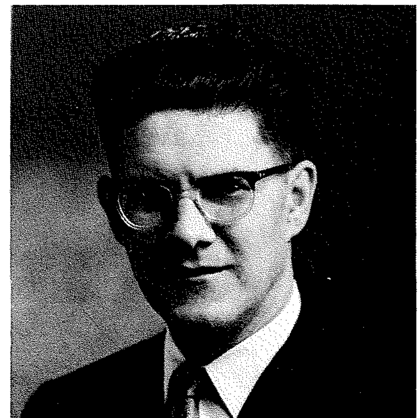

excursions do occur and constitute the main problem.

In modern communications systems, many channels must be multiplexed on a coaxial cable to make its operation economically feasible. The individual channels are multiplexed in a frequency division system in groups of 12 with appropriate bandpass filters and control pilots. The groups, in turn, are combined five at a time into super groups in a similar manner. The total transmission package then consists of similarly combined super groups. Within this kind of a system, most of the problems arise from the channel derivation equipment and noise.

The most common unit of bandwidth in the worldwide network is the voice channel. It is derived from wider bandwidths by frequency division multiplex. In the national systems the channel centers are 4 kHz apart yielding a channel approximately 3.3 kHz wide (100 to 3,400 Hz). To gain maximum voice capacity in the submarine cables for economic reasons, the channels are spaced 3.0 kHz yielding approximately 2.7 kHz (300 Hz to 3,000 Hz). The actual usable bandwidth for data transmission is determined by the differential delay as well as by the amplitude and is less than the figures given above. Typical delay curves are shown in Fig. 1 for the 3 kHz and 4 kHz cases.

Standard amplitude and delay limits have been established for data transmission by the CCITT. These range from voice grade (raw channel) to fully conditioned channels. The more important grades are shown in Fig. 2. These criteria are not difficult to meet on single links, but pose a problem when links are connected in tandem.

Three main systems of modulation are in current use on the commercial data network. Frequency modulation is the most common at speeds up to 1,200 baud. Phase modulation is used at higher speeds as well as lower, and vestigial sideband amplitude modulation works well at the higher speeds.

Frequency shift keying (FSK) is a carryover from the earlier HF systems. It is widely used at speeds up to 100 baud for telegraphy in narrow channels (120 Hz) and the CCITT has standardized an FSK modem *(Rec. V23)* for data transmission at speeds up to 1,200 baud. The modulation index at 1,200 baud is 0.66, which is a little less than optimum, but still provides good spectral distribution around the 1,700-Hz carrier frequency. Similarly, *V23* specifies a carrier frequency of 1,500 Hz and a modulation index of 0.66 at 600 baud. In the U.S., the Bell System Dataphone service utilizes the Western Electric 202 modem with FSK at a carrier frequency of 1,700 Hz and a modulation index of 0.83 at 1,200 baud. The Dataphone uses the two wire switched telephone network. Western Union Broadband exchange service (BEX) is a four-wire switched network providing service similar to the Dataphone. The Western Union modem utilizes FSK with a carrier frequency of 1,800 Hz and a modulation index of 0.66 at 1,200 baud.

Vestigial sideband amplitude modulation has been in use on international circuits for a number of years. Since it transmits only the lower sideband and a small remnant of the upper sideband, it enjoys nearly a two-to-one bandwidth advantage over FSK and PSK. With the carrier positioned at 2,800 Hz, a speed of 4,800 baud is possible on a well-conditioned voice channel.

Phase shift keying (PSK) is well suited to channels which have a low noise level permitting multilevel modulation. Demodulation is accomplished by comparing the phase of the incoming signal with a local oscillator or with the previous bit which has been stored. In the former, the absolute phase difference determines the element polarity; in the latter, the change in phase determines the polarity. When a local reference oscillator is used, it must be phase locked to the source by a pilot frequency or by timing recovery from the signal transitions themselves. The simplest PSK system uses phase criteria 180 degrees apart to denote mark or space. Two binary bit streams which are synchronous can be combined into a 2^2 (or 4-level) system which can be assigned phase criteria 90° apart. Similarly, three binary bit streams yield 2^3 (or 8-level) modulation with criteria 45° apart. The 8-phase system is marginal on many parts of the world network at the present state of the industry and represents an upper limit.

The instantaneous signal voltage at the detector regardless of the modulation system is the vector sum of the spectral components distributed throughout the band. Thus, any perturbation, static or time-varying, which alters the relative phase or amplitude of the components will produce distortion.

The channel derivation multiplex has filters which produce delay characteristics similar to that shown in Fig. 3. By adding delay to the center portion of the channel, the usable bandwidth can be increased. Commercial equalizers are available which are adjustable in 200 Hz steps from 750 to 2,950 Hz to a shape which approximates the inverse of the channel delay.

Each section of the equalizer is an all-pass network containing a resonant circuit which introduces up to about 2 ms delay at its resonant frequency. The relationship between delay and frequency for a typical section is shown in Fig. 4 for high and low settings of the adjustment. At a nominal 200 Hz between sections, a fairly smooth resultant curve can be obtained for moderate amounts of delay, but when the sections are adjusted near maximum, the overlaps are not as great and a ripple appears in the delay curve. For nominal bandwidths (600 Hz to 2,600 Hz) this problem is not too serious, but for the higher speed transmission where the delay must be compensated down to 400 Hz, the high rate of change of delay with frequency in the raw channel is a difficult problem. Equalizer sections which have a high rate of change of delay with frequency have a narrow bandwidth, requiring more closely spaced sections. Furthermore, the all-pass delay networks are similar in nature to the transmission channel networks making it rather difficult to produce conjugate characteristics. The sections nearest the center of the channel require the greatest delay, hence, have the steepest delay curve whereas the sections at the channel edges have the least delay and the flattest delay curves. This is opposite to the slope of the channel curve we are trying to correct.

Noise appears at the output of a cable channel from several sources. Underwater sections are relatively quiet but still generate about one picowatt of white noise per kilometer. Channels in terrestrial microwave and cable tend to be noisier because the thermal environment is less stable and the cables traverse telephone plant where they can pick up noise from other circuits. Impulse noise from telephone switch gear is also a serious offender. Permissible noise objectives are contained in *CCITT Recommendation G.222.*

The effects of the noise on a signal depends somewhat on the modulation method. A noise burst in the channel puts a noise spectrum on top of the signal spectrum which distorts the envelope. In the case of FM it may or may not result in a wrong decision in the detector. In an AM system, the space condition is always wiped out by a noise burst. The PSK detectors compare the signal with a reference oscillator or the previous bit which has been stored. This is a more powerful method of detection, because it searches out coherance, but it suffers if there are phase perturbations in the channel translation oscillators. The translation oscillators in a 12-channel group are controlled by a group pilot. When noise, such as cross products from other channels, lands in the pilot channel it may cause the AFC system to jitter and produce a sharp change in phase in the channel derivation oscillators. If this shift in phase occurs within the interval of one bit, and has a magnitude approximating the phase shift of the modulation, an error will occur. In practice, 2-phase and 4-phase modulation work fairly well, but the error rate on 8-phase modulation systems is marginal.

In recent tests between RCA Communications, Inc., San Francisco, and Tokyo using Rectiplex equipment furnished by their correspondent, the Kokusai Denshin Denwa Company error rates of 0.57×10^{-5} were obtained with a 4-phase system, and 1.18×10^{-5} with an 8-phase system. These rates were obtained on the submarine cable; however, the phase perturbations on the transcontinental network were too great to permit extension of the tests to New York.

SATELLITE CHANNELS

In April 1965, a new and important member joined the family of communications facilities. The EARLY BIRD satellite was launched from Cape Kennedy and took its place over the equator near mid Atlantic. After preliminary tests and alignment, channels were made available by Comsat to the International Record carriers to observe the performance of these channels in the regular data and telegraph services. RCA Communications, Inc. participated in these tests on June 9 and 10, 1965 and January 19 and 20, 1966. Some of the results, and a comparison with similar services using submarine cable channels, are quite interesting.

The channels used were 4B-equalized and were carried over AT & T facilities to the Andover, Maine, earth station. The June 1965 tests began using the European earth station at Raisting, Germany, but due to technical difficulties on June 10, 1965, the control was shifted to the Goonhilly Downs earth station in England. The January 1966 tests were carried out with Rome via the Fucino earth station.

Satellite channels are derived by frequency division multiplex similar to that used on terrestrial microwave links and submarine cables. The delay distortion due to the channel filters is therefore approximately the same. The channel centers are spaced 4 kHz, and when combined with equalized tails 4B end-to-end characteristics were easily obtained. Since the observing period was short most of the effort was concentrated on noise and stability observations. End-to-end noise levels were measured at −40 dBO or lower on the Frankfurt, Germany and Paris, France channels. The noise level on the Rome channel was −30 dBO.

Loop propagation time, which is an important parameter in error correcting data systems, was measured as follows:

New York-Paris-New York—530 ms
New York-Frankfurt-New York—540 ms

Transient phenomena are the most troublesome in data transmission. Impulse noise accounts for a large share of the errors received and occurs in a random fashion. The main source of impulse noise is not in the space segment, but is in the terrestrial links which share a common environment with automatic telephone and telegraph switching centers. Phase perturbations on the transmitted signals were expected and observed. The translation error in the multiplex was very low, 1 part in 200,000. Superimposed on this slow change in phase (1.8°/second at 1000 Hz) were short phase perturbations of about 5° to 10°. The measurements were limited by the time constant of the recorder so that the rate of change of phase could not be accurately determined; however, a phase shift of 10° in less than the duration of a bit would make 8-phase PSK with a 45° total deviation marginal.

The differential delay due to the voice multiplex channel filters is about the same as a terrestrial microwave system. However, superimposed on the delay curve is a variation of about 400 microseconds, peak-to-peak. Although the synchronous satellites remain substantially over the same point on the earth's surface, small movements along all three axes take place due to inaccuracies of the orbit and noninformity of the earth. This gives rise to small changes in path length and accompanying changes in propagation time.

CONCLUSION

We have discussed some of the more interesting characteristics of HF and cable transmission in the World network as it is today. We have had a preview into the satellite channel picture through the test programs and we are on the springboard ready to dive into data via satellite when the channels become abundantly available. The satellite story is short, but even as this paper goes to press, it will be growing with new "birds" going aloft.

The international working parties of the CCITT meet regularly around the world to set down standards of operation to assure that data can flow from country to country without hindrance. The worldwide network is indeed a crucial link in the information explosion of the sixties, and RCA Communications forms an important part of that network.

ACKNOWLEDGMENT

This author wishes to express his appreciation to all his colleagues without whose help this paper would not have been possible.

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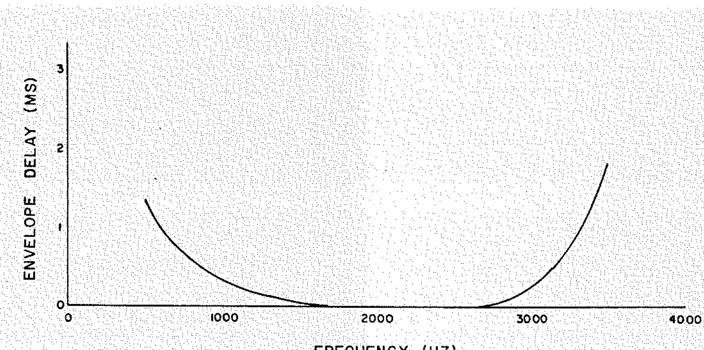

Fig. 3—Delay vs. frequency, single-carrier channel.

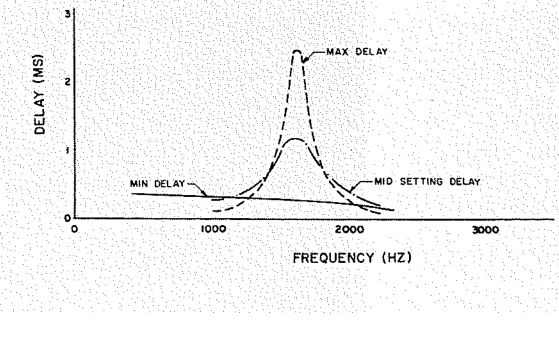

Fig. 4—Typical equalizer section; center frequency—1600Hz.

SATELLITE COMMUNICATIONS—WHAT NEXT?

The demand, applications, and technology for future satellite communications is discussed by the type of service. Transoceanic and military point-to-point telecommunication is described as the natural extension of the world-wide trunking networks. Intranational service is portrayed as a high capacity trunk or a TV distribution network. Direct-to-home TV and radio are examined under the broadcasting service application. The broadgathering service, exemplified by the world-wide weather network, emphasizes the needed growth in satellite receiver and antenna technology. The mobile service application indicates some of the unique possibilities offered by satellite communication. Key technological growth areas are identified in each service.

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The future growth of satellite communications depends on the same few factors which are historically common to the continued growth of any form of communication—an advancing technology, an economic satisfaction of demand, and unique service qualities or capabilities. Transoceanic telecommunications exemplifies these characteristics. In 1956, a submarine cable[1] (TAT-1), with 36 circuits, began the replacement of radio by cable as the primary means of meeting the growth demands of transoceanic service. In 1965, the EARLY BIRD satellite,[2] with a 240-circuit capacity, had 64 operational transatlantic circuits. This demonstrated that satellites can complement the submarine cable by implementing the capacity needed to meet future demands. Satellite communications will impact many existing communication services, and its growth is assured by exploding technology trends and unique service capabilities.

BACKGROUND

The growing population of communication satellites, shown in Fig. 1, proceeded from experimental objectives to first starts of commercial (EARLY BIRD) and military, the Initial Defense Communications Satellite Program—IDCSP[3] systems in just a few years. A steady increase in booster capability now provides thrust for heavier or multiple satellite launches. This increase in payload capability, shown in Fig. 2, permits increased prime and radiated powers at higher orbit altitudes. A satellite in an orbit altitude of 18,200 nmi provides a synchronism with the 24-hour earth rotation. With the satellite in the equatorial plane, it always remains over the same point of the earth's surface. At this altitude the satellite radiates a line-of-sight coverage of 32.8% with a ground antenna elevation angle of 5° of the earth surface (Fig. 3). For other applications, a narrow beam remaining over a fixed small area on the earth is within the state-of-the-art. Synchronous satellites correct initial launch injection errors and small perturbations due to solar and lunar influences and the earth's oblateness with on-board stationkeeping. Today's TITAN III-C booster capability easily launches six to eight satellites into synchronous orbit.

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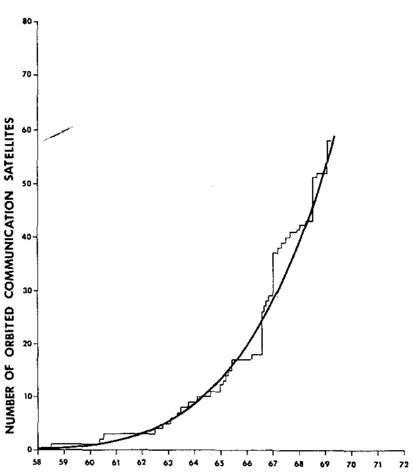
HIGH CAPACITY TRUNK ROUTES

The point-to-point transoceanic telecommunication service has been addressed by the TELSTAR, RELAY, SYNCOM, EARLY BIRD, and INTELSTAT II sequence of satellites. This commercial service includes telephone, television, telegraph, telex, and facsimile. Today, 195 million telephones ring the world. Estimates for 1980 indicate that there will be 500 million phones in use, an increase of 2½ times over the present number. Even more critical to transoceanic telephony is the prediction that 70 million overseas calls will be placed in 1980, nine times the 1965 amount. *How is this demand to be met?*

Fig. 1—Communication satellite launches.



In 1915, the radio telephone was demonstrated between Paris and Arlington, Va. Twelve years later, in 1927, commercial service established between New York and London, logged 2,300 calls. by 1955, users could reach 108 countries via radio and placed 1,700,000 calls. In September 1956, the first trans-Atlantic submarine cable (TAT-1) was placed in service linking the United States and Great Britain. A growth rate of 20% per year occurred and in 1965 over 8 million calls were placed. The cable network provided more circuits and faster service. Today the cable network, shown in Fig. 4, provides world-wide connectivity.

Cable technology has increased the circuits per cable from 36 initially in TAT-1 to 138 in TAT 3 & 4. Newest plans call for a cable between Florida and St. Thomas, with 720 circuits. The predicted demand for transoceanic circuits is shown in Fig. 5. Cables now supply about 600 North Atlantic circuits. These predictions[4,5] show the demand for the North Atlantic, the total United States terminations, and the total world. The figures do not include forecasts for television or the anticipated surge in computer-to-computer requirements. It is obvious from these estimates and the successful demonstrations of EARLY BIRD, that communication satellites can have a significant role in commercial transoceanic telecommunications.

Typical of the earth terminals is Canada's experimental earth station[6] at Mill Village, Nova Scotia (Fig. 6). This was completed in February 1966. Significant advances in this station were the

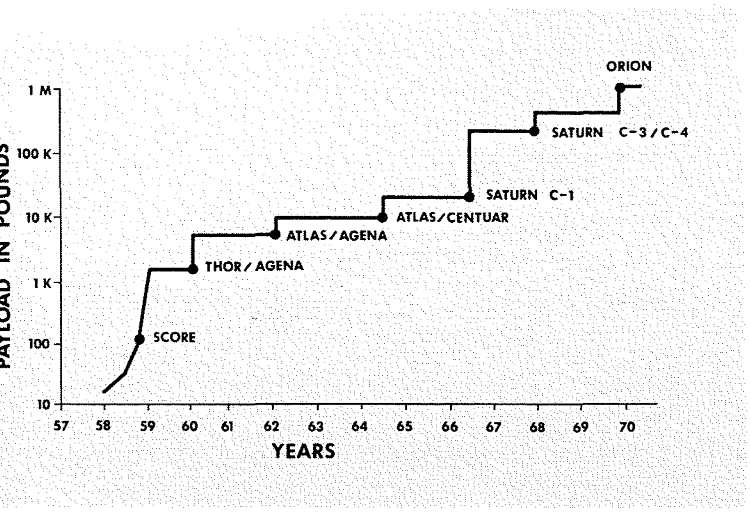


Fig. 2—Booster payload capability.

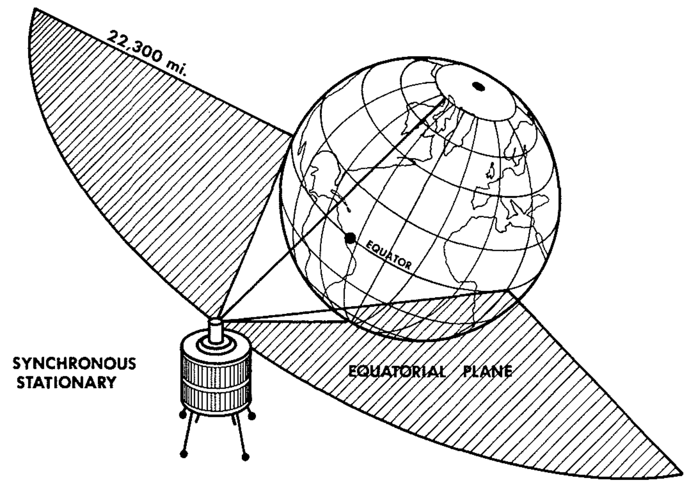


Fig. 3—Coverage zone-equatorial synchronous.

high efficiency cassegrainian feed system[7] and the wide operating frequency range (3.7 to 4.2 GHz with 200-MHz instantaneous bandwidth). An 85-foot-diameter parabolic antenna mounts on an azimuth-elevation type of pedestal. A 120-foot-diameter inflated radome encloses the antenna system. The antenna gain is 58.9 and 61.0 dB at 4.1 and 6.2 GHz, respectively. Total system noise temperature is 60° K at a 7.5° antenna elevation. A 27-dB gain parametric amplifier, cooled to 4.2° K by liquid helium is followed by a 20-dB, 4.5-dB noise-figure tunnel-diode amplifier. Transmit power is 10-kW peak for SSB and 8-kW carrier for FM. The SSB capability allows for tests with the NASA Application Technology Satellites. NASA will test with SSB up-link transmission with the satellite—converting the modulation to PM for down-link transmission.

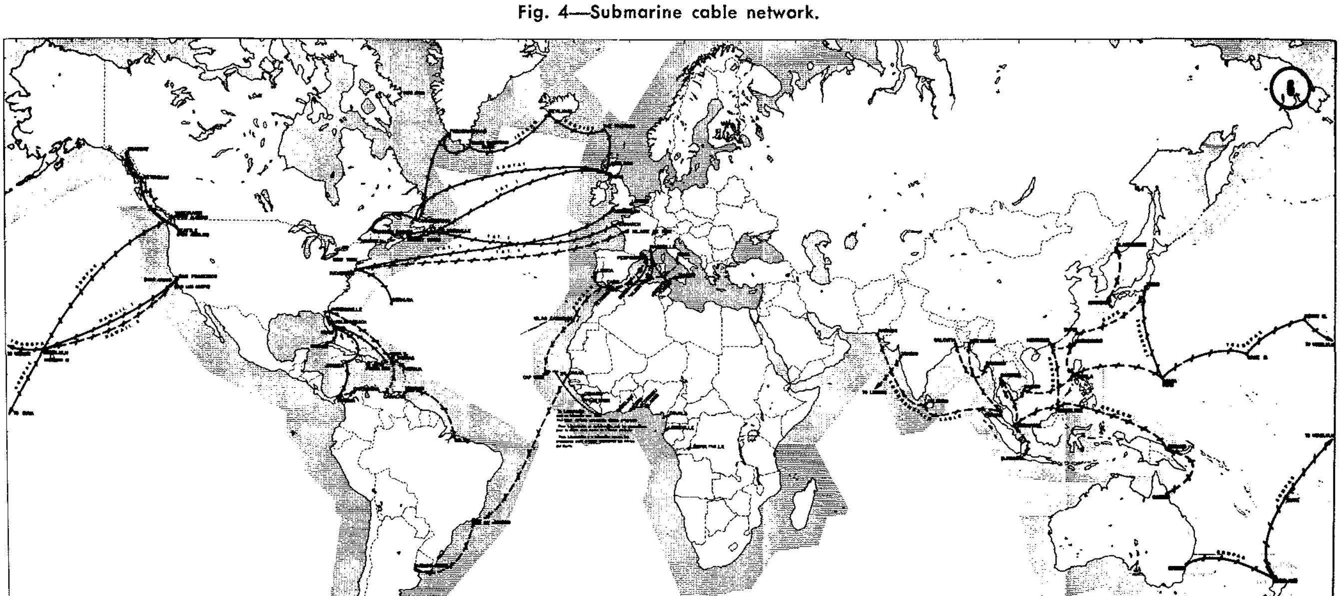
Presently, FM up, FM down is in use commercially. RCA Victor, Montreal, designed and developed this terminal for the Canadian Department of Transport. As the global commercial system develops, approximately 50 to 75 terminal sites, with 100 antennas, could be placed into service in the 1970's. Ground stations now in operation or under construction at sites are shown in Fig. 7.

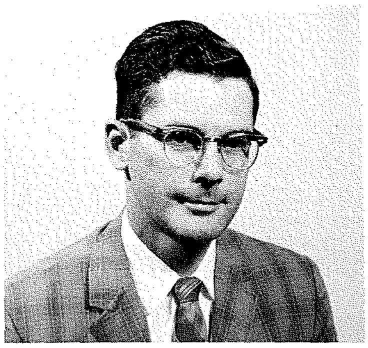
THE POINT-TO-POINT MILITARY REQUIREMENT

In contrast to the commercial point-to-point service, the military system adds requirements—classified traffic, high survivability, increased reliability. These characteristics have evolved into the Department of Defense developed Initial Defense Communication Satellite Project (IDCSP) under the Defense Communication Agency (DCA). Eight satellites, deployed in a circular orbit of 18,200 nmi on June 16, 1966, represent the start of operational military tests.

Transportable ground terminals developed for this service are the AN/MSC-46 (Mark 1-B) and AN/TSC-54 (Mark V). The Mark 1-B has a 40-foot-diameter parabolic antenna and can be transported by air. It has a 27-man operating crew. The Mark-V uses an 18-foot clover-leaf antenna structure with four separately feed dishes. A single C-130 aircraft transports the terminal, six operating personnel, and the support equipment.

Fig. 4—Submarine cable network.





THOMAS R. SHERIDAN received the BSEE from the University of Notre Dame in February 1950, and an MSEE from the Polytechnic Institute of Brooklyn in June 1957. He joined RCA Laboratories in May 1952 at Broad St., New York. He was a contributor to the development of RCA/MUX/ARQ Error Correcting Equipment for world-wide teletype channels. He obtained five patents in this area and received the RCA Achievement Award for this effort in 1955. At the Rocky Point, Long Island branch of the Laboratories, he engaged in digital communications research and headed up the System Simulation and Evaluation Center (SYSEC). As a sub-task leader on the Navy's Project Pangloss, he directed a research team evaluating low redundancy coding techniques. In 1962 Mr. Sheridan directed early laboratory support to the DEP Minuteman Program. After transfer to the New York Systems Laboratory of the Communications Systems Division, he had prime responsibilities for the performance evaluation and test of the Minuteman Communications Systems as Associate Director-Systems Engineering. He became Technical Director on the Air Force Dormant Missile Study, covering advanced command/control communications concepts. He is presently with SEER, as Staff Systems Manager, engaged in satellite communications and command/control communications efforts for DEP. Mr. Sheridan is a senior member of the IEEE and AIAA, and Vice-Chairman of the IEEE Translation Journals Committee.

This system can rapidly extend the basic DCA beltline communications network to any crisis area or reestablish emergency capability in the event of natural (earthquake) or man-made disruption. IDCSP provides communications to major headquarters, field commanders, and selected ship and airborne terminals.

Primary stations operate at Fort Dix, N.J., and Camp Roberts, Calif. These stations have 60-foot antennas and were originally built for the U.S. Army ADVENT program. Transportable terminals testing IDCSP are located in Hawaii, the Philippines, West Germany, and Asmara, Ethiopia.

The prior effort on moon bounce (DIANA), SCORE, COURIER, ADVENT, and the NASA SYNCOM programs contributed to the success of the IDCSP program. The ADCSP (Advanced Defense Communications Satellite Program) aims at providing a future system of increased capabilities for transmission of voice, telegraph, data, and graphic communications to an increased population of military users.

TRUNK ROUTE IMPLEMENTATION

Present satellite implementation of point-to-point trunk routes is not very efficient. The synchronous satellite antennas provide the 17.5° earth coverage (Fig. 8) plus an allowance for stabilization inaccuracy. The bulk of the transmitted energy dissipates everywhere but at the receiving antennas. High-gain directional antennas (with multiple beams), significantly increasing the energy received at gateway terminals, need development. The narrow beams to Andover and Goonhilly, in Fig. 8, show the improvement possibilities. Elimination of the stabilization error becomes more critical as the antenna angle is reduced. Beam pointing of the satellite antenna system provides additional and significant improvements in efficiency. While 1,200 MHz were allocated to satellite communication, only 50 MHz was allocated on an exclusive basis, therefore spectrum crowding will rapidly be a problem for satellites as it has been for the Earth radio spectrum.

INTRANATIONAL ROUTING

Intranational routing is the application of satellite communications for national telecommunication service, and contrasts with transoceanic service between two or more countries. Communication satellites would, in general, be competing here with the services provided by cable, microwave, and troposcatter. The USSR has begun operational experiments with the MOLNYA satellite. There are discussions in the United States to fly a domestic TV distribution satellite. This satellite would relay, from a few origination points, program material to the approximately 500 TV local transmitting stations in the United States. Such a system has been advanced by many organizations and its implementation is within the state-of-the-art. A satellite weight of about 1,500 lbs can provide about 10 simultaneous TV links. The power required would be about 300 watts.

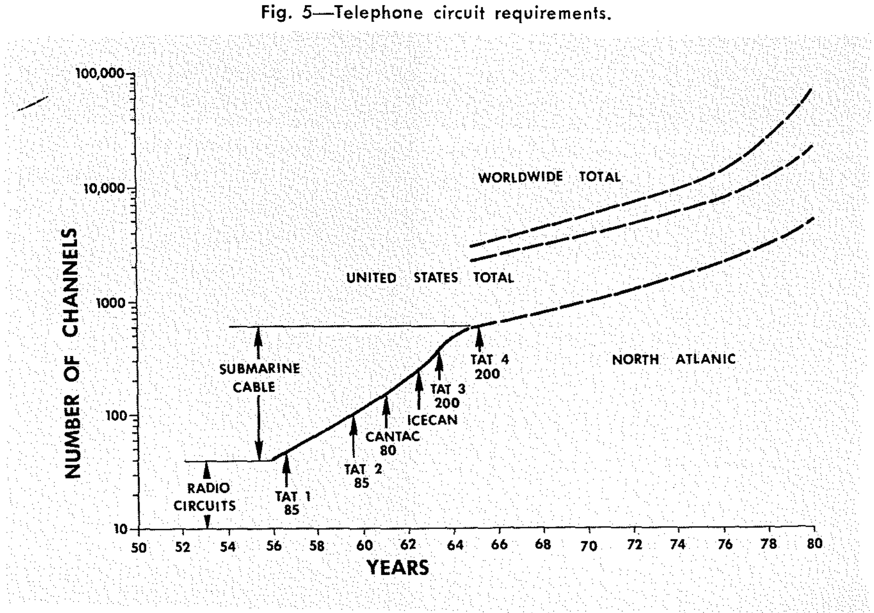
Conceivably, a TV distribution network could also be used for high-speed data between computer centers when such a service is required. Special encoding techniques will have to be employed to maintain error control over the time delays of satellite transmission. Communication satellites will be highly competitive to existing services in countries where a large capital investment in facilities has not been made. The underdeveloped countries, in particular, will have an advantage in selecting and comparing satellites for rapid expansion of their telecommunication plant. In congested terminal points, satellites alleviate a spectrum crowding which is nearing saturation in many cities.

BROADCASTING

High-powered, physically-large spacecraft provide the capability to transmit TV[8] and radio directly to home receivers. Achievement of this service will have tremendous impact in advancing the mutual understandings among nations. In the United States and Europe, TV for entertainment, education, news, and weather service will start a growth curve in the 1970's. In underdeveloped nations, where the cost of transmitters per person per square mile ratio is high, satellites offer an expansion of broadcasting services to significant percentages of the population. Use of this capability for education will have significant social impacts.

A TV broadcasting satellite requires a large antenna aperture and kilowatts of

Fig. 5—Telephone circuit requirements.



prime power. Currently, an experimental satellite to test the operational environment of this service has been studied. The VISTA satellite[9] has typical system parameters as shown in Fig. 9. Developments are already in progress in all technologies needed to implement this service.

BROADGATHERING

The inverse of the broadcasting mode finds many satellite applications. Broadgathering occurs when many small (relatively cheap) sensors or local report transmitters have wide geographic distribution and the satellite acts as a data gatherer. The satellite dumps its collection of large quanities of data to a few central processing stations for edit and information extraction. The world-wide weather system is an example of this service.

A satellite for the gathering service, by its functional requirements, will introduce complexity and high performance in the receiving portion, keeping the cost of the large number of ground terminals to a minimum. At synchronous altitude a radar antenna's pointing capability is needed on the satellite. If in lower orbit, then the satellite can be multi-purpose carrying sensors of meteorological interest. The RCA-developed TIROS and ESSA[10] satellites are examples of this community of interest.

MOBILE SYSTEMS

The military application of IDCSP and ADCSP is primarily of a strategic orientation and, although these systems work in tactical areas, neither is really "tactical" in the military meaning of the word. The size and complexity of the terminals limit their battlefield use. A truly tactical concept would place complexity in the satellite and extreme simplicity on the ground. Many mobile systems consist of a large number of low traffic users, to whom it is important to have instant service when needed.

Another example of introducing complexity in the bird to benefit the small user is the commercial satellite studies for ship-to-shore (156-MHz to 162-MHz) and air-to-ground (118-MHz to 136-MHz) service. In some concepts this service has been combined with high-density trunking (4 to 6 GHz) in the spacecraft. Provisions for possible crossover between two frequency bands are being considered. Switching selection between any of the microwave down-link transponders is made by up-link microwave frequency selection. This limited capability will be improved with the start of other types of switching, such as space division in the 1970's.

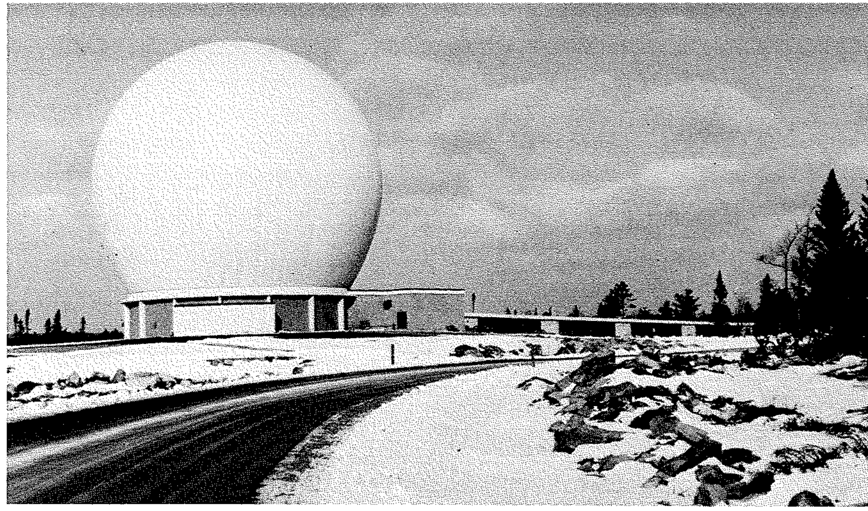


Fig. 6—Canada's Mill Village earth station.

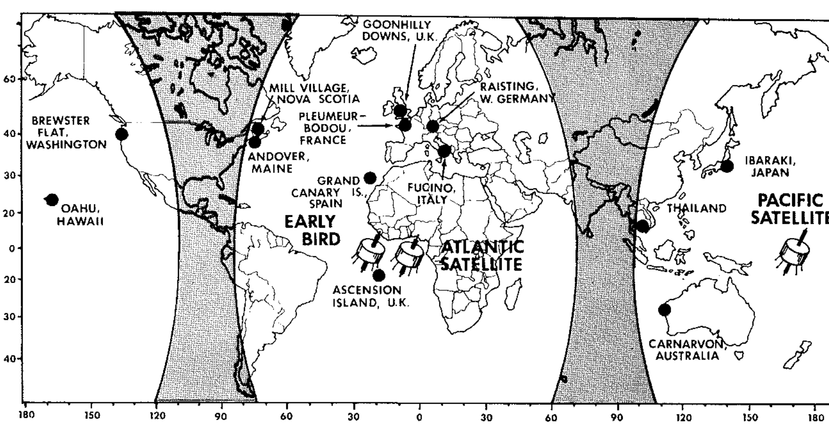


Fig. 7—Present commercial ground stations.

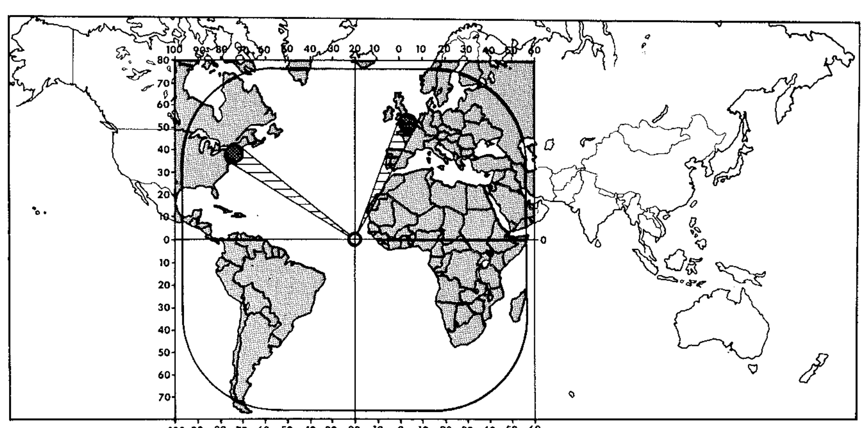


Fig. 8—Narrow beam trunking.

Impetus for this application comes from the air traffic control problem over the North Atlantic. The HF communications will be replaced by more reliable satellite communications, gaining needed safety for higher performance aircraft and increased flight densities. Satellite navigation concepts to augment the communication requirements are under study by a number of agencies. In mobile systems, the signal processing and multiple access problem are highlighted for effective utilization of satellite bandwidth and power.

TECHNOLOGY

One of the most important aspects of communication satellites is the efficient generation of power.[11,12] As shown in Fig. 10, the solid-state approach is favored at lower frequencies and power levels. At higher frequencies and powers the traveling-wave tube (TWT) is preferred. While improvements in solid-

FREE SPACE LOSS	182 DB
OTHER LOSSES	8 DB
TOTAL	190 DB
NOISE TEMP GROUND (560°K)	27.5 DB
GROUND ANTENNA GAIN	17.5 DB
SATELLITE ANTENNA GAIN	39.2 DB
(50° Lat.; 1.75° BEAMWIDTH)	
DETECTION LOSSES	6.7 DB
BASEBAND WIDTH (MHz)	4.25
REC SYST NOISE POWER (DOWN LINK)	134.9 DBW
RADIATED POWER FOR	
30 DB S/R AT SYNC TIP	
BLACK LEVEL	2.9 KW
SOUND	.5
TOTAL	3.4 KW

Fig. 9—Typical system parameters.

state devices will permit its selection to the right of the curve, development effort in TWT's promises higher efficiencies and broader bandwidth. The selection will be made on the basis of the specific application when near the locus curve. In the early 1970's, paralleling and phasing schemes will permit increasing the power (with reduced efficiency) to 500 watts at UHF and 100 watts at 7 to 8 GHz.

The need for directive antennas, phased arrays, and multiple beams appears to be established by the various applications already discussed. The entry of other than spin-stabilized control systems such as gravity or momentum systems provides an antenna platform which can be constantly directed to earth. Today, spinning satellites require electronic or switched phasing of a circular ring of elements, or mechanical despining the antenna to maintain the beam pointed at earth. The "stabilite" momentum control system[13] used a single reaction wheel to attain a three-axis attitude stability.

The achievement of an antenna pointing control system is increasingly important as narrower beams come into use. The NASA Advanced Technology Satellites and the Lincoln Experimental Satellites[13] are proving grounds for many of the new techniques which will have widespread applications in the 1970's.

CONCLUDING OBSERVATIONS

Other, hard-to-definitize applications are striving for concepts to use satellites. Many experts believe that data communications volume will outpace telephone requirements by 1980. The interconnection bandwidths needed between computers and for the expected wide influx of shared computer systems will place a heavy demand on data communications. The computer also brings the information retrieval applications. Document communication requires wide bandwidths. In the field of education where computers and communcations are just starting to integrate a heavy new demand for bandwidth will occur.

The observation of a critical operation with new and expert practice will be widely viewed by local medical practitioners. The music class will attend an operatic performance in Italy by educational TV. The art, archaeological, engineering—all levels of all knowledge tree branches can identify the one place or one time characteristics which currently prevents the viewing value from being widespread. Surely, communication satellites will find many opportunities in the 1970 and 1980's to satisfy needs which today are far out.

The field of satellite communications has a challenging future in its applications and technology. It has unique stature in its predicted impact on society and in the development of new classes of communications service. Satellite base systems offer solutions to the growing volume, variety, and complexity of communications. With increased booster capabilities, heavier and more sophisticated satellites will come under study for other advanced application concepts heretofore unable to be addressed economically by any existing communication media.

ACKNOWLEDGEMENTS

While the intent of this article is informative and the subject matter not novel in the field, a perspective of the future of satellite communications is not readily available. Significant orientations in my visibility of this field came from mutual discussions and efforts of a number of people within SEER and in the DEP divisions. Particular thanks is offered to Dr. R. Guenther, Chief Scientist of CSD, who aided in the inception of the paper, and to B. Glazer of CSD, to Dr. S. Spalding, and S. Gubin, of AED, and to T. M. Johnston of SEER, for their comments and helpful suggestions.

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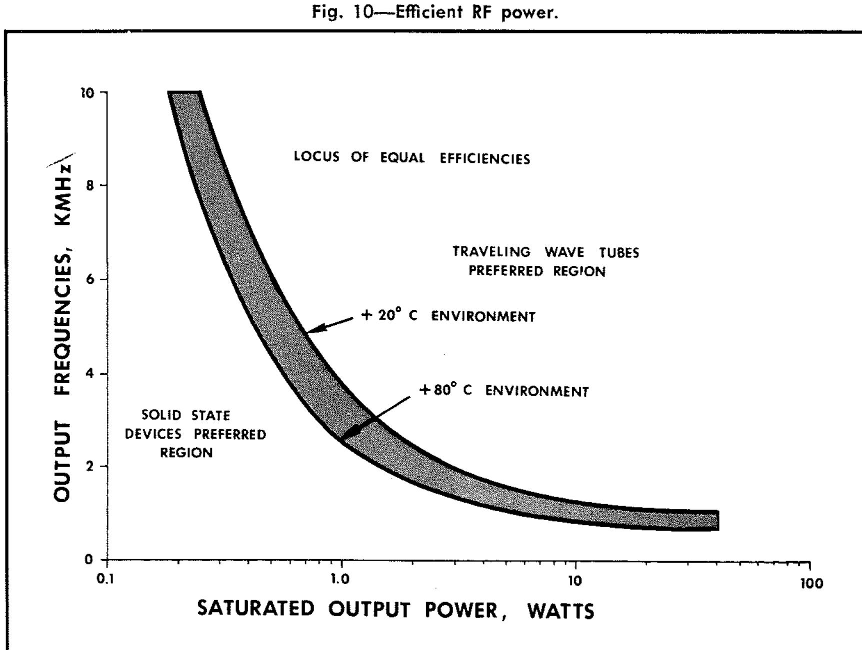
Fig. 10—Efficient RF power.



HDRSS—A HIGH DATA RATE STORAGE SYSTEM FOR NIMBUS B

The high data rate storage system (HDRSS) is being developed to provide storage and reproduction of data aboard the NIMBUS B Satellite. The system uses a 5-channel 2-speed tape recorder and a 5-channel frequency division multiplexer. The system is tested using a bench checkout unit, which simulates spacecraft signals and ground station processing functions.

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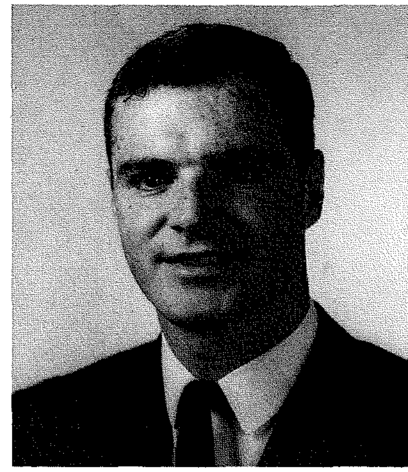


ROBERT C. CURRY received the BSEE from Newark College of Engineering in 1957 and the MSEE from the University of Rochester in 1961. Since Mr. Curry joined RCA in 1963, he has worked in two principal areas: communications for weapon systems and communications for satellite-to-ground links. In weapons communications, the emphasis was on reliable transmission of data over a survivable network while maximizing system safety. In satellite communications, he has been conducting studies to determine system configuration and modulation techniques for the transmission of video data over satellite communications links. At the Space Technology Laboratories (1961 to 1963), Mr. Curry served in the Minuteman program office as Assistant Project Engineer for the Communication Command Network. At the University of Rochester, where he was a graduate research assistant (1959-1961), and at General Dynamics/Electronics (1957-1959), he developed special purpose digital systems, and performed research on solid-state devices. Mr. Curry holds several United States patents and is a member of Sigma Xi and the IEEE.

THE high data rate storage system (HDRSS) (Fig. 1) is being developed by RCA for NASA to provide storage and replay of sensor data aboard the NIMBUS B satellite. NIMBUS B is the third satellite to date in the NIMBUS series.

The NIMBUS program involves stabilized, experimental meteorological satellites designed to circle the earth in a 600-nmi, near-polar, sun-synchronous orbit with an orbital period of approximately 107 minutes. The NIMBUS television and infrared sensors provide complete coverage of every point on the earth's surface once in daylight and once in darkness every 24 hours. During each orbital revolution, continuous infrared mappings and 31 single-frame television photographs are made.

On NIMBUS I, orbital coverage photographs were obtained using a rapid-readout (6.4 seconds) automatic vidicon camera system (AVCS) and a tape recorder with a 1:1 record-playback speed ratio, resulting in a video bandwidth of 60 kHz for an 800-line picture. This bandwidth, suitable for transmission to the two high-gain command and data acquisition (CDA) ground stations, is too large for the smaller weather stations provided around the world for use with TIROS automatic picture transmission (APT) satellites. The APT camera readout time is 208 seconds, resulting in a video bandwidth of 1.6 kHz.

To provide the capability both for real-time transmission to the local stations and complete orbital readout at the CDA station, the tape recorder speed-up concept was adopted. The HDRSS (Fig. 1) consists of redundant spacecraft tape recorders and multiplexers, ground station demultiplexers, subcarrier demodulators, digital decoders and video display equipment. Some of the ground station equipment was designed during the NIMBUS I and II programs; the spacecraft equipment was designed specifically for NIMBUS B. The spacecraft subsystem (Fig. 2) weighs 32 pounds and requires a power of 23 watts.

Final manuscript received December 1, 1966.

The heart of the HDRSS is a 5-channel tape recorder which stores and reproduces digital data and analog video signals for transmission to NIMBUS ground stations. During an orbital revolution the signals from four sensors and a clock are transmitted in real time to local stations and simultaneously stored in the tape recorder. Later, as the spacecraft passes over a ground station, the data is read from the recorder at a rate 32 times that at which it was recorded. The signals from the five tracks are frequency multiplexed and transmitted to earth via an S-band FM communications link.

FUNCTIONAL DESCRIPTION

The tape recorder (Fig. 3) operates in two modes: record and playback. The modes of operation are controlled by the record end-of-tape switch (reverts to off), the playback end-of-tape switch (reverts to record) and the *record on, playback on,* and *off* commands. Both recording systems may be operating simultaneously; alternatively, the two recorders may be operated sequentially to provide greater storage capacity.

Five signals are recorded in the HDRSS: The image dissector (ID) video signal (0 to 1,600 Hz); the high resolution infrared radiometer (HRIR) signal (0 to 360 Hz); the infrared interferometer spectrometer (IRIS) signal, a biphase 3,750-bits/s data signal; the medium resolution infrared (MRIR) signal, a biphase data signal at a rate of 1,600 bits/s; and the timing signal, a 2,500-Hz carrier, amplitude modulated by the 100-bits/s NASA Minitrack pulse width modulated (PWM) time code.

Recorder Electronics

Before the ID and HRIR signals are recorded, they are converted from video to frequency modulated subcarrier signals in the tape recorder electronics. The time-code, IRIS, and MRIR signals are amplified and are then recorded directly without further processing.

Tape Transport

The FM modulators and the input amplifiers are located on an electronics module in the tape recorder assembly but are not contained in the transport enclosure. The transport enclosure, which is pressurized to 2 psig with an inert gas, contains the drive motors, the record heads and head driving amplifiers, the playback heads and differential playback preamplifiers, and some of the motor control relay circuits. The playback phase equalizers (digital tracks) and the playback limiters are located on the electronics module outside the transport enclosure. The HRIR and ID channels have only playback amplifiers and limiters.

The HRIR signal, at playback speed, is a frequency modulated wave with a center frequency of 87.5 kHz with a frequency deviation of ±13.75 kHz at a maximum modulating frequency of 11.5 kHz. The ID signal, at playback speed,

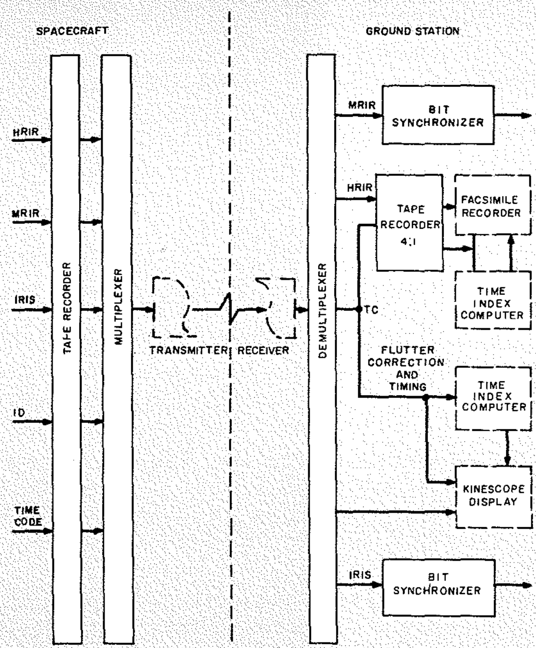


Fig. 1—High Data Rate Storage System, block diagram.

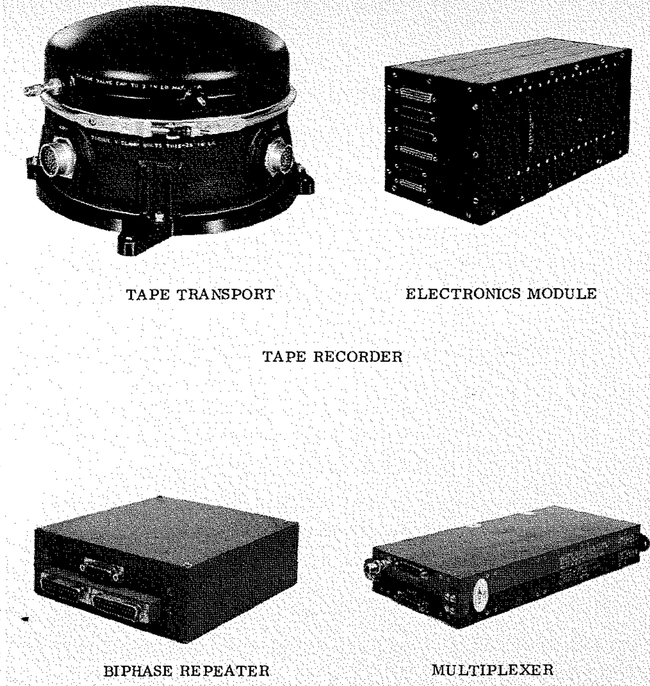


Fig. 2—Spacecraft subsystem of the HDRSS.

is a frequency modulated wave with a center frequency of 96 kHz, a peak frequency deviation of 24 kHz and a maximum modulating frequency of 51.2 kHz. The timing signal is an 80 kHz carrier, amplitude modulated by a 3.2-kilobits/s PWM waveform. The IRIS and MRIR playback channels are amplitude and phase equalized and are amplitude limited. They contain biphase data at 120 and 53.3 kilobits/s, respectively. The ID, HRIR, and timing signals from the recorder go directly to the multiplexer; while each biphase signal first passes to a repeater which removes some of the high-frequency jitter introduced by flutter in the tape recorder.

Biphase Repeater

The two biphase repeaters (Fig. 4) are identical except for the data and clock bandwidths. The incoming signal and noise are amplified, band limited, and amplitude limited. The output of the limiter is then processed simultaneously on two separate paths: a timing extraction path and a data path. In the timing extractor the incoming data is differentiated and used to drive a one-shot pulse generator. The standardized pulses from the one-shot are combined with the output of a voltage controlled (saw tooth) oscillator (VCO) in a multiplier. The output of the multiplier is filtered to develop an average bias voltage which controls the frequency of the VCO. The output of the VCO (the extracted timing) is used to sample the incoming signal in the data path. Additional flip-flop stages are used to generate pulses during tape dropouts so that the decoder at the ground station will not lose synchronism. If this feature were not provided, the ground station decoder would lose data during reacquisition of timing in addition to the data lost during the dropout. The output of the IRIS repeater, a square wave digital data signal, still in biphase-level format at 120 kilobits/s (53.33 kilobits/s for MRIR) is passed to the spacecraft multiplexer.

MULTIPLEXING

The IRIS, MRIR, HRIR, timing, and ID signals are fed to the multiplexer (Fig. 5) where they are heterodyned, filtered, and added to form a composite signal with a frequency spectrum ranging from DC to 700 kHz. The input signal amplitudes are approximately 3 volts peak-to-peak, while the composite output signal amplitude is 5.5 volts peak-to-peak.

IRIS Signals

The IRIS signal is assigned the lowest frequency multiplexer channel, covering the range from DC to 132 kHz (within −3 dB). No heterodyning is required for the signal; this multiplexer channel consists simply of a low-pass filter.

In the ground station demultiplexer (Fig. 6), the IRIS channel consists of a low-pass filter and an equalizer which corrects the phase response of the cascaded multiplex-demultiplex filters to within ± 1 μs.

Timing Signal

The second channel is the timing channel with a passband from 171 to 203 kHz. The input signal consists of an 80-kHz carrier, amplitude modulated by a PWM time code which has a pulse repetition frequency of 3.2 kHz and pulse durations of 60 and 180 μs for *0* and *1*, respectively. The timing signal is heterodyned with a 267-kHz local oscillator; the difference frequencies are selected by the 171-to-203-kHz bandpass filter.

At the demultiplexer, a 171-to-203-kHz filter selects the timing signal; no further heterodyning is required as the signal is passed directly to the time code detectors and the flutter discriminator at 187 kHz.

MRIR Signal

The MRIR signal is placed in channel 3 of the multiplexer. The MRIR biphase data, at a rate of 53.3 kilobits/s, modulate a 300-kHz local oscillator. The resulting double-sideband AM signal is passed through a 235-to-365-kHz filter. The equivalent baseband width is ±65 kHz, or approximately ±1.2 times the bit rate.

At the demultiplexer, the signal is selected by a filter with a passband from 235 to 365 kHz and is heterodyned with a 475-kHz local oscillator. The difference frequencies are selected by a low-pass 270-kHz filter and are passed through a phase corrector having a passband from 110 to 240 kHz. This channel, used in NIMBUS I for an FM video subcarrier, is phase-equalized to within ±5 μs of constant time delay. At this point, the signal is a double-sideband AM wave with a center frequency of 175 kHz. This signal is envelope detected to recover the biphase MRIR data which is then passed to a bit synchronizer for

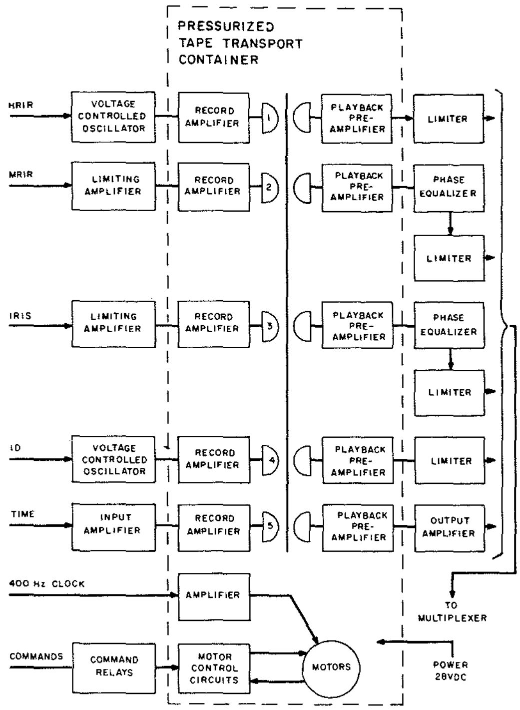


Fig. 3—HDRSS tape recorder, block diagram.

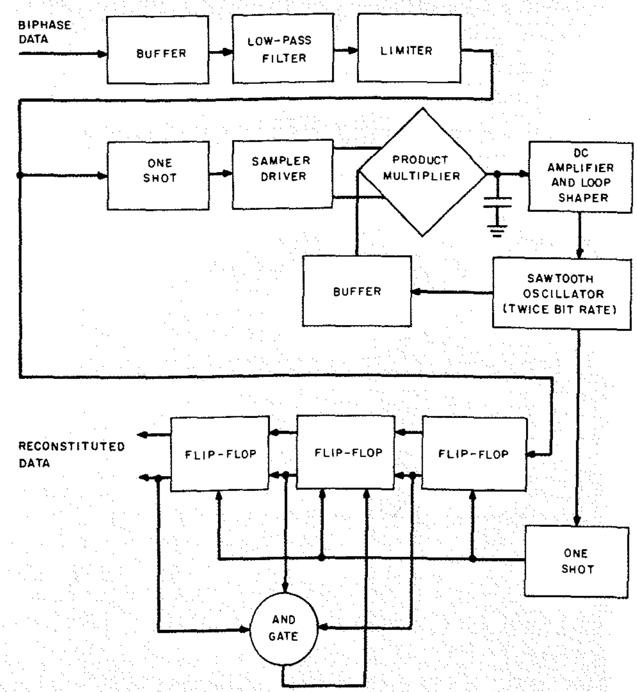


Fig. 4—Biphase repeater, block diagram.

timing extraction and conversion to non-return-to-zero format.

ID Signal

The ID signal uses channel 4 of the multiplexer, with a passband from 400 to 530 kHz. The input signal from the tape recorder is an FM wave with a center frequency of 96 kHz and a peak deviation of ±24 kHz at a maximum rate of 51.2 kHz. This signal is frequency doubled and heterodyned against a 640-kHz local oscillator; the difference terms are then passed through a bandpass filter having −3 dB points of 400 and 530 kHz. The bandwidth is such that the signal received at the demultiplexer and demodulator is a "single-sideband" FM carrier for certain combinations of peak deviation and modulating frequencies. For example, for video signals superimposed on a white (120 kHz) average field, the signal becomes SSB-FM for any modulating frequency, since all upper sidebands corresponding to the video signal are removed by the multiplexer filter which cuts off at 400 kHz.

At the demultiplexer, an input filter with a passband from 400 to 530 kHz, selects the signal which is then heterodyned with a 640-kHz local oscillator; the difference frequencies are selected by a 270-kHz low-pass filter and are fed to an equalizer having a passband of 110 to 240 kHz, which corrects the phase response of the cascaded multiplexer-demultiplexer filters to within ±5 μs of constant delay. At the demultiplexer output, the signal is an FM wave, with a center frequency of 192 kHz and a peak frequency deviation of ±48 kHz at a maximum rate of 51.2 kHz.

HRIR Signal

The HRIR signal uses channel 5 of the multiplexer. The incoming signal is an FM wave with a center frequency of 87.5 kHz and a peak deviation of ±13.7 kHz at a maximum rate of 11.5 kHz. This signal is frequency doubled, then heterodyned against an 805-kHz local oscillator and the difference frequencies are selected by a filter with a passband from 565 to 695 kHz.

At the demultiplexer, after mixing with the 805-kHz local oscillator, the HRIR signal is phase corrected to within ±5 μs and passed to an FM discriminator. At the demultiplexer output the HRIR signal is an FM wave with a center frequency of 175 kHz and a peak deviation of ±27 kHz at a maximum rate of 11.5 kHz.

RF LINK

The five HDRSS signals, translated to their selected frequency slots and combined in the multiplexer, are used to frequency modulate an S-band transmitter having an output of power of 5 watts at a frequency of 1710 MHz. An 85-foot antenna and a 3-MHz receiver are used at the ground station. The overall carrier-to-noise ratio (CNR) for the RF link is 18.1 dB. For a given CNR and fixed bandwidths, the detected signal-to-noise ratio for a particular channel is determined by the fraction of the total RF deviation which is allocated to that channel. The amplitudes of the five signals are adjusted in the multiplexer to provide the appropriate RF deviation to each channel. The HDRSS signal frequency deviations and the resulting signal-to-noise ratios are given in Table I.

TABLE I—HDRSS Signal Frequency Deviations and Signal-to-Noise Ratios

Channel No.	Signal	Peak RF Deviation (kHz)	Baseband SNR (dB)
1	IRIS Data	100	34 (P/RMS)
2	Timing	75	—
	Time Code	—	15.4 (RMS/RMS)
	Flutter	—	30 (RMS/RMS)
3	MRIR	400	28.2 (P/RMS)
4	ID	300	34.4 (B-W/RMS)
5	HRIR	200	47 (B-W/RMS)
Total Peak RF Deviation		1,075	—

GROUND STATIONS

Two command and data-acquisition stations provide coverage for nearly 93% of the NIMBUS orbital passes at an altitude of 600 nmi. The Ulaska station is located at Gilmore Creek near Fairbanks, Alaska; the other, Rosman, is located at Rosman, North Carolina. The Ulaska station acquires the spacecraft on an average of 10 out of 14 orbital passes each day. Rosman acquires the spacecraft on an average of two orbital passes a day of the four missed at Ulaska, and two orbital passes for back-up. Both stations have 85-foot-diameter parabolic antennas to track and command the spacecraft. Data transmitted by the spacecraft are received at Ulaska and Rosman and are relayed over long lines or a wideband microwave link to Goddard Space Flight

Center and to the Weather Bureau at Greenbelt, Maryland.

As the S-band signals are received, they are converted to VHF, FM-detected, demultiplexed, and passed to their respective subsystems for further processing. The ID signal is fed to an FM demodulator and then to a kinescope display unit, while the HAIR signal is recorded on a facsimile recorder. An on-line computer is used to assist in the analysis of the data and to provide latitude and longitude grids for the sensory data. The ID subcarrier is demodulated and the video is applied to the kinescope electronics and to a kinescope monitor for visual observation. In the kinescope electronics the horizontal and vertical sync pulses are detected and applied to a deflection generator. The output of the deflection generator provides sweep voltages to the kinescope in the kinescope recorder. The video signal is amplified and displayed on the kinescope.

The timing signal is fed to an envelope detector to recover the time code; an index computer uses the time code to generate a numerical print out on the ID picture. The raw timing signal is also fed to an FM demodulator which generates a flutter correction signal. This signal is applied to the ID kinescope sweep circuits to compensate for the effects of tape recorder flutter on the video time base.

The video displayed on the kinescope is projected onto the 70-mm film in the film processor with unity magnification. At the end of the picture, a decimal readout from the ID index computer is illuminated and focused onto the film below the video display. The picture size is approximately 2 by 2 inches with a vertical sweep period of 6.75 seconds and a horizontal sweep rate of 133⅓ lines per second. Each picture contains approximately 800 lines. At the end of each decimal-display exposure, the film is automatically advanced. As successive pictures are taken, the film is developed, fixed, and dryed. In approximately one minute the film is processed and displayed in the viewer. A take-up reel collects the completed pictures.

The HRIR signal, along with the timing signal, is recorded on a MINCOM recorder, and played back at one quarter of the record speed. The slow-rate HRIR is fed to an FM demodulator and then to a Westrex facsimile recorder. The slow-rate timing signal is envelope detected and used to drive the facsimile recorder motor. The PWM time code is also passed to the HRIR index computer which generates time data to be impressed on the facsimile record.

The two biphase digital signals, IRIS, and MRIR, are passed to bit synchronizers (Fig. 7) where they are filtered, bit timing is extracted, and the data is converted to non-return-to-zero format by a sampling detector. The bit synchronizers are commercially available units selected for their flutter tracking capability and low bit error rate in the presence of FM noise.

Bit timing for the biphase data is extracted by a phase-locked loop which is synchronized with the zero crossings of the incoming data stream. The extracted timing pulses are then used to sample the polarity of the incoming data.

The decoder has three selectable phase-locked-loop bandwidths to ensure optimum performance for various operating conditions. The maximum permissible frequency modulation of the bit rate for which lock can be maintained with a PCM transition density of 50%, for various bandwidths, is presented in Fig. 8. These curves determined the maximum flutter which could be exhibited by the HDRSS tape recorder while working with this decoder.

The bit error rate of the HDRSS biphase decoder, operating with FM noise, is shown in Fig. 9. The biphase channels of the complete HDRSS have been demonstrated to have a bit error rate of 1 $\times 10^{-5}$ from tape recorder input to decoder output while subjected to simulated power supply noise, sensor noise, and RF-link noise.

The decoded IRIS and MRIR data are provided, along with the extracted timing signal, to a NASA interface for transmission over the microwave link and further data processing at the Goddard Space Flight Center.

SYSTEM CHECKOUT

The HDRSS performance is verified at the system level using the HDRSS bench check unit. This unit simulates all

Fig. 5—Multiplexer, block diagram.

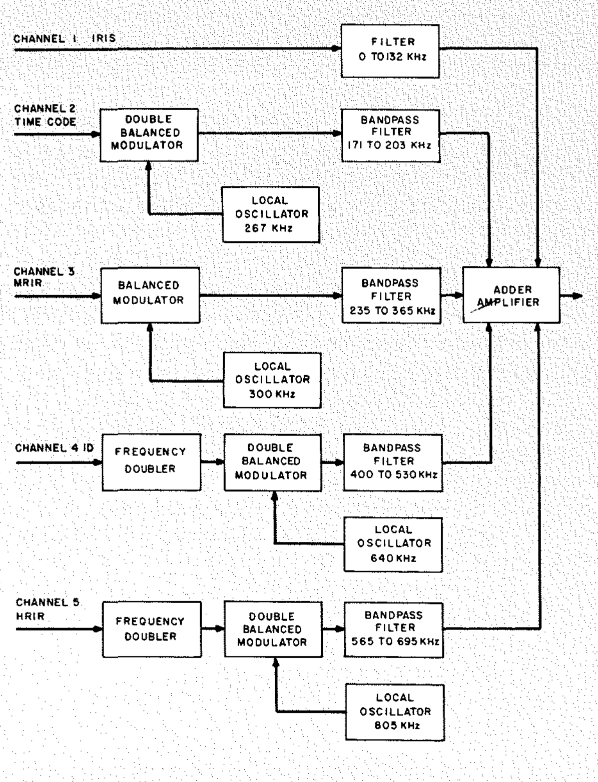
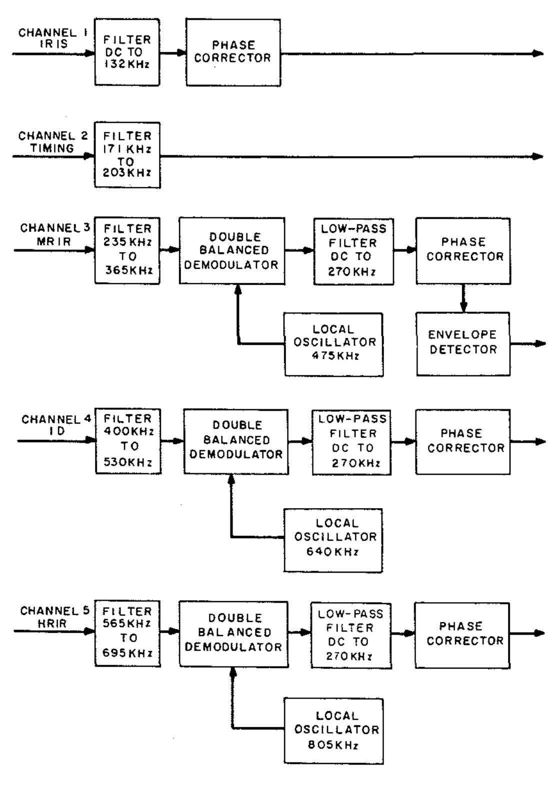


Fig. 6—Demultiplexer, block diagram.



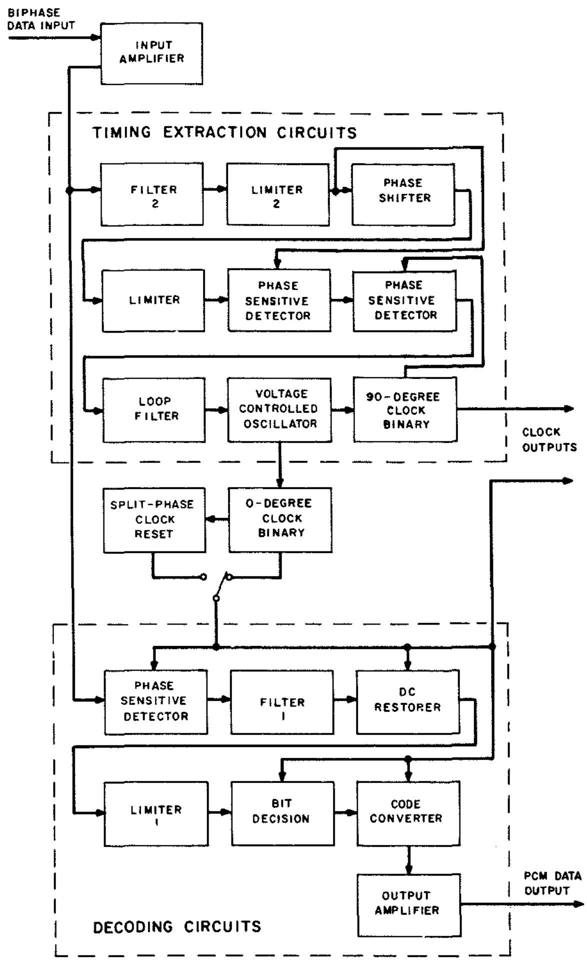


Fig. 7—Biphase decoder, block diagram.

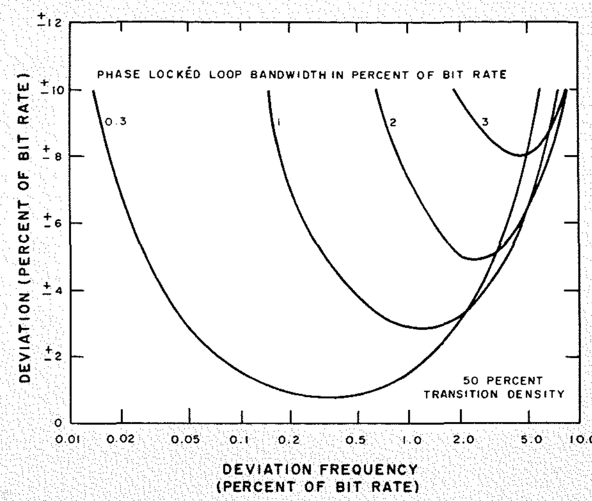


Fig. 8—Maximum lock hold-in range of HDRSS biphase decoder.

ACKNOWLEDGEMENT

The author, in writing this paper, has reported the work of a large team of people many of whom have made very significant contributions to the HDRSS program. The team included several design groups, Test Engineering, the NIMBUS Program Office, and Space Communications System Engineering. While it is not possible to list here the specific contributions, it is a pleasure to acknowledge the total effort made by each person involved in helping to bring the HDRSS to its present state of development.

spacecraft-HDRSS interfaces, including signals, commands, power, and telemetry, and provides quantitative measurement of all specified parameters for the HDRSS. All subsystem interfaces are simulated and brought to breakout boxes to facilitate fault isolation to the subsystem level.

The bench check unit includes a link simulator in which shaped noise is added to the multiplex-demultiplex interface to simulate the noise of the FM communication link. In addition, noise is added to each of the signal inputs and the power supply to verify HDRSS operation under realistically noisy conditions.

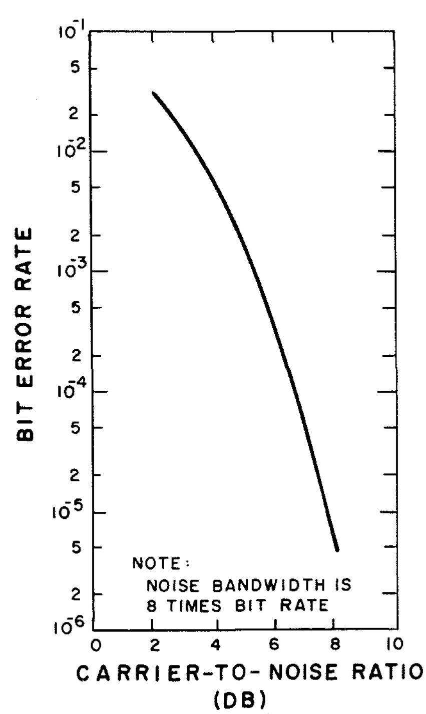
Operation of the ID video channel is demonstrated by recording a simulated test pattern signal in the HDRSS and playing it back for display in the bench check unit kinescope assembly. A hard copy of the displayed picture is also made in the kinescope recorder as a permanent record of test results. In addition, quantitative measurements are made of system linearity, drift, frequency response, transient response, and signal-to-noise ratio.

The HRIR infrared channel operation is similarly demonstrated by quantitative measurements of overall system parameters, and photographic records of the channel output signals.

Timing channel operation is demonstrated by detection of the simulated Minitrack time code with an acceptable bit-error-rate and by a subjective evaluation of the HDRSS flutter correction capability as observed on the kinescope display. The IRIS and MRIR biphase data channels are tested by generating and recording a pseudo-random code on the HDRSS tape recorder. The code is then played back and passed through the multiplexer, the simulated RF link, and the demultiplexer. The received signal is decoded and compared bit by bit with the original sequence while a count is kept of the number of errors. The bench check unit also contains detectors which indicate whether the phase-locked-loop timing extractor has fallen out of lock.

Certain parameters, such as tape recorder flutter, limiting level, intersymbol interference, harmonic distortion, etc., are measured on the subsystems before the bench check unit is used to test the HDRSS at the system level.

Fig. 9—Bit error rate as a function of carrier-to-noise ratio for the HDRSS biphase decoder.



APPLICATION OF AUTOMATIC SPEECH RECOGNITION TECHNIQUES TO BANDWIDTH COMPRESSION

For certain military applications, speech communications must be achieved with the minimum possible bandwidth. A modest amount of bandwidth compression (from 5:1 to 20:1) can be achieved with various types of vocoders; for example, the channel vocoder, correlation vocoder, formant-tracking vocoder and voice-excited vocoder. The greatest possible reduction in bandwidth is achievable, however, by a system capable of recognizing the individual phonemes in speech. These phonemes occur at an average rate of about 10 per second, resulting in a potential bandwidth reduction of about 100:1. This paper describes the current status of a continuing investigation to develop automatic speech recognition techniques. Earlier work considered isolated words only, while current work is directed toward the recognition of phonemes in continuous (conversational) speech.

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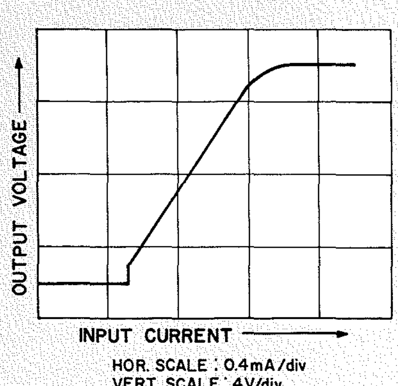
STUDIES are under way to develop a machine which will recognize conversational speech. This capability will make it possible to control machines by voice commands. Such a speech recognition machine would also permit voice inputs to computers via telephone links. Another application currently being developed by Applied Research for the Post Office Department is a spoken ZIP code recognition system which will improve the efficiency of parcel sorting operations. A speech communications system could also be developed utilizing a speech recognition system at one end of a communications link, transmitting the recognition results over a narrow bandwidth channel, and resynthesizing speech at the receiver. Such a system would require the minimum possible bandwidth for speech communications but would produce a synthetic speech, with none of the original speaker characteristics.

Final manuscript received December 1, 1966.

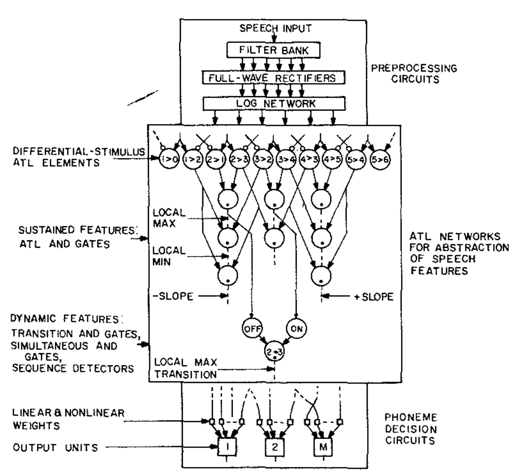
Fig. 1—ATL element transfer function.



SPEECH RECOGNITION SYSTEM

The processing technique employed in the speech recognition studies is called analog-threshold logic (ATL), because of the characteristics of the basic computing element used for most of the logical functions. The transfer function of the ATL element (Fig. 1) has an output which is linearly proportional to the net sum of excitatory and inhibitory inputs, provided that this net sum is greater than an adjustable threshold. The discontinuity in the transfer function ensures reliable discrimination between quiescence and a minimum output value, thereby avoiding sequential amplification of thermal drifts in cascaded networks.

Fig. 2—Simplified block diagram of speech recognition system.



A simplified block diagram of the speech-recognition system is shown in Fig. 2. The speech spectra are divided into 19 segments by overlapping bandpass filters whose outputs are full-wave rectified. The rectified outputs of each filter are logarithmatized before the various speech features are derived. This logarithmatization permits compression of the input dynamic range of speech without destroying the transitional properties of the speech signal. With this technique, speech signals having a 60-dB dynamic range can be processed. The ATL networks are connected in various logical configurations to abstract pertinent features of speech such as local energy maxima, local energy minima, local regions of increasing and decreasing energy, as well as the transitional properties of these features. The operations shown in Fig. 2 demonstrate symbolically how these features are abstracted. Recognition decisions are based on the presence or absence of certain critical features which are weighted appropriately in the decision networks.

CONTINUOUS SPEECH RECOGNITION

Previous Applied Research work[1] showing the feasibility of highly accurate speech recognition based on feature-abstraction techniques was achieved under conditions of: 1) isolated speech, 2) six male speakers, 3) high signal-to-noise ratio, 4) known recording characteristics, and 5) a 200-Hz to 9-kHz bandwidth channel.

The continuous-speech recognition system is being developed to operate reliably under conditions of poor signal-to-noise, possible restrictions of bandwidth, and very poor enunciation; yet the system must be capable of recognizing the speech of many speakers. Because of these restrictions, as much information should be provided in the recognition process as is possible concerning the nature of speech in which the individual sounds become increasingly obscured by these detrimental factors.

In continuous speech, sounds can easily modify surrounding sounds so that the waveform of a sound produced in continuous speech can differ significantly from the waveform of that sound produced in isolation. In reality, even for isolated words, many sounds are influenced by the context in which they appear; i.e. they are modified by preceding and following sounds.

Many additional problems exist in recognition of continuous speech; sounds can be eliminated, added, combined, or substituted for one another. Stress and intonation of the speaker also lead to variability of the spoken words. Continuous speech is characterized by considerable variations between the ideal phonetic translation of a given text and the same text as spoken by many speakers. The greatest variability occurs in the characteristics of vowels and in the liberal substitution of a voiced consonant for its minimal-pair unvoiced equivalent (such as */d/* or */t/*), or vice versa. (The diagonals enclose individual phonemes; a phoneme is the smallest unit in speech distinguishing two similar words; for example, the */p/* in *pin* and the */f/* in *fin* are two phonemes.) In the case of vowels, a very similar vowel commonly is inserted for the "ideal" vowel, or a sequence of several similar vowels is often generated. Also, vowels often are generated in acoustic contexts which tend to produce very short unemphasized vowel features. In such cases, the high-frequency features of the vowels are reduced in intensity in a manner quite similar to that found in vowel-like consonants.

The critical features differentiating very similar phonemes are usually contained in a very narrow portion of the total speech spectrum. For example, the feature distinguishing */r/* from */w/* occurs in only the low-frequency portion of the speech spectrum. Similar critical features are located only in the high-frequency portion of the speech spectrum for the distinction of phoneme pairs such as */f, s/*, */t, k/*, etc. It is easy to determine that one of a small group of phonemes has occurred from broadband spectrum features, but the critical

Authors (l to r) A. Nelson, T. Martin, R. Cox, and H. Zadell testing the typewriter output of the speech recognition system.

T. B. MARTIN graduated magna cum laude from the University of Notre Dame in 1957 with a BS in EE. He received the MSEE in 1960 on the RCA Graduate Study Program at the University of Pennsylvania and is currently pursuing studies for the PhD at the University of Pennsylvania. Since joining RCA, he has worked on applied research in the field of semiconductors. He has also contributed to a project concerned with all phases of a long-range communications system. In 1959 Mr. Martin began work in the area of pattern recognition and adaptive machines. These studies resulted in a system for speech recognition by means of analog threshold logic networks. He has developed numerous such networks for abstraction of the invariant features of patterns. These networks are constructed of a universal logic element designed by Mr. Martin that makes possible parallel processing networks that possess both analog and binary properties. He has investigated numerous adaptive devices, and adaptive decision networks. For the past two years he has served as project engineer for several contracts dealing with the recognition of phonemes and words in continuous speech. He is also the project engineer on a Post Office contract for the recognition of spoken ZIP codes. He is a consultant on a speaker authentication contract using techniques that originated from the speech recognition studies. He has authored or co-authored ten published papers and has given numerous talks to professional societies, conventions and universities. He has been awarded two patents and has several pending.

A. L. NELSON received a BEE degree from the University of Minnesota in 1958. He has done graduate work at the University of Minnesota and has completed course work requirements for an MSEE degree at the University of Pennsylvania As an undergraduate, Mr. Nelson worked in the acoustics laboratory at the University of Minnesota and at Western Electric where he worked with automatic switching equipment. In 1959 he joined the Surface Communications Division of RCA and participated in studies of vocoder evaluation, including subjective evaluation of vocoders, extensive articulation tests of many different vocoders, and talker recognition studies. He also contributed to a study of automatic speech recognition. Since joining Applied Research at RCA, he has been participating in studies of speech recognition by feature-abstraction techniques and of speech synthesis techniques which would be used with the Applied Research speech-recognition equipment. Most recently he has been participating in studies including the application of feature abstraction techniques to the problem of speaker authentication and a digit recognizer to be used in translating spoken ZIP codes in parcel post sorting instructions. Mr. Nelson is a member of the Acoustical Society of America, the IEEE, Eta Kappa Nu, and is a Professional-Engineer-in-Training (Minnesota).

R. B. COX graduated with honors from the University of Florida in 1964 with a BS in EE. He is currently pursuing studies for the MSEE at the University of Pennsylvania under the RCA Graduate Study Program. Since joining Applied Research, Mr. Cox has worked on the synthesis of filters from electronically variable delay lines, and the use of these techniques for noise reduction and pitch extraction from speech. He has also worked on an RCA sponsored program to incorporate adaption in the RCA speech recognition equipment. For the past two years, Mr. Cox has been working in the area of analog threshold logic and speech recognition by feature abstraction techniques. He has been a principal contributor to the work performed under several contracts dealing with the investigation and recognition of continuous speech. He is responsible for the development of many analog threshold logic networks for the recognition of features, phonemes and words occurring in continuous speech. He is presently working on a program for the recognition of spoken ZIP codes under contract with the Post Office.

H. J. ZADELL graduated cum laude from the University of Pittsburgh in 1959 with the BS degree in Electrical Engineering. He received the MSEE in 1963 at the University of Pennsylvania as a participant in the RCA Graduate Study Program. Joining RCA in 1959, he worked on transistor circuits and digital communications systems in the Aerospace Communications and Controls Division. He also has participated in the design of airborne digital communications systems and special-purpose computers, including interface problem analysis and compatibility studies. Since 1963, Mr. Zadell has been working in the area of neural logic and pattern-recognition systems. He has been a principal contributor to the work performed on speech recognition using Analog threshold logic feature-abstraction techniques. He has been directly responsible for developing many of the speech-recognition networks as well as the construction of several speech recognition equipments. Recently, Mr. Zadell has been engaged in the analysis of phonemes and the abstraction of words in continuous speech. He has contributed materially to the RCA Applied Research digit-recognition program. He is co-author of several papers on speech recognition using neural logic.

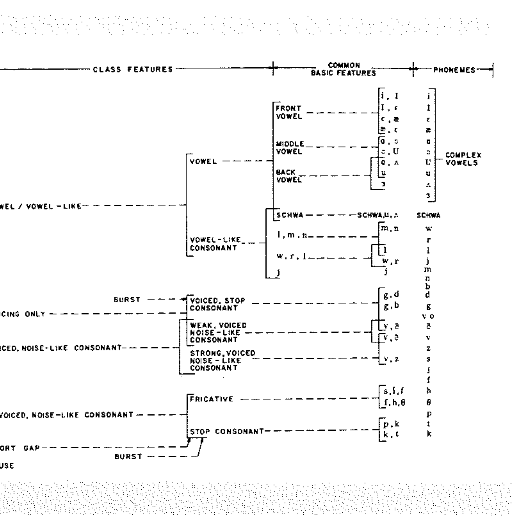


Fig. 3—Speech sounds grouped by class features, common basic features, and as individual phonemes.

phoneme-differentiating features are localized to narrow spectral regions.

Under conditions of restricted bandwidth or poor signal-to-noise ratio, the first noticeable effect in humans is a loss of discrimination between very similar phonemes. This loss does not change overall intelligibility with a proportional effect, however, since understanding in many cases does not depend on such critical discrimination. Indeed, such substitution is quite common under the very best conditions. For example, enunciation of the common word *was* by a group of humans will involve a frequent substitution between /*s*/ and /*z*/ for the final consonant. These interchanges are considered to be natural speaker substitutions which do not affect overall intelligibility. However, if conditions should occur in which noise or spectrum filtering has made it impossible to distinguish between voiced and unvoiced speech, it is sufficient to indicate that either the phoneme /*s*/ or the phoneme /*z*/ has occurred. It would be impossible to select either of the phonemes when the critical features differentiating them have been obscured. On the other hand, it would be a total loss of information to produce no response because the input speech itself did not contain acceptable criteria for the final selection of either the /*s*/ or the /*z*/.

Therefore, recognition networks for continuous speech must be organized to produce some information about degraded speech even when the quality is not sufficient to achieve final phoneme recognition with a high degree of certainty. The human under conditions of degraded speech is restricted to grouping of similar phonemes or use of context to decide which phoneme has occurred.

The capability to provide as much information as is possible, depending on the quality of the speech to be analyzed, has been achieved by organizing the feature-abstraction networks on a hierarchy of three processing levels, each level functioning optimally under different signal-to-noise and restricted bandwidth conditions.

The three processing levels abstract features which are classified as follows:

1) Broad class features
2) Common basic features
3) Unique phoneme features

Fig. 3 shows the specific types of features abstracted in this hierarchy and indicates how they are used to classify the speech sounds.

The broad class features are relatively insensitive to localized noise which involves only one or two channels of the recognizer. These features possess varying degrees of sensitivity to broader spectrum noise and spectrum filtering. Under conditions of degraded speech, abstraction of the broad class features is more reliably achieved than abstraction of either the common basic features or the unique phoneme features. Broad class features may be the only information that can be provided under poor signal-to-noise conditions.

Common basic features are those which are common to very similar phonemes such as /*ʃ*/ or /*s*/ but do not serve to differentiate between those phonemes.

Unique phoneme features are the localized spectral characteristics which differentiate between the various similar phonemes. These features can be reliably extracted for high-quality speech with sufficient bandwidth, signal-to-noise ratio, and speaker enunciation.

Two oscillograms which show the manner in which continuous speech can be segmented into broad classes are shown in Figs. 4 and 5. The first example shows the response of eight of the experimental networks for the sentence, *"Rice is often served in round bowls."* This speaker initiated the sentence with a vocal cord vibration (voicing only) followed by the vowel-like consonant /*r*/. Detection of the phoneme /*r*/, indicated in channel 8, can occur only after the transitional characteristic of this phoneme into the following vowel has been observed. The response, therefore, occurs near the boundary between the vowel-like consonant class feature and the vowel class feature (channels 2 and 3, respectively). One of the peculiarities that occur in the transition from /*aI*/ to /*s*/ is shown by the appearance of a voiced noise-like consonant at the initiation and termination of /*s*/. This characteristic is due to the inertia of the vocal tract and indicates an overlap of voicing with the unvoiced noise-like feature. This peculiarity also occurs later in the sentence for the next /*s*/. The phoneme /*v*/ produces an output on the vowel-like consonant class feature network because no high-frequency noise occurred in the pronunciation of this text.

The first example did not contain either affricatives or unvoiced stop consonants. An example containing both of these types of phonemes is shown in Fig. 5, which is a recording of the eight features for the sentence, *"These days a chicken leg is a rare dish."* Separation of the affricative into an unvoiced fricative occurs for the affricative /*tʃ*/. The /*k*/ in *chicken* is erroneously identified as an unvoiced fricative because of the absence of a high-quality stop and burst characteristic. A "voicing-only" response occurred prior to the /*tʃ*/ phoneme because of a tendency for the speaker to terminate the vowel /ʌ/ with a reduced emphasis. This response occurs quite often at the terminal portions of some of the vowels and is a valid response from a recognizer which utilizes only the acoustic properties of the speech.

Fig. 4—Responses of class features to the sentence: "Rice is often served in round bowls."

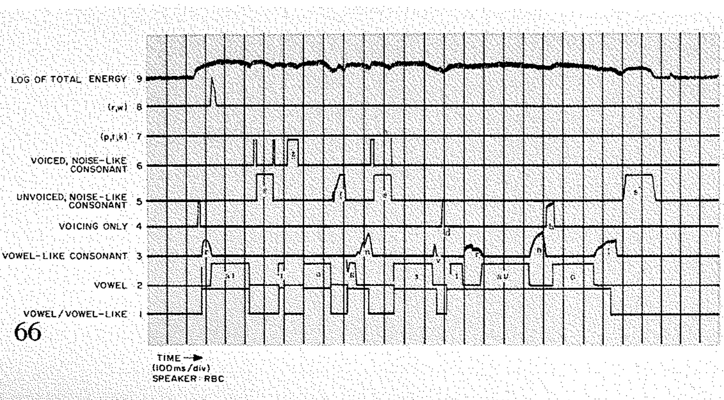
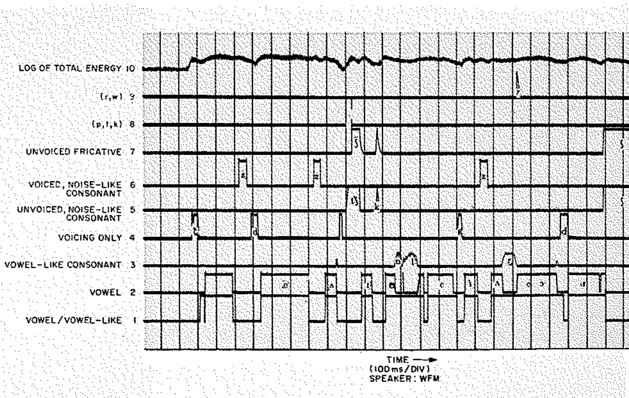


Fig. 5—Responses of class features to the sentence: "These days a chicken leg is a rare dish."



LIMITED VOCABULARIES

In addition to the major goal (recognition of phoneme-by-phoneme sequences in continuous speech) it is possible to utilize the results that have been obtained up to this time for recognition of a limited vocabulary. For many limited vocabulary applications, only the class features and a few common basic features are necessary. In many limited vocabularies no pair of entries differ only by a single phoneme, so that no critical phoneme decisions are required. Since both the class features and common basic features can be abstracted with greater accuracy than can be obtained from a phoneme-by-phoneme decision process, it is neither necessary nor advantageous to utilize phoneme decision processes for limited vocabulary applications. The minimum possible processing should be used which provides the discrimination necessary to separate the various entries in the limited vocabulary.

Common basic features and class features can be applied to the problem of recognizing the digits *0* through *9*. A recording was made of the digits spoken in isolation by eight male speakers, and these data where analyzed to develop the rules for digit recognition shown in Fig. 6. For example, the sequence recognized for *zero* consists of /z/ before either of two front vowels /I/, or /i/ followed by /r/. It was not necessary to include the final vowel after /r/ in the recognition criteria because it can be recognized uniquely from the limited vocabulary of the digits without the final vowel. As an indication of the phoneme /z/, a voiced, noise-like consonant feature was utilized. Similar broad class features were utilized for the front vowels and /r/.

Networks were then interconnected to abstract the sequences of features used for recognition of each digit. The diagram shown in Fig. 7 is an example of a three-input sequence-detecting network with its time constants adjusted to respond to the digit *4*. Except for variations due to the length of sequence presented, similar networks were developed for recognizing the remaining digits, including both *oh* and *zero*. In general, the elements may be cascaded as shown in Fig. 7 to respond to a sequence of indefinite length. For example, N1, N2 and A1 form one stage of the *4* sequence; and N4, N5 and A2 simply repeat this function for the next feature in the sequence.

The recognition criteria employed in the networks which abstract the specific features of each digit were chosen such that a response is always obtained for the expected feature, and responses may also occur for other similar sounds. For example, multiple vowel responses are allowed in those areas in which it is known that the distribution of recognition features between similar vowels overlaps for a large sample of speakers.

The sequences employed for the recognition of the individual digits are unique sequences which do not overlap. However, additional sequences are possible and must be considered for the recognition of a broad population of speakers with various dialects. For example, an alternative pronunciation of the digit *4* is /fo/, and for *5* a possible pronunciation is /fav/.

Fig. 8 shows an example of the responses of the recognition networks to one utterance of each digit. These responses occur in real time after the proper sequences of features have occurred for each digit. Both a typewriter output and a visual display were also used to monitor machine performance. When the data recorded by the eight male speakers were used, a perfect score was obtained for the total number of 253 spoken digits. The number of repetitions of the 11 digits ranged from two to five for the various speakers.

zero: z → (ɪ + i) → r
oh: (ɔ + ʌ) → ʊ → (silence ON)
1: w → ʌ → n
2: t → (i + ɛ + ɪ) → u
3: θ → r → i
4: f → (ɔ + ʊ) → ɜ
5: f → a → (ɪ + ɛ + ɜ) → v
6: s → (ɪ + ɛ) → ks
7: s → (ɛ + ʌ) → v
8: (ɪ + ɛ) → i → (silence ON)
9: n → ʌ → (æ + ɛ + ɪ) → n

SYMBOLOGY: A → B = A before B

Fig. 6—Rules for feature sequences used for digit recognition.

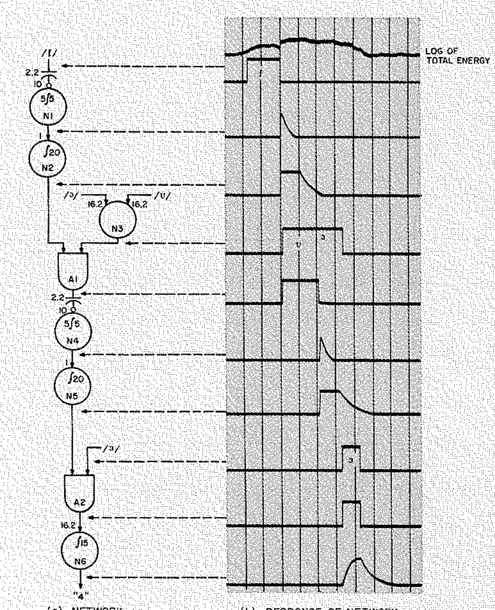


Fig. 7—Typical isolated digit recognition network.

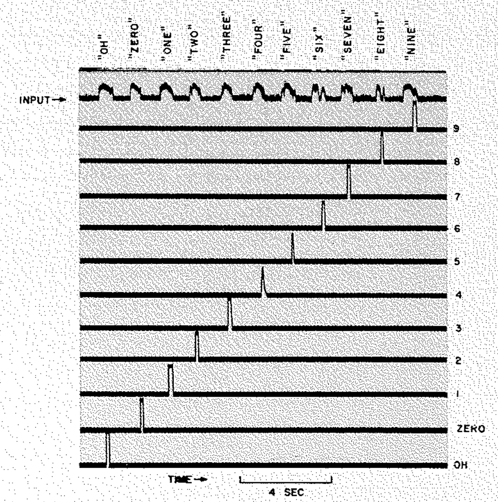


Fig. 8—Outputs of digit recognition networks in response to isolated digits.

CONCLUSIONS

A system of logic, called analog-threshold logic, was developed which originally demonstrated the feasibility of recognizing consonant-vowel-consonant words spoken in isolation. Highly accurate phoneme recognition was achieved using speech recorded by six male speakers.

Studies are now under way to extend the techniques to the recognition of continuous speech. Although this problem is made more difficult by the way the phonemes are degraded in continuous speech, the approach has already been shown feasible for the recognition of limited vocabularies and words in continuous speech. At the present time, an investigation is under way to develop recognition networks for individual phonemes in continuous speech. An automatic speech recognition capability will make possible the ultimate development of phonetic typewriters and numerous other limited vocabulary applications. Among these are ZIP code translation, voice control of all types of machines, and voice programming of computers.

ACKNOWLEDGMENT

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IMPROVING THE EFFICIENCY OF PRINTED ENGLISH LANGUAGE TRANSMISSION BY DICTIONARY ENCODING ON THE RCA 601

The ability of a dictionary encoder to reduce the redundancy of printed English text is evaluated by simulation on an RCA 601 computer. The dictionary encoder matches segments of the input text to entries of a stored dictionary which contains frequently occurring sequences of letters. The text is thus defined as the succession of code designations corresponding to the selected dictionary entries. Since, for a normal piece of text, fewer bits are needed to specify the code designations than the text itself, the encoding produces a compressed equivalent of the original input. For a broad type of English language text (news dispatches prepared for newspaper publication) the number of bits required to represent a piece of text can be reduced by 50% when using a 1,000-entry dictionary. While a better compression than 50% is theoretically possible it may be difficult to realize, but a compression of the input text to 60% to 70% of its original size appears to be easily realizable with a small dictionary.

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It is generally accepted that a large fraction of what we write or say is redundant. There are two kinds of language redundancy; the first involves the value of what is said or written, while the second does not question the subject value but appears because frequently occurring words and phrases may be specified in an abbreviated form. It is the latter type of redundancy, as crudely illustrated by the shortened form "*Yo ca rea thi sentenc*," that is of interest here. The difficulty with this example, however, is that it is not uniquely decipherable for many words are identical with the exception of the last letter. A better example, used to translate spoken language into an abbreviated written form, is shorthand as used for stenographic purposes.

The present interest in printed English redundancy was initiated by a need to improve the efficiency of command and control communication links. It is assumed that all possible messages cannot be catalogued, and thus a message dictionary type of encoder is not possible. The vocabulary normally used for command and control communication is small, however, and can be used to form a word and phrase type of dictionary. If this dictionary also includes the characters of the alphabet standing alone, any possible message can be encoded and uniquely decoded. In addition to the written language redundancy, such an encoder takes advantage of message format redundancy; headings and classification, for example; and redundancy introduced by machine operations such as carriage returns, line feed, upper case, and lower case. A word-phrase type of dictionary encoder, utilizing a small dictionary appeared to be useful for command and control communication links. Experience with this type of encoding suggested it may have more general applications, and this led to a broader evaluation which is reported here.

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BASIC APPROACH

A word-phrase type of dictionary encoder was simulated on a general purpose digital computer. The encoder operates by looking up segments of the input text in a stored dictionary that contains commonly occurring words and phrases as well as letters and letter combinations. The encoder obtains the longest possible character match between the input text and a dictionary entry and sends out a code designation for the selected entry.

On the average, the code designation can be specified with fewer bits than the entry itself, and thus a compression of the input text is obtained. The encoder output is transmitted to the receiving terminal where it is expanded by a decoder which uses a stored dictionary identical to the one used for encoding. For example, the input text, "*Work_carefully_on_the_fundamental_problem*" might appear at the encoder output as *VKCROV'ISHP'OU*RTF* if *VK* is the codeword for "*Work*," *CR* is the codeword for "*_care*," *OV* represents "*fully*," *'I* represents "*_on_the*," etc. Thus, this input text may typically be broken into sections, each matching an entire dictionary entry, as indicated by the slashes in the following sentence: "*Work/_care/fully/_on_the/_fun/da/ment/al/_problem/*."

An important feature of this type of encoding is that no information is lost, but the time required to transmit a page of text will vary with the amount of compression achieved. Although a variable transmission rate is acceptable for printed English, it would not be usable for either speech or television transmission.

Most of the dictionary encoding evaluation was done experimentally using the encoder-simulator, and encoder logic and dictionary entries were optimized as a result of statistics collected by computer simulation. The printed English used as input to the encoder consisted of Associated Press news dispatches which are distributed by teletype transmission to newspaper offices across the country. A brief analysis of a word-type dictionary encoder will be followed by a description of the encoder-simulator design and its performance for the news dispatch type of input.

DICTIONARY ENCODER ANALYSIS

A theoretical analysis of printed English redundancy was reported[1] by C. E. Shannon who estimated that printed English is at least 75% redundant. The redundancy analysis was done as a function of constraint length, one more than the number of characters of text used to predict the following character, and the 75% value was obtained for a constraint length of 100 characters. If the constraint length is limited to only eight characters, a 50% redundancy is obtained. These values include all statistical effects within the constraint length and thus would indicate the performance of an ideal encoder with the same constraint length. It is unlikely, however, that a practical encoder will utilize all statistical effects of the input language and in addition the constraint length of a dictionary encoder, one that matches

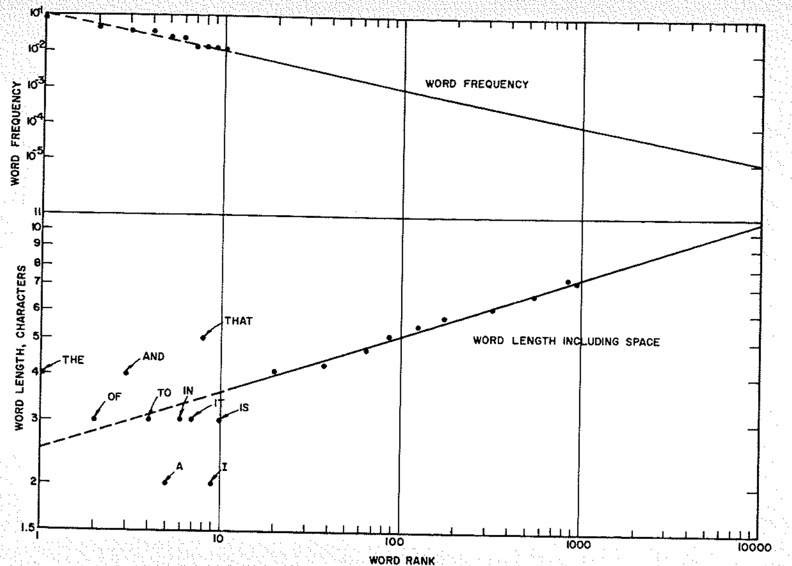

Fig. 1—Word frequency and word length statistics.

dictionary entries to segments of the input text, is not well defined. For these reasons, a better estimate of encoder performance is obtained by using word length and word frequency statistics.

An accepted empirical expression for word frequency in English is $0.1/N$ where N, the word rank, is the order of the word according to frequency of occurrence.[1] Thus, according to this expression, *"the,"* the most frequent English word, occurs 10% of the time, and the second most frequent word, *"of,"* occurs 5% of the time. A plot of word frequency versus word rank, given in Fig. 1, shows the curve, $0.1/N$, and measured frequencies of occurrence for the ten most frequent English words.[2]

Normally, frequently occurring words have fewer letters than words that are seldom used. This dependence was estimated by finding average word length as a function of word rank by using a list of the 1,000 most frequently used English words.[2] Average word lengths (which include an inter-word space) are shown by points on Fig. 1 for several word ranks greater than 10, and exact word lengths are shown for each word with rank less than 10. The straight line drawn in Fig. 1 to approximate word length L for word rank N greater than 10 is given by:

$$L(N) = 2.45\,N^{0.167} \qquad (1)$$

A word-dictionary encoder with dictionary size N_d should include all words of frequency rank N_d and less. Thus, word matches between the text and dictionary entries should occur for a fraction M of the input words where:

$$M = \sum_{N=1}^{N_d} \frac{0.1}{N} \qquad (2)$$

and the average character length of these words is

$$L_M = \sum_{N=1}^{N_d} \frac{0.1}{N} L(N) \qquad (3)$$

Values of M and L_M, plotted in Fig. 2, are based on the measured values of frequency of occurrence and word length for $N < 10$ that are given by Fig. 1. The sums are continued for $N > 10$ by using the two curves of Fig. 1. The upper limit of these sums, where $M = 1.0$, is $N = 10{,}430$ and for this N a value of $L_M = 5.75$ is obtained. This is comparable to the usually accepted average word length of 4.5 characters (which does not include a space as part of the word).

For a dictionary size N_d, a total of $\log_2 N_d$ bits are required to specify any of the N_d entries when using a constant length code. Thus, for the fraction of words M, when a match is obtained the average encoder output/input data rate ratio is:

$$R_W = \frac{\log_2 N_d}{L_M K} \qquad (4)$$

where K is the number of bits used to specify each input character. Values of R_W for $K = 5$ and $K = 6$ are shown in Fig. 3. These may be interpreted as the data rate reduction possible if the text vocabulary is limited to the N_d most frequent words.

When there are no restrictions on the vocabulary of the input text, a word match is not obtained for a fraction, $1 - M$, of the input words. The average length of these words is:

$$L_x = \sum_{N=N_d+1}^{10{,}430} \frac{0.1}{N} L(N) \qquad (5)$$

If these words were transmitted letter by letter, the average output/input data ratio is:

$$R_x = \frac{\log_2 N_d}{K} \qquad (6)$$

which is an increase of data rate. If, however, pairs of letters called digrams, and triplets of letters called trigrams, and longer letter combinations are included in the dictionary, words may be spelled in groups of letters. The simulation analysis discussed later shows that a data-rate ratio resulting from letter combination entries:

$$R_{LC} = \frac{\log_2 N_d}{2.32\,K}, \quad N_d > 400 \qquad (7)$$

can be achieved if 400-letter combination entries are included in a dictionary. The 400 entries include every letter standing alone and every letter preceded by a space. The factor 2.32 of Eq. 7 arises because the average length of character agreement, the match length, of the letter combination entries was observed to be 2.32 characters. The number of entries, 400, was selected by simulation analysis as a good proportion of a 1,000-word dictionary to devote to letter combinations.

For a word-letter-combination diction-

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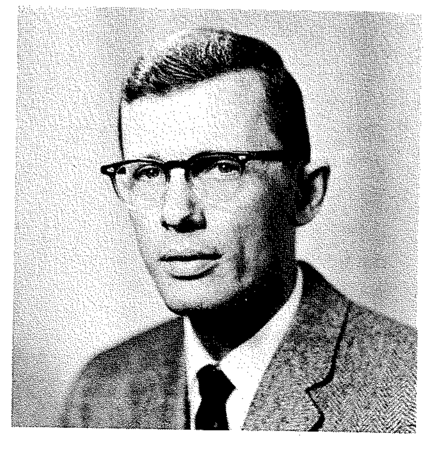


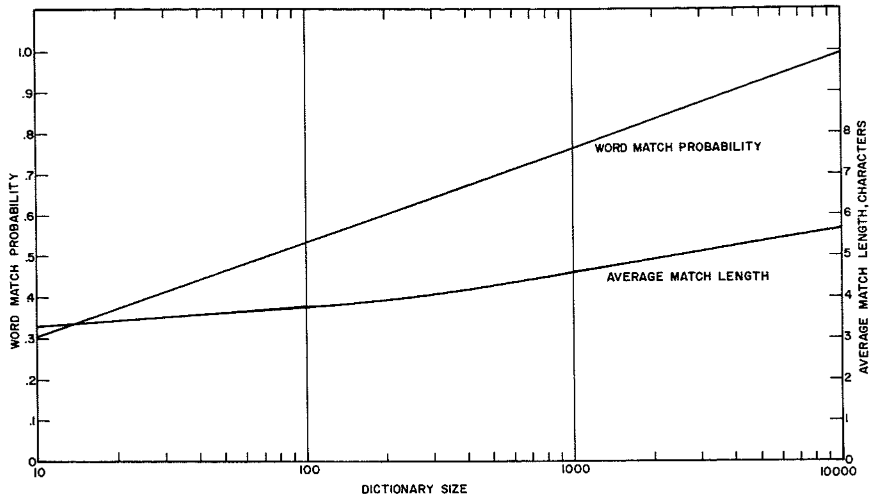


Fig. 2—Word match probability and match length for a dictionary encoder.

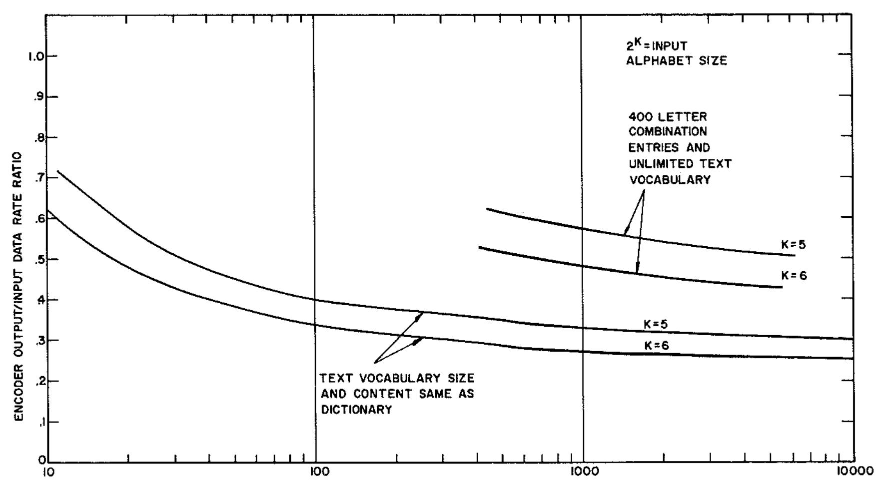


Fig. 3—Dictionary encoder performance as a function of dictionary size.

ary encoder using 400-letter-combination entries, the overall output/input data rate ratio is:

$$R = R_W' M_W + R_{LC}(1 - M_W) \quad (8)$$

where:

$$R_W' = \frac{\log_2 N_d}{K \cdot L_W}$$

$$M_W = \sum_{N=1}^{N_d - 400} \frac{0.1}{N}$$

$$L_W = \sum_{N=1}^{N_d - 400} \frac{0.1}{N} L(N).$$

Fig. 3 shows curves of R for $K = 5$ and $K = 6$ as a function of total dictionary size. Since 400-letter-combination entries are included in the dictionary independent of total size, the curves are more accurate for $N_d = 1{,}000$. This effect is probably small (except for $N_d = 500$) and the effect of punctuation, numerals, and machine operations in the text and multiple word entries in the dictionary, factors not included in Eq. 8, may be of more importance. Fig. 3 will be compared to simulation results discussed in the last section.

ENCODER-SIMULATOR DESIGN

The encoder-simulator design as well as its dictionary entries are greatly influenced by the characteristics of the input text. The basic alphabet of the encoder must be the same as that of the input, and for the news dispatch text this consists of the 64-character linotype code composed of letters, numbers, punctuation, and typesetting symbols. The original text, obtained on punched paper tape, was not modified in any way and was first translated to the computer code. A partial listing of the linotype alphabet with the assigned computer symbols is given in Fig. 4 to permit interpretation of later examples.

A general block diagram of the encoder-simulator, which is programmed in the RCA 601 assembly language, is given in Fig. 5. This includes subsections to read the dictionary and the input text, logic operations to recognize key characteristics of the text, a dictionary search routine, and a statistics collection subprogram. The output from the simulator consists of the original text, the encoder output, and the dictionary including the collected statistics.

The dictionary input coding recognizes an arbitrary number of major dictionary sections by the presence of divider cards among the dictionary entries. Four major dictionary sections are used: a word dictionary, a capital word dictionary, a suffix dictionary, and a special dictionary. Use of these dictionary sections is controlled by the encoder logic. The presence of a space in the input text identifies the beginning of a word and a search of the word dictionary follows. A word is also identified by a hyphen (not at the end of a line), and a carriage return and elevate; but in these instances the space, understood to precede entries of the word dictionary, is omitted. If none of these characters (space, hyphen, carriage return and elevate) is detected, the search follows in the suffix dictionary which includes a majority of the letter combination type of entries. The suffix dictionary not only supplies word endings but is also used to spell out words not found in the word dictionary after the space and first letter or letters are taken from the word dictionary. The use of the word and suffix dictionaries changes when the beginning of a sentence or paragraph is detected. If the first letter of the new sentence or paragraph is capitalized the next entry is taken from either the capital word dictionary or the word dictionary, and if it is not a capital letter the next entry comes from the suffix dictionary. This option permits any entry of the word dictionary to be used at the beginning of a sentence or paragraph as an entry with its first character capitalized.

Capitalized words not at the beginning of a sentence or paragraph are recognized prior to searching the word dictionary, and instead the capital word dictionary is searched. If a good match (more than the capital letter) is not found, the word dictionary is searched; and any entry found there may be capitalized by prefixing the entry by an entry of the special dictionary. If a good match is not found in the word dictionary, the capital letter, consisting of three characters (upper case, the letter, lower case), is taken from the capital word dictionary.

The special dictionary also contains four kinds of interword space entries which are used in linotype material to

INOTYPE CHARACTER	COMPUTER SYMBOL
DD-THIN-SPACE	"
ARRIAGE RETURN	%
LEVATE (LINEFEED)	↑
M SPACE	# OR
N SPACE	@
YPHEN	-
PACEBAND (SPACE)	
HIFT (UPPER CASE)	<
NSHIFT (LOWER CASE)	>

Fig. 4—Partial listing of linotype alphabet and assigned computer symbols.

justify lines of type. The same spacing, however, is generally used throughout any one line and thus only changes of space variety need be sent. The four space entries in the special dictionary include two single-character spaces (spaceband and add-thin-space), one double-character space (spaceband, En space), and one triple-character space (spaceband, En space, spaceband). The code designations of these entries appear in the encoder output when a change of the space character(s) is found in the input text. The current space character(s) is stored in a space register, and at the decoder the contents of this register are understood to precede all entries from the word dictionary and capital word dictionary. This applies unless the character(s) of the input text preceding a word is a hyphen or a carriage return and elevate when the space is omitted, or if a new paragraph is detected in which case the indentation is included with the carriage return and elevate entry and no space is needed.

The narrow columns used for newspaper text cause many words to be hyphenated at the end of a line. These words cannot be found in the word dictionary unless the hyphenation is recognized before the dictionary search. This is done and the word is adjusted to appear in continuous form preceded by the hyphenation characters (lower case, hyphen, carriage return, elevate). The hyphenation characters are encoded using the special dictionary which includes nine entries that indicate hyphenation of the following word after its first, second . . . , ninth character. The remaining unhyphenated word is now located in the word dictionary using the standard search procedure. Obviously, much of the logic in the encoder-simulator, such as the hyphenation routine, is peculiar to the input text used, and may not be required for other applications.

The same dictionary search routine is used for each of the four dictionaries. It begins with a jump to the dictionary entry which matches the first character of the input text. The following entries are searched until a character disagreement occurs at a character length equal to or less than the length of a previous agreement. Multiple word entries are identified by a star following the last word, and spaces between words of the entry are changed to agree with the space register if it is one of the single character spaces. Otherwise, because multiple character spaces are seldom used, the multiple word entry is ignored. After the best match is achieved between the text and an entry the code designation for that entry is moved to an output register, and the coding continues with the next character of the input text.

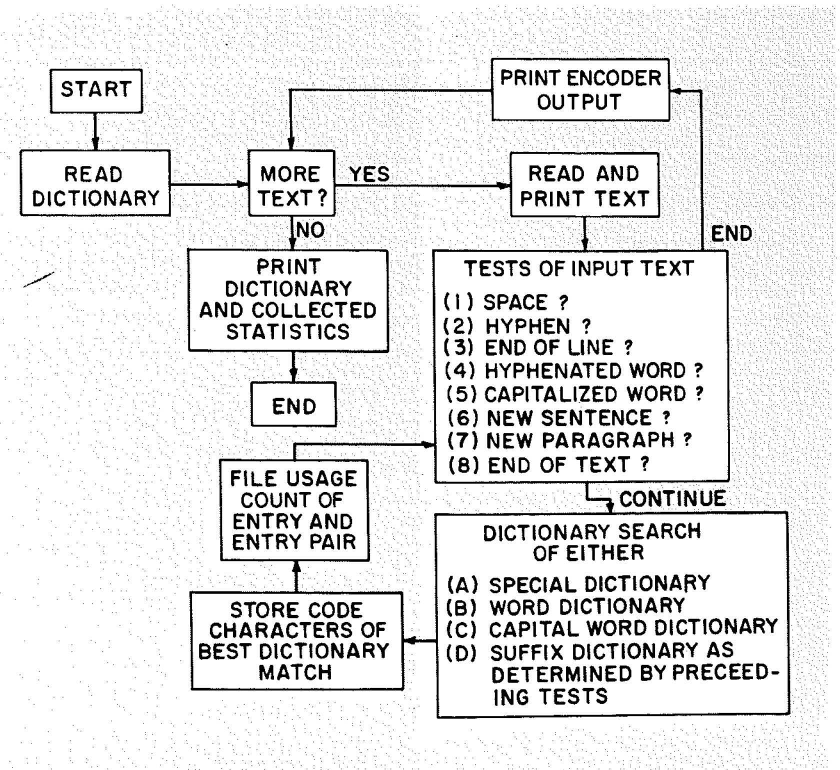
The preceding description covers the actual encoding operations. The additional function of the encoder-simulator is to collect text statistics that are used to improve the encoder logic and dictionary entries. This is done by recording the usage count of each dictionary entry and also recording the code designations of the entries that followed it in the input text each time it was used. For a 1,000-entry dictionary a million combinations of adjacent entries are possible, but only a small part of this number will normally occur. Thus, storage locations are not reserved in advance for counting the occurrence of each pair of entries, but a new storage location is made available each time a new pair of entries is used. This is done by spreading out the dictionary, which is originally stored in a close-packed form. The dictionary print-out, Fig. 6, which occurs after all the input is encoded (or the memory is full) includes the original dictionary entries as well as the collected statistics. An example of recorded usage counts is given in Fig. 6 by the word *"today"* which occurred 8 times in the previously encoded text, and one of these times it was followed by the word *"to"* which has the code designation *"RC."*

ENCODER-SIMULATOR PERFORMANCE

If little were known about the text statistics, the initial encoder dictionary could be limited to the characters of the alphabet. The usage counts for pairs of entries would then show which character pairs would make good entries in an expanded dictionary. If more statistics were then collected using the expanded dictionary, longer entries in a still larger dictionary would be suggested. If this process were repeated many times for a representative sample of input text, an optimum dictionary would be developed for that type of text. Ideally a dictionary of fixed size should be altered after each iteration to increase the weights of lighter entries at the expense of heavier ones. In this connection entry weight is defined as the frequency of usage for the entry multiplied by the number of entries needed to encode the entry if it were omitted from the dictionary. By such an optimization some entries which are used infrequently will be retained if they otherwise must be spelled by several entries, and other entries although they are used more frequently will be discarded because they are easily spelled.

For the present encoder simulation, the initial dictionary contained 400 of the most frequent English words[2] and

Fig. 5—General block diagram of dictionary encoder.



```
L
E     E
N  C  N
G  O  T       USAGE COUNTS
T  D  R
H  E  Y

05 Q/ THOSE   0004 LX01PU01A&01=P01
05 R= THREE   0002 KF01Z101
07 RA THROUGH      0004 ==01X&01=D01XU
04 RB TIME    0008 %S01Y101ZF01RC02TP
02 RC TO 0038 MT01HB01HE01!N01QA01XB
                VE01VH01NH01=001E!02
04 RD TO%•    0011 EP01X=01J001KB01GU
04 RE TO A•   0003 1R01•Z01LK01
05 RF TO BE•  0002 DJ01PZ01
06 RG TO GET•
06 RH TO THE•0005 DK01DA01XB02•B01
05 RI TODAY   0008 RC01!I01ZZ01ZX01TK
03 RJ TOO     0002 EJ01=•01
```

Fig. 6—Section of word dictionary and collected statistics.

200 commonly occurring digrams and trigrams in addition to the letters of the alphabet. The digrams and trigrams were taken from lists that were tabulated for cryptographic purposes.[3] This initial dictionary was then modified by adding and removing entries as suggested by the entry weights. The optimization was done by inspection of the collected statistics and not by the computer. Thus, an exact optimization was not obtained; but certain unappropriate entries (proper names, for example) which occurred frequently in one piece of text, but seldom elsewhere, were avoided. The amount of text used for optimization purposes was gradually increased until 115,000 characters were used in the later steps. All of this text, consisting of 50 news dispatches from 15 August 1966, was picked at random. There appears to be little overlap in subject material.

An example of this input text, one sentence followed by the first word of the following paragraph, is given by Fig. 7. The 208 characters of input text at the top of Fig. 7 produce the 64 character pairs of encoder output shown in the center of the same figure. Each character pair specifies one dictionary entry. For this example there are 1,340 entries in the dictionary, and thus $\log_2 1340 = 10.4$ bits are used to specify an entry. The compression achieved for the example of Fig. 7 is an output/input data rate ratio:

$$R = \frac{64}{208} \cdot \frac{\log_2 1340}{6} = 0.53 \qquad (9)$$

The encoder operation is easily followed by examining the additional simulator output at the bottom of Fig. 7. This consists of the original input with square brackets, [, inserted to show the matches obtained between the text and dictionary entries. All characters between a pair of brackets are encoded as one dictionary entry. This additional output also shows which entries of the word dictionary are capitalized (they are preceded by [<>] which is encoded by an entry of the special dictionary) and how hyphenated words are modified prior to the dictionary search. Examples of capitalization and hyphenation are given by the words *"sea"* and *"border"* in the simulator output at the bottom of Fig. 7.

The compression of 0.53, the output/input data rate ratio, obtained for the example of Fig. 7 is similar to the overall compression of 0.53 obtained for the 115,000 characters of text used for optimization of the 1,340-entry dictionary. This figure was further checked, to guard against undue bias in dictionary selection, by encoding an additional 13,000 characters of text and a compression of 0.55 was obtained. Judging from this latter result and the fairly consistent values of compression obtained for the 50 news dispatches used for dictionary optimization, a compression of 0.53 appears to be representative of the selected dictionary. With further optimization of dictionary entries and minor improvements in the encoder logic it is estimated that a compression of 0.5 can be obtained with a dictionary containing 1024 entries.

The dependence of compression on the dictionary size was determined by removing 509 entries from the originally developed encoder-simulator dictionary. The choice of which entries to remove was based on a weight threshold of 40 units. Thus, any entry with a weight (as defined earlier) less than 40 was removed, and the resulting dictionary contained 831 entries.

As for the original dictionary selection the removal was done by inspection and thus is subject to error because the removal of an entry affects the weights of the remaining entries. Examination of the entry usage counts using the smaller dictionary, however, suggested that a reasonable optimization (comparable to that of the larger dictionary) had been obtained. The same 115,000 characters of input used earlier were encoded with the smaller dictionary and a compression of 0.57 was obtained. If these two points; a compression of 0.53 for a dictionary of 1,340 entries, and a compression of 0.57 for a dictionary of 831 entries; are compared to the curves of Fig. 3, the variation of compression with dictionary size is similar to the theoretical curves. A correspondence in the actual compression values should not be expected because of differences between the simulator input and the theoretical model. It appears, however, that a compression gain resulting from multiple word entries almost offsets losses produced by numbers and punctuation in the text, and thus fairly close agreement is actually obtained.

It is interesting to compare the dic-

Fig. 7—Example of encoder-simulator input text and output code.
upper: input text
center: encoder output
lower: input text divided by square brackets to indicate dictionary entries used for encoding.

```
208
<I>N <D>AMASCUS, <S>YRIA, AN ARMY%•SPOKESMAN"REPORTED"A"THREE-%•HOUR BAT
TLE BETWEEN <S>YRIAN BOR>-%•DER"POSITIONS"AND"TWO"<I>SRAELI%•GUNBOATS ON
 THE EASTERN SHORE OF%•THE <S>EA OF <G>ALILEE. • # #/#/•%•#<T>HE


64
XPXE'E'O•SOYYD300K%UZXAQBA3QPS=IN&=WR=4MH/==BUCAYD300K'F=PCE0E&'=IN=-=0%
RWXO0W0B&V:AZ'HC•A'Q0W==M=EW'02=+LPE&DLV=LOWLTXL:O&RZMYM


[<I>N[ <D>[AM[AS[CU[S,[ <S>[Y[R][A[,[ AN[ ARM[Y%•[SPOKESMAN["[REPORTED["
A["THREE[-%•[HOUR[ [BATTLE[ BETWEEN[ <S>[Y[RI[AN[ >-%3[BO[RD[ER["[POS[IT
ION[S"AND["TWO["<I>[S[RA[EL[I[%•[GUN[BO[AT[S[ [ON THE[ E[AS[TER[N[ SHOR[
E[ OF%•THE[ <>[SEA[ OF[ <G>AL[I[L[EE[. • # #/#/•%•#[<T>HE[
```

tionary encoder performance with linotype text to the performance of a variable length code such as described by Huffman.[4] If a variable length code were used on a per letter basis (shorter codewords to represent frequently occurring letters) the lower bound on the average number of bits required per letter of text is given by:

$$F_1 = -\sum_{i=1}^{64} p_i \log_2 p_i \qquad (10)$$

where p_i is the probability of letter i. For the linotype text used for dictionary encoder simulation an F_1 of 4.99 bits is obtained. The ratio of this number to 6, the number of bits required by the constant length linotype code, is 0.83. Thus, the upper bound on the data rate reduction using a variable length code is 17%, but this could only be obtained if the variable length code matched the letter probabilities exactly. An upper bound on the average number of bits per letter for a Huffman code is $(F_1 + 1)$ or 5.99 bits, and this indicates that in the worst case no compression is obtained.

A variable length code may also be considered as an adjunct to dictionary encoding. The lower bound on the number of bits required to represent a dictionary entry would be:

$$I_1 = -\sum_{i=1}^{N_d} p_i \log_2 p_i \qquad (11)$$

where p_i is the probability of usage for entry i, and N_d is the number of dictionary entries. For the smaller dictionary, 831 entries, a ratio of I_1 to $\log_2 831$ (the maximum value of I_1 corresponding to $p_i = 1/N_d$) of 0.92 was obtained and for the larger dictionary the ratio of I_1 to $\log_2 1340$ was 0.91. These numbers indicate that a variable length code could reduce the dictionary encoder output data rate by as much as 8% or 9%. This is equivalent to a maximum possible improvement in the overall encoder compression of about 4%.

Further insight into the encoder-simulator operation is provided by the word frequencies of the 25 entries from the word dictionary with largest frequency of usage. This is of interest because it illustrates the effect of multiple word dictionary entries on word statistics. Fig. 8 shows these observed word frequencies and the curve $0.1/N$ used earlier for encoder performance calculations. The total frequency of usage of the word "*the*" is also shown on Fig. 8. This point, which is obtained by adding the usage of all entries containing "*the*," is close to the previously measured frequency of "*the*" plotted in Fig. 1.

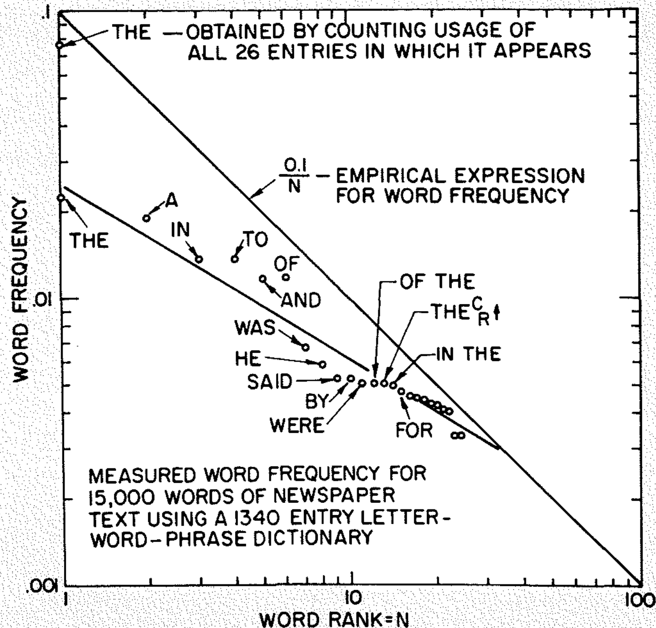


Fig. 8—Word frequency as measured by encoder-simulator

An indication of the encoder-simulator complexity is given by its dictionary memory requirements. The larger dictionary, 1,340 entries, requires 2,300 half-words (24 bits each) of storage; the smaller dictionary, 831 entries, requires 1,600 half-words. These memory sizes are required for normal encoder operation which does not include statistics collection.

DISCUSSION

The simulation of a dictionary encoder indicates that a printed English compression of 50% is realizable for a broad type of English language text when using a 1,000-entry dictionary. Best performance of the dictionary encoder is obtained after careful dictionary optimization for the type of text to be encoded. This can be done by using computer collected statistics or could be entirely done, including the selection of better dictionary entries, by a computer. For input texts with a limited variety of subject material a compression better than 50% should be expected when using a 1,000-entry dictionary. The gradual variation of compression with dictionary size indicates that a compression of the input text to 60% to 70% of its original size should be possible with a dictionary consisting of a few hundred entries. While not considered a part of the encoder evaluation it should be noted that the compressed encoder output will be more susceptible to transmission or processing errors than the original text because less language redundancy is available to facilitate visual error correction.

A significantly better compression than 50% for a broad type of text is possible as demonstrated by Shannon,[1] but this would require additions to the dictionary encoding logic. One such addition would be a memory of the recently encoded text. This could be practically done by searching back in the encoder output to find identical strings of code words, and replacing the repeated string by a reference to the earlier occurrence. Such a technique would reduce the number of code words needed to send words or phrases which are not in the dictionary, but are used more than once in a particular piece of input text.

ACKNOWLEDGEMENT

This investigation of printed English compression was done in the Communications Research Laboratory of RCA Laboratories and received support and assistance from Dr. K. H. Powers, Director, as well as other members of the Communications Research Laboratory. The present paper benefited from the constructive comments and suggestions of several reviewers. The paper tapes of the input text and assistance in interpreting them were supplied by Mr. James Monteverde of Trenton Times Newspapers. These tapes were repunched in a form suitable for computer use with the help of Mr. J. C. Hepburn and Mr. George Shapiro of RCA Communications.

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2. G. Dewey, *Relative Frequency of English Speech Sounds*. Cambridge: Harvard University Press, 1923, Table 4.
3. Fletcher Pratt, *Secret and Urgent*. New York: The Bobbs-Merrill Company, 1942.
4. D. A. Huffman, "A Method for Construction of Minimum Redundancy Codes," *Proc. IRE*, vol. 40, pp. 1098-1101, September 1952.

DESIGN AND INTEGRATION OF A LOW-POWER DTL-JK FLIP-FLOP

A flexible integrated flip-flop circuit has been successfully designed and fabricated. Close marriage of circuit design and integrated component limitations had to evolve to enable a practical design implementation. Improvements in component fabrication capability had to be realized. Yield improvements due to greater circuit complexity also had to be realized. Worst-case circuit designs using loose tolerance components had to be accomplished. The overall results of these efforts is a flip-flop design which, from a logic and systems point of view, allows simplified improvision of a large variety of counter, register, and control operations.

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Integrated Digital Circuits and Subsystems
Defense Microelectronics, DEP, Somerville, N.J.

THE diode transistor logic (DTL-JK) flip-flop (see Table I, *Glossary of Terms*) is one of a complete family of low-power integrated logic circuits designed at RCA. Other family members include a dual four-input gate, a quad two-input gate, a dual four-input buffer gate, a dual three-input phantom-*or* gate, and a gate-input expander module. The design criteria used for the flip-flop circuit were minimum power dissipation, 1-MHz minimum clock complementing rate, high-threshold and pulse-noise immunity at all inputs (primarily at the clock inputs), maximum logic flexibility in counter, register, control and other applications, and operation over a −55°C to +125°C temperature range at power supply voltages from +3.8 to +6.3 volts.

LOGIC CAPABILITY

Fig. 1 shows a schematic diagram of the flip-flop circuit and Fig. 2 shows a photomicrograph of an integrated flip-flop chip. The circuit of Fig. 1 may be operated as either a DC *set-reset* (RS) flip-flop, or as a clocked JK flip-flop having DC set-reset capability. In the RS mode, the circuit provides two DC *set* terminals and two DC *reset* terminals. The features of the flip-flop in the clocked JK mode are as follows:

1) Split-clock or single-clock input capability.
2) Internal self-steering of clock trigger circuit.
3) Two external steering controls at each clock trigger circuit.
4) A DC set-reset capability for either high or low clock inputs.
5) Steering control independent of clock-input state (clock triggering impossible without proper steering levels).
6) Clock negative-edge triggering (false triggering from steering lines impossible).
7) A JK mode master-slave equivalency (race conditions impossible).
8) Fan-out of each flip-flop output to six-unit gate loads under all operating conditions (all gate and buffer types in the logic family present a unit gate load).
9) Simple loading rules:
   a) Any DC set or reset input to the flip-flop represents one unit gate load.
   b) Any steering input to the flip-flop represents one-half of a unit gate load.
   c) In the single-clock mode of operation, the common-clock input line of the flip-flop is defined as two unit gate loads.
   d) In the split-clock mode of operation, each clock input line of the flip-flop is defined as two unit gate loads.
10) Well defined output logic levels: *1* level typically 0.6 volt less than V_{cc}, *0* level typically 0.1 volt ($V_{ce\text{-}sat}$).

Final manuscript received December 21, 1966.

CIRCUIT CAPABILITIES

Low-Power Dissipation

The power dissipation in the flip-flop circuit with the clock and steering lines low is typically 8 mW for a V_{cc} of 4.0 volts.

High-Noise Immunity

The clock-line threshold immunity to noise is typically 1,500 mV over the full operating temperature range of the circuit. The threshold immunity increases to approximately twice this value for noise pulses less than 100 ns wide. The DC set-reset line has a typical threshold immunity greater than 1 volt over the full operating temperature range. The threshold level for this line increases approximately 50% for noise pulses less than 50 ns wide. For the steering line, the threshold immunity is typically 1,200 mV at +25°C, and 700 mV at −55°C with the line high; or 1,900 mV at +25°C, and 1,100 mV at +125°C with the line low. For a high-steering line, the threshold immunity is increased approximately 50% for noise pulses less than 100 ns wide.

Buffer-Circuit Output Impedance

The use of buffer-circuit output impedances at both flip-flop outputs assures relatively constant output-drive characteristics under conditions of high capacitive loading and also provides a low impedance to any AC coupled noise. These circuits provide complementary outputs and provide logic swings compatible to all gate and buffer family members.

One MHz Complementing Rates

The flip-flop will properly complement for clock signal rates up to 1 MHz over the full temperature and fan-out range in a system utilizing compatible family

GLOSSARY OF TERMS

SET-RESET FLIP-FLOP: Two cross coupled inverting gates, each having additional inputs to permit level control of the set and reset operations.

JK FLIP-FLOP: A flip-flop having provision for proper toggling operation using only a clock signal input (J's are the steering control inputs on the O-side steering gating; K's are the steering control inputs on the I-side steering gating).

STEERING CONTROL: External and internal flip-flop signal connections to permit control of a clocked trigger input signal.

SPLIT-CLOCK CAPABILITY: A flip-flop having separate I and 0 clocked input triggering circuits.

MASTER-SLAVE FLIP-FLOP: A double flip-flop (dual-rank) mode of operation which permits entry into the "master" flip-flop during one clock input signal phase and transfer from "master" to "slave" (output) flip-flop during the opposite clock input phase. Such operation avoids the occurrence of possible race conditions in shift-register, counter, and control applications.

SUBSTRATE: Basic integrated-circuit silicon-wafer starting material.

FORCED BETA (β): Circuit operation in a condition where both base and collector currents are fixed by circuit conditions rather than natural transistor beta conditions.

FORCED ALPHA (α): Circuit operation in condition where both emitter and collector currents are fixed by circuit conditions rather than natural transistor alpha conditions.

members. The flip-flop provides typical *turn on* and *turn off* propagation delays of 150 ns.

Reliable Triggering Under Widely Varying Clock Waveshape Conditions

The trigger circuit is relatively insensitive to variations in the clock trigger pulse. Satisfactory operation can still be obtained when the clock pulse has been degraded to the point that the amplitude is less than 2.5 volts and the trigger-edge transition time is greater than 300 ns.

CIRCUIT OPERATION

The flip-flop circuit (Fig. 1) consists of two inverter gates cross coupled to form a bistable circuit, together with a trigger-control circuit. (In the discussion that follows, positive logic is assumed; $0 \approx 0$ volts and $1 \approx V_{cc}$.)

Each inverter gate circuit operates as a three-transistor inverter amplifier, with an emitter-follower circuit used to supply the output drive for the high-level (*1*) state and a saturated transistor used to supply the low level (*0*) output. Gate fan-in is provided by three-input pnp substrate transistors. The gate circuit output goes to a low level when all fan-in points are at a high level, and goes to a high level when any fan-in point is at a low level. The use of an emitter-follower, high-level output circuit provides relatively constant turn-off speed under conditions of high capacitive loading. Moreover, any noise voltages that occur during the high level output of the gate, whether as a result of AC coupling or of loading variations, will be damped by the low-impedance emitter follower. Another advantage of the emitter-follower load circuit is that it provides negligible dummy-load current to the saturated output transistor during the low-level output state. Noise voltages occurring during the low-level output of the gate will be damped by the low impedance saturated output transistor.

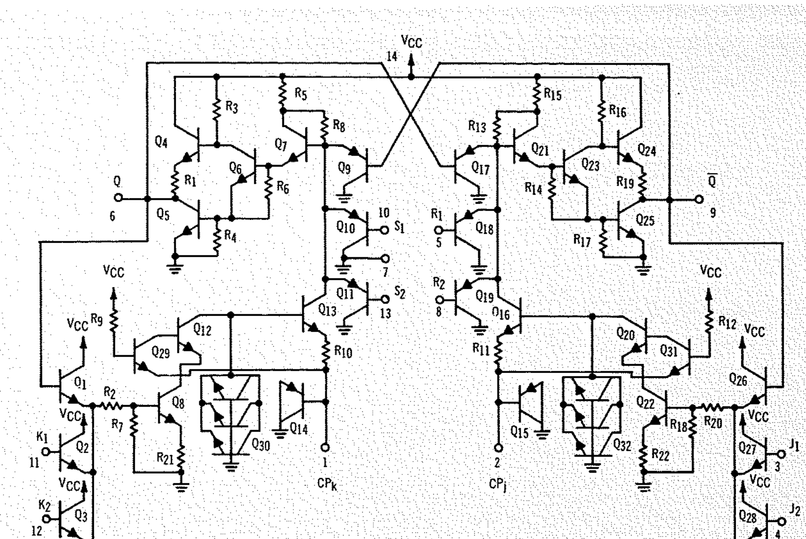


Fig. 1—Flip-flop schematic diagram.

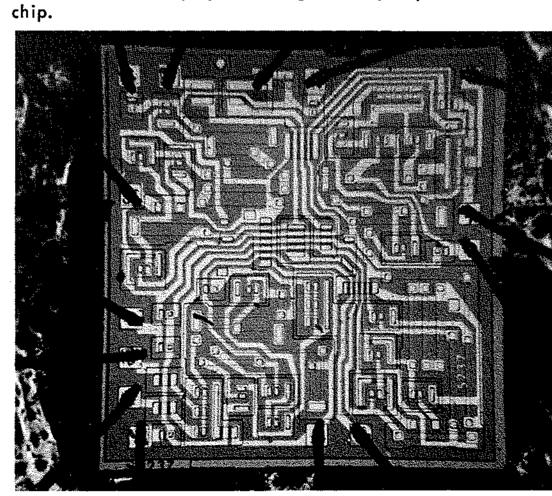
R. C. HEUNER received his BSEE from the City College of New York in 1959. He is now taking graduate courses toward an MSEE at the Newark College of Engineering. Mr. Heuner joined the RCA Defense Microelectronics activity of DEP in 1964. He was responsible for the circuit design, analysis and evaluation of a low-power family of monolithic logic circuits. In 1965 Mr. Heuner became Group Leader of the Defense Microelectronics Integrated Digital Circuits and Subsystems Group. In this capacity, he is responsible for the circuit design and development of new monolithic digital logic families intended for use by the DEP divisions of RCA. This responsibility has included the design and development of monolithic circuit, logic and layout techniques as applied to integrated digital arrays. Prior to coming to RCA, Mr. Heuner worked in the Digital Systems and Switching and Terminal Equipment groups at International Telephone and Telegraph Laboratories, Nutley, New Jersey. At I.T.T., Mr. Heuner initially worked on the logic design, debugging and system testing of an electronic digital switching center. He later worked in the circuits group on the design of discrete saturating and non-saturating type logic circuits, memory driver circuits, drum and tape read circuitry and other digital circuits. Mr. Heuner is a member of the IEEE.

Direct-current setting of the flip flop is accomplished by the application of a low-level input at either terminal 10 or 13. The DC resetting is accomplished when a low-level input is applied to either terminal 5 or 8. A high-level input at a DC set or reset input terminal will have no effect on the operation of the flip-flop circuit.

Trigger setting or resetting is accomplished by a change in the clock-line input from a high state to a low state. The flip-flop circuit will be triggered at the negative-going edge of the clock line pulse only when steering conditions are correct. The flip-flop is normally set up for split-clock operation. For single-clock operation, terminals 1 and 2 must be tied together externally. Internally connected self-steering controls provide the proper toggling action for single- or split-clock inputs. Connections for two additional external steering controls are available on each triggering circuit. Any steering control input at a high level on the *1* side triggering circuit (at terminals 11 or 12) will inhibit the triggering of the flip-flop to the *1* state. Any steering control input at a high level on the *0* side triggering circuit (at terminals 3 or 4) will inhibit triggering the flip-flop to the *0* state. The corresponding triggering circuit will be enabled only when the two external steering inputs and the internal self-steering input are all low. The circuit is then triggered at the negative-going edge of the clock-line input waveform.

The positive clock level charges the proper trigger capacitor under control of steering circuit conditions. Note that the *1* and *0* side-trigger capacitors are comprised of the junction capacities of *Q30* and *Q32*, respectively. The negative going clock edge changes the flip-flop state in accordance with the trigger capacitor states. The negative going clock signal also prevents any further steering input effect on either trigger capacitor until the clock line returns to a high level. This sequencing of operation simulates master-slave type of flip-flop operation, thus eliminating the possibility of race conditions. The clock trigger lines, on both the *1* and the *0* side, contain RC activating circuits designed to eliminate effects of transient noise pulses of amplitudes greater than the 1,500-mV noise-threshold level that occur at the clock inputs. When a clock line is raised to a high state and is held in this state, the flip-flop trigger capacitors will change to the triggering potential unless

Fig. 2—Photomicrograph of integrated flip-flop chip.



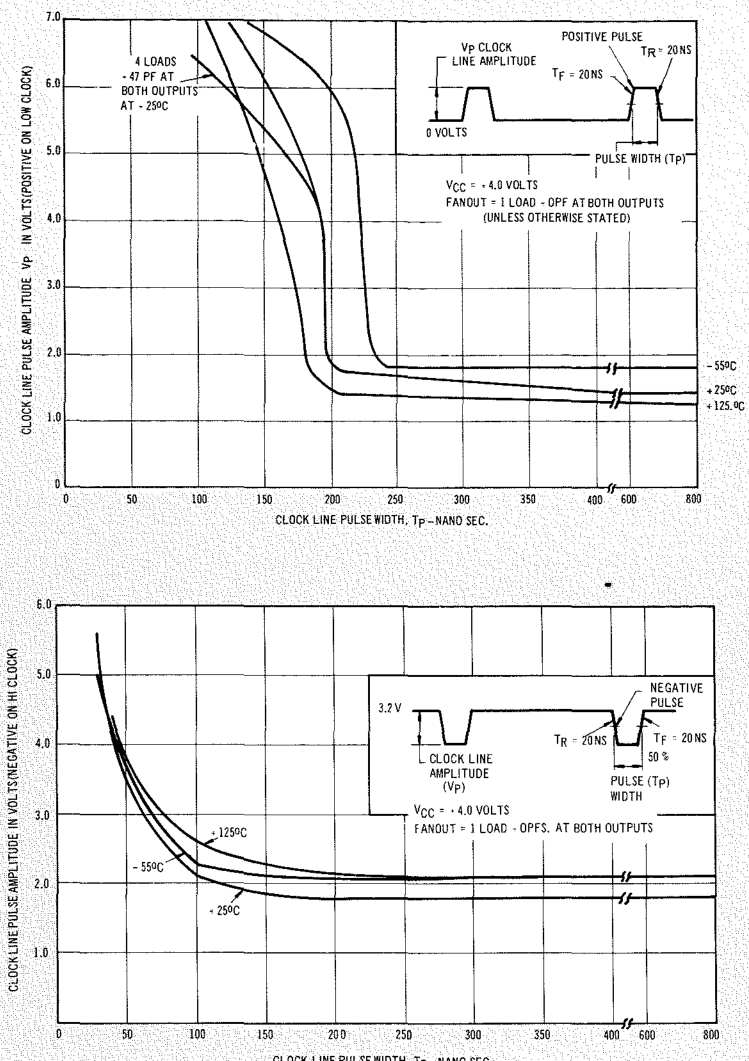


Fig. 3—Noise immunity characteristics of the clock line (typical trigger-circuit sensitivity). Amplitude of positive clock-line pulse required to trigger flip-flop as a function of clock-line pulse width.

prevented by inhibiting steering conditions. The charge path is from V_{cc} through R_9 and the base-collector junctions of *Q29* and *Q12* into *1* side trigger capacitor *Q30* and similarly through *R12*, *Q31*, *Q20*, and *Q32* on the *0* side.

It is impossible for any steering input to charge either trigger capacitor. Rather, these steering inputs will merely dictate whether the high-level clock input will be allowed to set the associated trigger capacitor. This type of steering control is achieved by the saturation or cutoff of transistor stages *Q8* or *Q22* which are effectively paralleled across junction trigger capacitors *Q30* and *Q32* respectively. For example, saturation of *Q8* sets the emitter of *Q12* to approximately one $V_{ce\text{-}sat}$ above ground. All charge current now diverts into the base-emitter of *Q12* and through *Q8* to ground. Trigger capacitor *Q30* consequently charges no higher than $V_{ce\text{-}sat}$ of *Q8*. During triggering, to ensure proper master-slave operation, the negative-going-clock trigger edge also pulls down the emitter of *Q29*, thus diverting all base-collector charge current and cutting off *Q12*. Trigger capacitor *Q30* is now completely isolated for all effects on the K side steering circuit regardless of the state of *Q8*. This condition exists until the clock line returns high thus allowing the trigger capacitor charge path to again function under control of the K side steering conditions. An identical situation exists on the *0* side trigger circuit. As a result, triggering of the flip-flop can never be falsely initiated or prevented by any steering line signal, regardless of clock-line levels. Similarly, when the steering inputs are inhibiting, the flip-flop can never be erroneously triggered from the clock lines. The internal steering connections assure that the existing state of the flip-flop will enable either the trigger circuit of the *1* side or the trigger circuit on the *0* side. Thus, the proper JK function is guaranteed.

Negative clamp diodes included on both clock input lines eliminate the possibility of double triggering of the circuit because of ringing at the negative-going edge of the clock input. Some presently existing integrated flip-flops have been noted to be quite susceptible to such clock line ringing.

OPERATING CHARACTERISTICS

Figs. 3 and 4 show the noise margins measured on typical flip-flop units. Fig. 3a illustrates clock-line noise immunity characteristics of the unit for positive noise pulses that occur on a low clock line. Fig. 3b illustrates clock-line immunity characteristics for negative noise pulses that are developed on a high clock line. Fig. 4 shows the DC set and reset input threshold level and the steering control input threshold level.

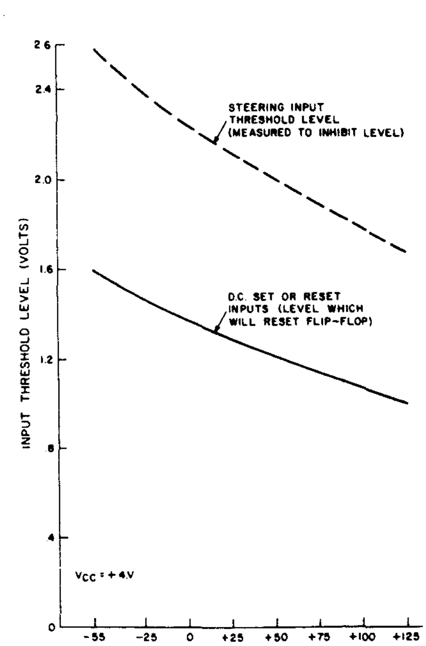
Fig. 5 points out the allowable range of variations in the clock waveform characteristics in typical flip-flop units. Note that proper binary operation exists for trigger edge transition times varying from 0 to 600 ns under all allowable temperature and clock pulse amplitude conditions.

Fig. 6 defines the required clock, steering, and DC set-reset input pulse widths and their respective timing relationships for proper flip-flop operation under all allowable operating conditions.

LOGIC EQUIVALENT

Fig. 7 shows the equivalent logic diagram and Boolean expressions which de-

Fig. 4—Typical noise-immunity threshold levels for the steering inputs and for the DC set and reset inputs of the flip-flop circuit.



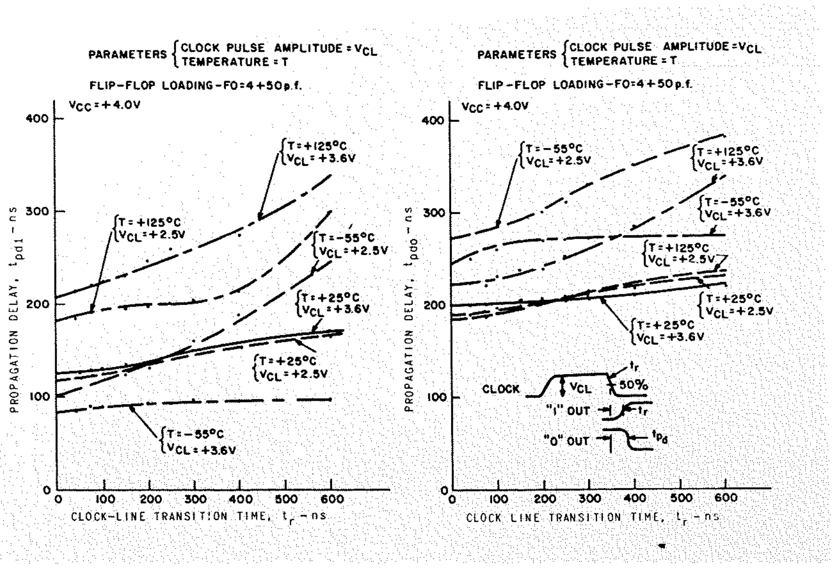


Fig. 5—Variation in the propagation delays of the flip-flop with changes in the characteristics of the clock-input waveform: (a) typical flip-flop propagation delay as a function of clock-pulse rise time; (b) typical flip-flop pair delay as a function of the clock-pulse rise time.

fine the logic functions and capabilities of the flip-flop circuit. This equivalency illustrates the master-slave JK mode of operation. In this case, the master storage unit is actually comprised of *1* and *0* side trigger capacitors.

MONOLITHIC INTEGRATED CIRCUIT COMPONENT CONSIDERATIONS

Many circuit design restrictions are imposed by the limitations of achievable integrated component characteristics (see Fig. 2).

Trigger Circuit Capacitors

The flip-flop configuration utilized reflects a definite effort to minimize the size and tolerance of the trigger capacitor elements required. Utilizing capacitors *Q30* and *Q32* in the base circuit of transistor stages *Q13* and *Q16* respectively, and *R10* and *R11* in the emitter circuits, yields trigger circuit energy time constants of β times the *RC* value. In addition, utilizing the trigger capacitors with one end grounded, augments the actual junction or metal-oxide capacitor with the back-biased n+ to substrate junction capacitance. Both the above circuit techniques allow the use of small (20-pF nominal), relatively loose tolerance (-0%, $+75\%$) capacitors. The range of capacitor values selected guarantees proper switching of the slowest flip-flop element while at the same time permitting operation at 1-MHz repetition rates.

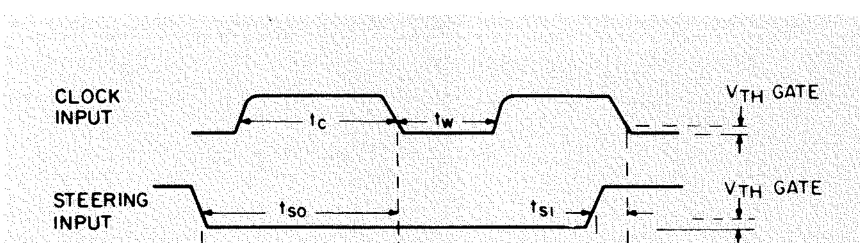
Integrated capacitors may be fabricated using either a metal-oxide or junction technique. The metal-oxide technique utilizes a n^+ diffusion region and the metalization pattern as the capacitor plates, with a controlled silicon dioxide insulating layer acting as the dielectric. The n+ region is fabricated during the emitter diffusion step. Such a capacitor, although not dependent on voltage biasing, requires close control of the silicon dioxide dielectric thickness for good capacitor tolerance. Presently, metal-oxide capacitor values of 0.07 pF/mil^2 can be reproduced. Thus, to fabricate a capacitor equal to 20 pF, a 286 mil^2 silicon chip area is required, excluding contact areas. To guarantee a minimum value of 20 pF, the parasitic junction capacitance from the n+ plate to the substrate was paralleled across the metal oxide capacitor. Measurements show this capacitance to be about 10 pF at a +3.6-volt bias, corresponding to the fully charged bias of the trigger circuit capacitor. This junction capacitor increases to about 15 pF as voltage bias decreases to a $V_{ce\text{-}sat}$ potential. Objections to the metal-oxide capacitor are the extra process steps required in fabricating the specially controlled silicon dioxide dielectric thickness as well as the loss of some units due to pin-hole shorts in the dielectric layer. Fig. 8 shows a photomicrograph of an early version of the flip-flop circuit using two metal-oxide capacitors along with corresponding paralleled n+ to substrate junction capacitors.

The junction technique utilizes paralleled sets of back biased silicon junctions, be they collector-substrate, base-collector or emitter-base junctions. Such capacitors are dependent on the voltage bias across them. No special process steps or pin-hole problems exist with this type capacitor. Care must, however, be given to maintain fabrication procedures and diffusion doping levels reasonably constant in order to attain good control over capacitance values.

Two 20-pF junction capacitors were fabricated for use in the final flip-flop circuit shown in Fig. 2. Design goals were for total junction capacitances of 24 pF at a +3.6-volt bias. This value increases to about 35 pF at a $V_{ce\text{-}sat}$ bias. To implement such a capacitor, three emitter-base junctions, three collector-base junctions and one collector-substrate junction were paralled together. Area of the collector-substrate junction was enlarged to yield the desired capacity value. Total area used for one 20 pF junction capacitor is 165 mil^2.

Resistors

Total resistance value utilized per flip-flop chip is 123 kilohms nominal. Inte-



Symbol	Definition	Recommended Minimum Pulse Width	Recommended Minimum Advance of Clock Trigger Edge
t_c	Trigger Circuit Set-Up Time	650 ns	
t_w	Minimum Width of Clock Pulse	350 ns	
t_{so}	Steering Allow Set-Up Time	850 ns	850 ns
t_{s1}	Steering Inhibit Set-Up Time	550 ns	550 ns
$t_{S\text{-}R}$	DC Set or Reset Pulse Width	500 ns	

Fig. 6—Required pulse widths and timing relationship of the inputs to the flip-flop circuit.

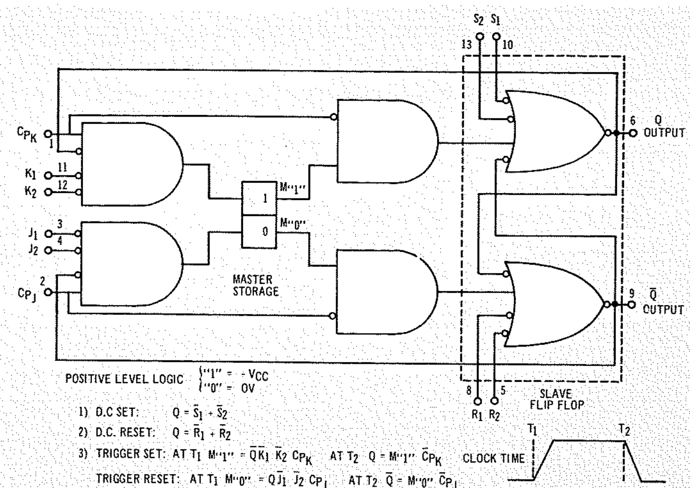


Fig. 7—Equivalent flip-flop diagram and Boolean expressions.

grated circuit resistor resistivities are typically 175 ohms per square, meaning 700 squares are required to implement the 123 kilohms. If layout rules are carefully observed, resistor absolute tolerances of ±20% and ratio tolerances of ±3% can be achieved. At the initiation of the flip-flop design, these layout rules called for minimum resistor widths of 1 mil and single-piece, straight-run resistor shapes. To practically achieve 123 kilohms resistance in the allowable chip size, resistor widths of 0.6 mil as well as "weaved" resistor patterns had to be utilized. All resistors requiring close ratio tolerances were made equal in width. A total resistance of 97 kilohms was fabricated using 0.6-mil widths. A total resistance of 26 kilohms was fabricated using 1-mil widths. Results to date on the flip-flop design show that resistor absolute and ratio tolerances are falling within the required worst-case values.

Transistors

The saturating output transistors used on the flip-flop circuit are designed for especially low $V_{ce\text{-}sat}$ values. An n+ collector contact ring is placed almost entirely around the base region, with collector metalization contact along this ring maximized. In addition, an n+ epitaxial collector layer is used beneath the entire n-area to further lower collector saturation resistance. On the flip flop units, $V_{ce\text{-}sat}$ values of 0.2 volt under maximum loading and temperature conditions at $V_{cc} = 4.0$ volts are typical.

The K steering circuit transistors, Q_1, Q_2, and Q_3 are fabricated as one unit having a common-collector region. An identical situation exists on J steering-circuit transistors, *Q26*, *Q27*, and *Q28*. In both cases chip area is reduced by such a scheme.

In the flip-flop steering circuit it is required that β's of the pnp substrate transistors of *Q29*, *Q12*, *Q13*, *Q31*, *Q20*, and *Q16* be low. In additon the inverse β of transistors *Q29*, *Q31*, *Q13*, and *Q16* must be low. Both of these characteristics are obtained by using a gold doping diffusion fabrication process. Pnp and inverse β's of 0.2 or less have been measured on the flip-flop of Fig. 2.

Metalization Pattern

Metalization widths of 1 mil or more were utilized wherever possible. In certain areas, however, metal widths of between 0.5 and 1 mil were necessitated by the density of signal wiring. Such runs were made 1 mil or more wherever possible. Metal run separations of 1 mil or more were used in all cases. Metal run crossovers were accomplished by the use of tunnels. Note that the low-power dissipation level of the flip-flop eliminates power distribution problems associated with either narrow metal patterns or tunnels.

MONOLITHIC INTEGRATED CIRCUIT DESIGN CONSIDERATIONS

Trigger Circuit Energy

To guarantee proper triggerable flip-flop operation under all allowable operating conditions, the steering circuit must be capable of holding the on-gate side *off* until the flip-flop unit latches itself. Specifically, at the clock trigger edge and assuming the flip-flop is to be switched from a *0* to a *1* state, all *1*-gate *on* current drive from V_{cc} through *R5* and *R8* must be diverted through *Q13*, and *R10* into the low CP_k line. This condition must be maintained until the *0* gate turns on and changes the base potential of *Q9*, thus diverting the *1*-gate on-current drive and causing the flip-flop to latch. The general expression derived for the steering circuit trigger capacitor discharge time is:

$$i_D\ (t) = (\mathrm{I}_E - I_C)\ exp\ (-t/\tau)$$

$$\tau = R_{eq}\, C = \left[\frac{R_{10}}{1-\alpha(t)}\right] C\,(t) \tag{1}$$

where: $i_D\ (t)$ = trigger capacitor discharge current; I_E = initial *Q13* emitter current through *R10*; I_C = initial *Q13* collector current from *R5* and *R8*; R_{eq} = equivalent resistance reflected into *Q13* base circuit; $\alpha(t)$ = forced $i_C(t)/i_e(t)$ of *Q13*; and $c(t)$ = trigger capacitor *Q30*. Note that τ contains two expressions, namely $\alpha(t)$ and $c(t)$ which are functions of time.

As capacitor *Q30* discharges, the emitter current, $i_e(t)$ through *Q13* decreases. The $i_c(t)$ will decrease slightly in value as a function of time as the overdrive factor of *Q13* decreases. The overall resulting effect is an increase in the value of $\alpha(t)$ as a function of time. Effectively, the increase in $\alpha(t)$ corresponds to an increase in the forced β of *Q13* with time. Qualitatively what is taking place is a reduction in capacitor discharge current with time and consequently an increasing discharge time constant. Subsequently the trigger capacitor discharges at progressively slower rates toward a zero-current level.

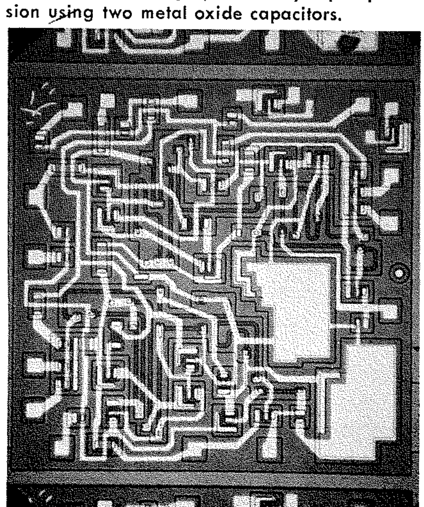
Since capacitor *Q30* is a junction capacitance, its value will increase with time during the discharge period. Thus, the $C(t)$ expression in Eq. 1. Note that the increase in the value of $C(t)$ acts to increase the discharge time constant of the trigger circuit.

To simplify solution of Eq. 1, two conservative assumptions were made. First, $C(t)$ was assumed constant at its initial fully charged or minimum value with time. Second, the $\alpha(t)$ value was broken into four time segments and assumed constant at its initial value in that time segment. Eq. 1 was then solved for time t over the four separate time segments using Eq. 2:

$$t = R_{eq}\, C\ \ln\left[\frac{I_{final} - I_{initial}}{I_{final} - I_{critical}}\right] \tag{2}$$

The critical value of current used for one $\alpha(t)$ section corresponded to the

Fig. 8—Photomicrograph of early flip-flop version using two metal oxide capacitors.



initial current value of the next $\alpha(t)$ section (initial current values calculated from initial $\alpha(t)$ values). The final current value is zero and thus can be dropped from Eq. 2.

Worst-case conditions exist at $-55°C$ temperatures. This is due mainly to the decrease in β of *Q13* at lower temperatures. Based on experimental and calculated results, the slowest flip-flop unit is found to have a maximum latch time of 450 ns at $-55°C$. The four time segments at which Eq. 2 was evaluated were for $\alpha_1 = 0.14$, $\alpha_2 = .714$, $\alpha_3 = 0.833$ and $\alpha_4 = 0.91$, corresponding to forced β values of 0.16, 2.5, 5, and 10 respectively. The final α value of section four was 0.938 corresponding to a β of 15, which is the minimum value assumed at $-55°C$. Using the above criteria, a trigger capacitor *Q30* of 20 pF minimum, along with a worst case *R10* value of 73% of the 2,000 ohms nominal, can maintain enough base drive through *Q13* to sink all *1*-side on-gate current for 566 ns, thus guaranteeing proper worst-case toggling action. As soon as the flip-flop toggling action has occurred, and when the clock line is returned to a high level, capacitor *Q30* will discharge rapidly through saturated stages *Q12* and *Q8*. Thus, no penalty is paid in the discharge cycle if the β RC discharge time constant is appreciably higher than the flip-flop latch time. An identical situation exists on the *0*-side trigger circuit.

Inverse-β And pnp Requirements

As mentioned in the preceeding section, it is necessary to maintain a low pnp substrate β at steering circuit transistors *Q29, Q12, Q13, Q31, Q20* and *Q16*. In addition, a low inverse β at transistors *Q29, Q13, Q31,* and *Q16* is required.

For example with the clock line at a high level, the flip flop in the *1* state and $V_{cc} = 4.0$ volts the following situation arises. The *Q29* is biased in an inverse direction, while both the pnp substrate transistors at *Q29* and *Q12* are forward-biased. Clock current now flows into the emitter of *Q29*. If the inverse β of *Q29* takes on a maximum value of 0.2, a 16-μA clock-line current drain can occur. This has a completely negligible effect on the high clock-line voltage potential. On the other hand, a maximum inverse β of 2 would cause a 156-μA high clock-line current drain to flow. A gate circuit driving three flip-flop clock-line loads may then be forced to pass 468μA in the high state. This in turn would drop the clock-line *1* level by about 187 mV, an allowable but notable effect.

A second situation arises in the charging time constant of the trigger capacitor circuit. In this case, assume a high clock level and a *0*-state flip-flop exist. This again biases *Q29* on in the inverse direction while also forward biasing both the pnp substrate transistors of *Q29* and *Q12*. Assuming an inverse β and pnp β of 0.2 maximum slows down the trigger capacitor charge time constant by 16.5%, a tolerable value. On the other hand, a pnp β of 1 maximum (inverse β assumed approximately equal to pnp β) would cause a 50% slowdown in trigger-capacitor charge-time constant.

SUMMARY OF APPLICATIONS

In summary, the following applications illustrate some of the logic flexibility of the flip-flop circuit:

1) Fig. 9a shows the logic diagram of a five stage shift register configuration. Feedback connections for either ring counter or unidistant decade register operation is permissible.
2) Fig. 9b shows the logic diagram of a three-stage ripple counter configuration; *nand*-gate assembly of clock gated counts 5 and 8 are also shown.
3) Fig. 9c shows the logic diagram of a three-stage parallel-carry counter configuration; *nand*-gate assembly of clock gated count 8 is also shown. Note that all counter gating is accomplished through use of the multiple J and K steering inputs. Six- and nine-stage parallel carry counters can be implemented using only three and six external *nand* gates, respectively.
4) Fig. 9d shows the logic diagram of a ripple-through decade counter. The counting sequence is also given. For this application the split clock capability is used to facilitate resetting the fourth counter flip-flop.

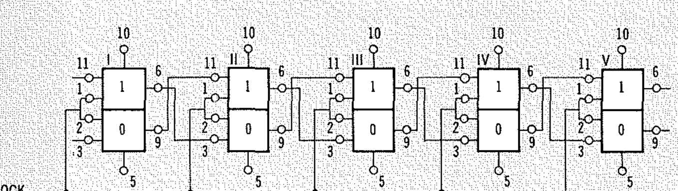


Fig. 9(a)—Logic diagram of a five-stage shift register configuration.

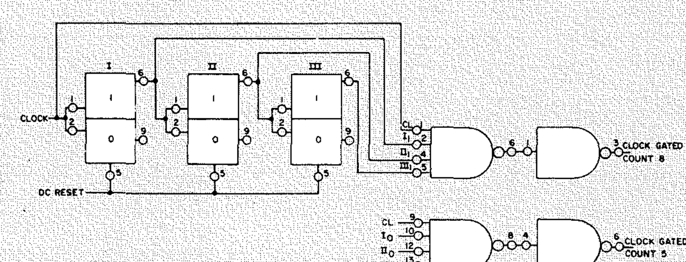


Fig. 9(b)—Logic diagram of a three-stage ripple counter and two assembled counts.

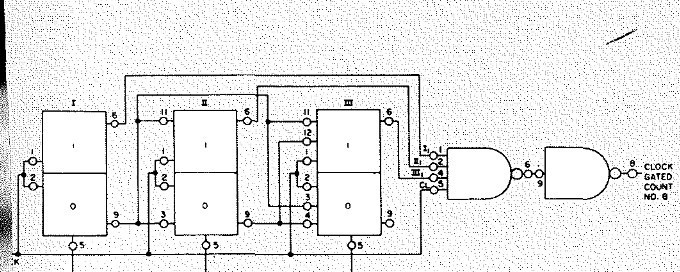


Fig. 9(c)—Logic diagram of a three-stage parallel carry counter and assembled count.

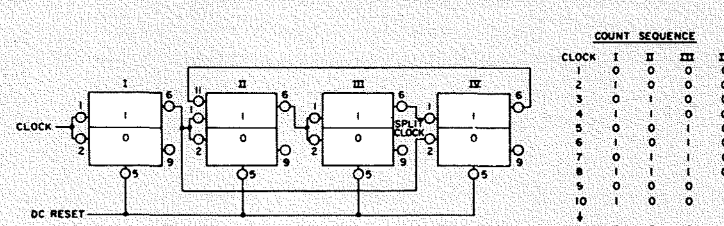


Fig. 9(d)—Logic diagram of a ripple-through decade counter.

ELECTRONIC ZOOM BY VARYING VIDICON RASTER SIZE

An imagining camera used in airborne or missile applications requires low weight components. When both wide-angle and narrow-angle zoom viewing is desired, the lens system usually becomes heavy and complicated. This paper presents a means of obtaining the desired zoom feature by electronically varying raster size on an imaging tube. An electronic zoom camera has been designed and built at RCA using a $1\frac{1}{2}$-inch vidicon and magnetic deflection at standard commercial frame rates. A beam-current servo loop and automatic target control allow zoom-ratio variation from 1:1 to 5:1 without tube adjustments.

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When variable fields of view are required for a TV camera system, the space and weight requirements of the optical lenses can become large compared to the electronics. If the field of view can be varied electronically (without optical "zoom" lenses) obvious advantages are gained in specialized TV camera applications (such as airborne or portable cameras where weight and size are critical), assuming that the image resolution and quality so obtained are adequate for the application involved.

ELECTRONIC ZOOM—ADVANTAGES AND DISADVANTAGES

The $1\frac{1}{2}$-inch-vidicon zoom camera described herein is more compact and lightweight than one with the customary optical zoom lens. It uses magnetic deflection at standard commercial frame rates. Control flexibility is inherent, since a beam current servo loop and an automatic target control allow zoom-ratio variation from 1:1 to 5:1 without tube adjustments. This gives many possibilities for feedback control of field of view. The response time of the zoom is fast in the forward direction, since no mechanical parts must be moved.

A fundamental limitation of the electronic zoom technique is that an extremely good image tube is required in order to obtain high resolution in the zoomed condition. However, this is partly offset by the fact that the resolution is usually much better at the center of the tube and optics than at the outer circumference. A relatively slow response time is obtained when reversing zoom ratio, since the unscanned photosurface area had been integrating large amounts of light which must be discharged by more than one frame of electron scan. In general, higher resolution may be obtained by optical zoom lenses, but at the obvious cost of weight and flexibility.

Raster burn because of underscanning may or may not be a limitation depending upon the particular camera application and upon the particular tube being used. For instance, in the tactical missile guidance application for which the zoom camera was constructed, raster burn is not important since zoom operation occurs only once; moreover, raster burn has not occurred in more than two years of operation with this camera. Slight raster burns have, however, been observed on a similar vidicon used in a similar application. These facts, coupled with the observation that tube manufacturers do not warrant their standard vidicon photoconductive surfaces against changes in target sensitivity resulting from underscanned operation, suggest that the designer should be aware of these pitfalls before undertaking a design.

This zoom technique has many applications where the size and weight advan-

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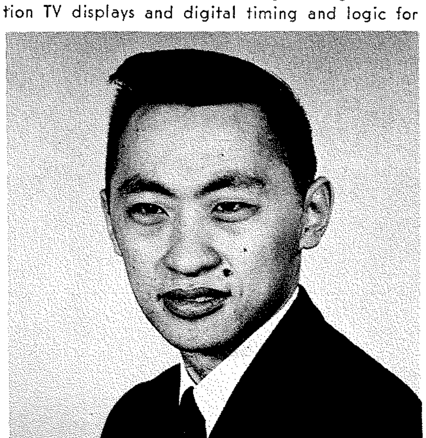
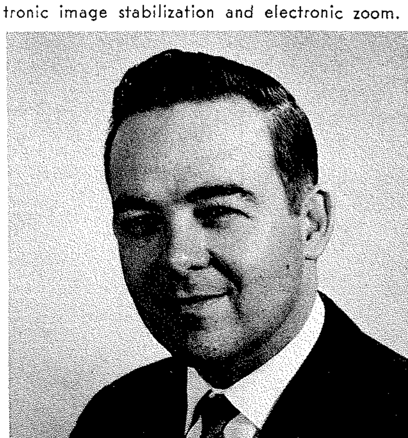
R. C. KEE received his BSEE from Worcester Polytechnic Institute in 1955 and his MSEE from Northeastern University in 1961. Since joining RCA in 1956, after a year at Bell Telephone Laboratories, Mr. Kee has worked on the design and development of magnetic modulator and transistor circuits for airborne conversion equipment, logic and digital circuits for weather balloon telemetry, and transistor circuits for video tape servo and infrared seeker systems. He has completed work on Video Image Processor circuits and logic for a satellite borne Optical Tracker. Recently, he directed the design of circuits and logic for a master timer synchronizer, and slow image motion processor components, for the AN/FSR-2 Optical Surveillance Subsystem. He was concerned with the study and design of lightweight, high efficiency DC-to-DC converters and DC-to-AC inverters for the LEM project, using semi-conductor and magnetic techniques. He supervised the design of a cooled infrared vidicon camera and the development of a deflection intensifier vidicon camera with electronic zoom and image deflection capability. At present he is directing a group in Electro-Optical Engineering, responsible for the design of high resolution TV displays and digital timing and logic for project Night Life. Mr. Kee is a member of IEEE, Eta Kappa Nu, Tau Beta Pi, and Sigma Xi.

DONALD DION received the BSEE from the University of Maine in 1951, and did graduate work at Brooklyn Polytechnic Institute. He became associated with RCA in 1951 as an inductive component engineer. On a research fellowship from 1956 to 1957 he did graduate work in the field of network synthesis. Returning to RCA in 1957, Mr. Dion did logic design on automatic test equipment and time division data link instrumentation. From 1958 to 1959 he worked on communications problems of a satellite interceptor system and on nose cone discrimination procedures including an examination of applicable radar polarization techniques. From 1960 to 1963 Mr. Dion worked in the radar area where work included performance evaluation of monopulse radar, digital radar signal processing, and correlation techniques. Since 1963, Mr. Dion has worked in the Electro-Optical group where work has included statistical analysis of the low contrast performance of image and photomultiplier tubes, circuit design on an infrared vidicon camera chain, and implementation of techniques for electronic image stabilization and electronic zoom.





tage may be used. Airborne or missile cameras may be used for surveillance in the unzoomed conditions and when any suspicious objects or targets appear, may be zoomed in for closer examination. Portable TV camera equipment may be made lighter and more compact where the resolution is sufficient. The technique may be extrapolated for use in other image tubes with raster scans.

TECHNIQUE DESCRIPTION

A diagram of the electronic zoom camera is shown in Fig. 1. A simple, fixed focal length lens is used to focus an image from the light box onto the photoconductive surface of the 8521 vidicon tube. Magnetic deflection coils are used to deflect the vidicon beam across the target surface to obtain a standard television raster of 525 lines at standard rates. The vidicon beam current is kept constant by use of a beam current servo circuit consisting of the vidicon, R_1, R_2, and DC amplifier A. The grid No. 1 voltage determines the amount of beam current flowing in the vidicon. By varying the potentiometer R_1, the output voltage of A which feeds G_1 may be varied, thus adjusting the beam current to a desired value. The beam current flows out of the cathode (K) through R_2 which generates a voltage across R_2 proportional to the current. This voltage across R_2 is sent to A as an error voltage to keep the current constant through R_2 and thus also constant out of the vidicon.

Normal magnetic deflection coils are used for sweeping the electron beam to produce the standard raster. A normal format of 3:4 aspect ratio is used and, in the unzoomed position, the rectangular format is just circumscribed by the vidicon target ring. By varying a single voltage potentiometer in the raster zoom control, the rectangular sweep raster on the face of the vidicon may be varied continuously by a factor of five in linear dimensions. This reduction in vidicon raster size essentially reduces the field of view as seen by the monitor. That is, if a 25° field of view is focused onto the vidicon face in the unzoomed condition, only 5° field of view is looked at by the reduced scanning raster. This, in essence, provides a zoom image on the monitor. There raster zoom control consists of the horizontal and vertical deflection coils. Each deflection circuit is continuously adjustable over a 5:1 sweep range and is voltage controlled from a common zoom control potentiometer. When zooming, the 3:4 aspect ratio is kept and the raster remains centered.

In order to keep good picture quality on the monitor while zooming, an automatic target loop was designed. The purpose of this feedback loop is to insure a relatively constant amplitude video signal output for the changing conditions within the vidicon during zoom. The video signal taken off the target (T) is amplified through the video preamp and amplifier to be used by the monitor. This amplified signal is also rectified and filtered in the integrating amplifier to obtain a DC voltage proportional to the average video out. This DC voltage is then fed back to the target lead through resistor R_3 to keep the video a constant average value. For ease of loop evaluation, an auto-manual switch is provided to open up the target loop and adjust the target voltage with potentiometer R_4. Thus, with a beam servo loop to keep the beam current constant and an automatic target loop to keep the output video constant, the electrical zoom control may be used without fear of having to readjust other controls at each zoom setting.

ELECTRONIC CAMERA CHARACTERISTICS

Zoom ratios of 5:1 have been achieved in the electronic camera. Higher zoom ratios are possible; however, the maximum usable zoom ratio will be limited by the loss in picture resolution as the camera is zoomed. The loss of resolution comes about primarily because of the reduced vidicon raster size as the picture is zoomed. This reduction in raster size reduces both the effective resolution of the vidicon itself and the effective resolution of the lens. The loss of vidicon resolution results because the size of a resolution element, referenced to the scanned portion of photoconductive surface of the vidicon, becomes smaller as the raster is zoomed, while the size of the read-beam spot stays relatively large. This large read-beam spot scanning over a small resolution element results in a reduction in picture contrast and, therefore, in resolution. The loss in resolution of the lens is a result of the smaller lens angle subtended by the zoomed image.

The effective resolution of the vidicon by itself is inversely proportioned to the zoom ratio. For an unzoomed vidicon raster, the size of the scanning spot approximates the size of a resolution element. The size of an effective resolution element referenced to the scanned portion of the vidicon photosensitive surface is, however, inversely proportional to the raster size. As a result, the effective resolution of the vidicon is inversely pro-

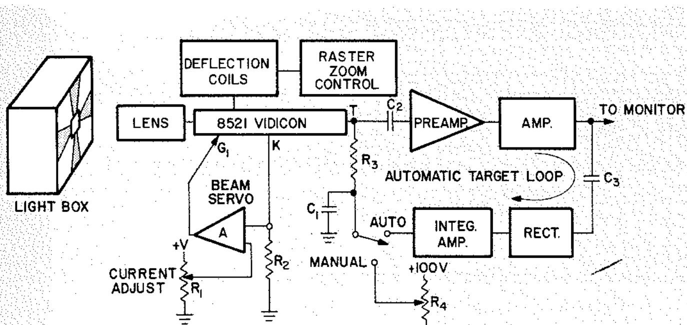
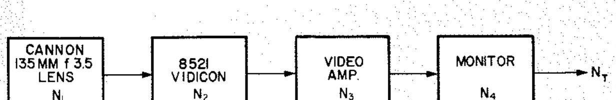


Fig. 1—Electronic zoom camera.



$$\frac{1}{N_T^2} = \frac{1}{N_1^2} + \frac{1}{N_2^2} + \frac{1}{N_3^2} + \frac{1}{N_4^2}$$

N_T = TOTAL SYSTEM LIMITING RESOLUTION IN TV LINES

N_1 = LIMITING RESOLUTION OF LENS IN TV LINES

N_2 = LIMITING RESOLUTION OF 8521 VIDICON IN TV LINES

N_3 = LIMITING RESOLUTION OF VIDEO AMPLIFIERS IN TV LINES

N_4 = LIMITING RESOLUTION OF TV MONITOR IN TV LINES

Fig. 2—System resolution.

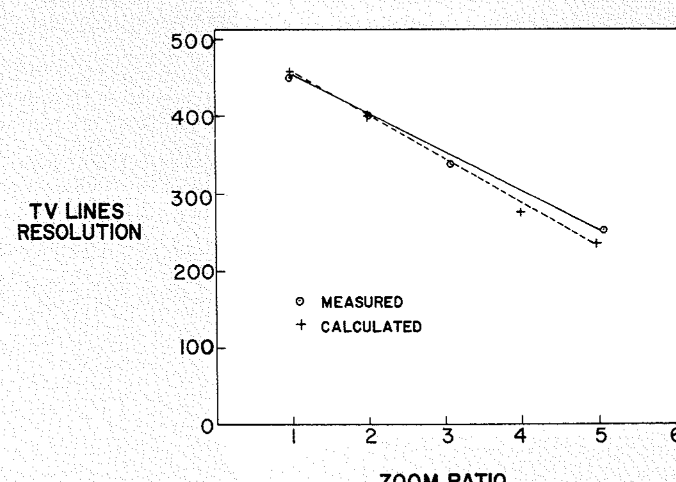


Fig. 3—Zoom camera resolution vs. zoom ratio.

portional to the raster size or to, what is the same, the zoom ratio.

The effective resolution of the lens also is inversely proportional to the zoom ratio. This is true because the resolution of the lens is determined by the lens angle subtended by the image projected on the vidicon photosensitive surface. Since this angle is approximately inversely proportional to the zoom ratio, so also is the lens resolution. The loss in overall camera resolution occasioned by zooming can be quantified by considering the vidicon and lens as being low pass filters in a cascade of low-pass filters. If each of the filters in such a cascade (see Fig. 2) have high-frequency rolloff rates approaching Gaussian—that is, $R(f) = \exp[-f^2/2\sigma^2]$—then it can be shown that the limiting resolution (2% response) is given by:

$$\frac{1}{N_T^2} = \frac{1}{N_1^2} + \frac{1}{N_2^2} + \frac{1}{N_3^2} + \frac{1}{N_4^2} \tag{1}$$

where: N_T = total system limiting resolution in TV lines, N_1 = limiting resolution of lens in TV lines, N_2 = limiting resolution of (8521) vidicon, N_3 = limiting resolution of video amplifier, and N_4 = limiting resolution of TV monitor.

Representative numbers for the limiting resolutions for the camera constructed are:

$N_1 = 2700$ TV lines (50 line pairs/mm)
$N_2 = 1500$ TV lines (8521, 1 ½-inch vidicon)
$N_3 = 640$ TV lines
$N_4 = 800$ TV lines

These give an overall (unzoomed) resolution, from Eq. 1, of:

$$N_T = 456 \text{ TV lines} \tag{2}$$

If the camera is zoomed 5:1, the new limiting resolutions are:

$$N_1 = \frac{2700}{5} = 540 \text{ TV lines}$$
$$N_2 = \frac{1500}{5} = 300 \text{ TV lines}$$
$$N_3 = 640 \text{ TV lines}$$
$$N_4 = 800 \text{ TV lines}$$

This gives an overall zoomed resolution of:

$$N_{T\,(\text{zoomed})} = 232 \text{ TV lines} \tag{3}$$

The measured resolutions and those calculated are compared in Fig. 3 showing that the resolution at any zoom ratio can be reasonably well predicted from Eqs. 1 and 2.

The target current as a function of zoom ratio is shown in Fig. 4. For this curve the target voltage was kept constant while the light level was varied so as to maintain a constant amplitude video output. The curve shows that as the camera is zoomed, more light is required to maintain the video output constant. More specifically, the curve shows that the amount of light required to maintain the output video amplitude constant is approximately proportional to the area scanned. This proportionality exists because the target current corresponding to a particular light level (or the target current difference corresponding to two different light levels) is proportional to the scanning speed or to the area scanned.

Fig. 4 shows that as the raster is zoomed the dark current drops. One would expect that the dark current would drop in proportion to the raster area scanned, since it is primarily from the scanned raster area that the dark current originates. The dark current curve of Fig. 4 shows a dropoff in dark current that is somewhat less than would be expected from area considerations alone. Fig. 5 shows the signal output voltage as a function of zoom ratio as the target voltage is raised to maintain the dark current constant. Raising target voltage as the raster is zoomed tends to increase the output video amplitude and partially counteracts the inherent dropoff in video amplitude occasioned by zooming. Comparison of Figs. 4 and 5 reveals that raising the target voltage when zooming (constant dark current) results in a gain in video amplitude of approximately five over that with constant target voltage.

The vidicon target voltage as a function of zoom ratio is shown in Fig. 6. The dotted curve was obtained by manually adjusting the target voltage to obtain an optimum picture at each zoom setting. The solid curve is the target voltage obtained at each zoom setting when the automatic target control servo loop is closed. Fig. 6 shows that the automatic target control servo maintains an image that is close to optimum during zoom.

CONCLUSIONS

A technique has been described for obtaining image zooming electronically which may be useful in special applications. Measurements and quantitative data have been discussed about the laboratory findings. Qualitative pictures, to date, indicate that the technique is definitely feasible for particular applications, with savings in weight and size.

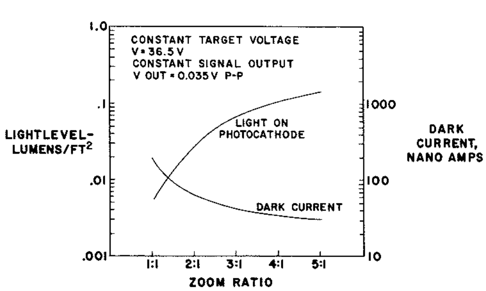


Fig. 4—Sensitive vs. zoom.

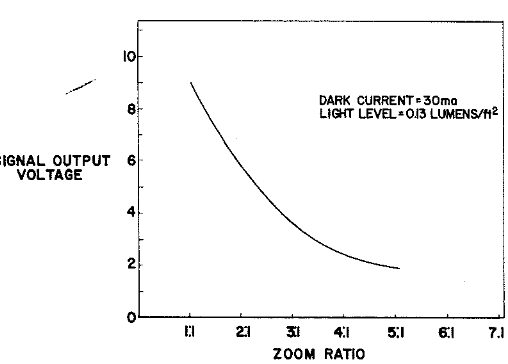


Fig. 5—Signal output voltage vs. zoom ratio.

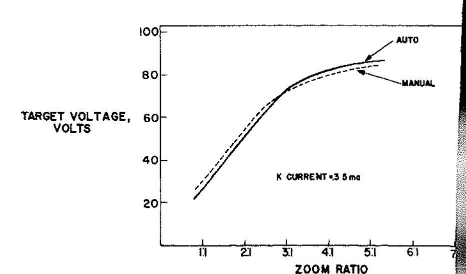
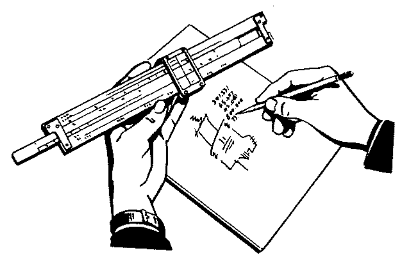


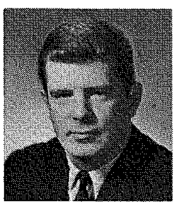
Fig. 6—Electronic zoom camera—target voltage vs. zoom ratio.

Engineering and Research NOTES

BRIEF TECHNICAL PAPERS OF CURRENT INTEREST



Stable Field-Effect Transistor Amplifier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Final manuscript received October 19, 1966

A linear integrated MOS amplifier has been developed which offers considerable promise for both analog and digital applications. Details of amplifier design which features the use of an MOS transistor for DC feedback are presented. The degenerative feedback so obtained stabilizes the amplifier's operating point against variations in supply voltage and possible drifts in device threshold (pinch off) voltage. This bias stabilization technique can also be employed in TFT integrated circuits. The amplifier in its most useful form is shown in Fig. 1.

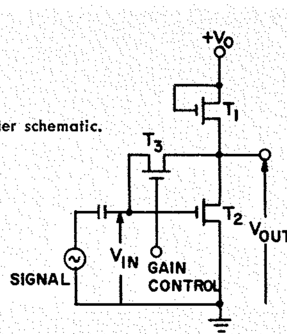
The basic amplifier consists of transistors T_1 and T_2, with T_1 acting as a load impedance for the input transistor, T_2. As the MOS (and the TFT) is a square-law device, the small-signal transconductance of T_2 varies linearly with bias voltage, which gives rise to distortion when a resistor is used for a load as in a conventional amplifier circuit. In this circuit, however, the nonlinear MOS load compensates for this nonlinearity in transconductance of T_2 in such a way as to maintain a constant small-signal voltage gain.

Stability is achieved by transistor T_3, which biases the amplifier so as to maintain the DC condition $V_{out} = V_{in}$. Drift in the characteristics of either T_1 or T_2 results in an automatic adjustment of the bias voltage so that the operating point of the amplifier is kept in its high-gain linear region. The gain of the amplifier is a function of the resistance of T_3. By varying the gate voltage of T_3, the gain of the amplifier can be controlled. Several amplifier stages can be cascaded by direct coupling with a feedback loop around the entire amplifier.

In the experimental amplifiers, it has been found convenient to make T_1 and T_2 identical except for channel width. The stage gain is then the square root of the ratio of the channel widths. Because the circuit requires no passive components, very little substrate area is used. Since all transistors can be of the same type, the number of processing steps is minimized.

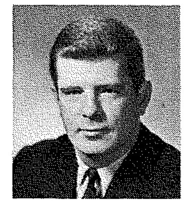
Measurements of distortion of experimental amplifiers have shown harmonic distortion to be less than 0.1% at 1 volt peak-to-peak output, and less than 1% at 7 volts peak-to-peak output. Because of its good linearity and voltage-controlled gain characteristics, the circuit might find application in I-F and AGC circuits in AM receivers, in computer sense amplifiers, and possibly in video amplifiers.

Fig 1—Amplifier schematic.



Complementary-Symmetry Silicon-on-Sapphire MOS Scratch-Pad Memory Cell


J. F. Allison


J. R. Burns


F. P. Heiman

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RCA Laboratories, Princeton, N.J.

Final manuscript received November 17, 1966

Most large high-speed computers have multiple means of data storage. Magnetic tape, discs, or cards serve as auxiliary mass storage, while magnetic cores are generally used for the main high-speed random-access memory. Increased system performance can be obtained through the use of an additional memory—the so-called *scratch-pad memory*. The main memory usually has a capacity of 4,000 to 128,000 words with access times of about 1 or 2 microseconds. The scratch-pad memory is smaller (64 to 512 words) but is considerably faster than the main memory. It is used to perform highly repetitive or closed-loop iterative calculations where its high speed of operation can result in a significant improvement in computer efficiency.

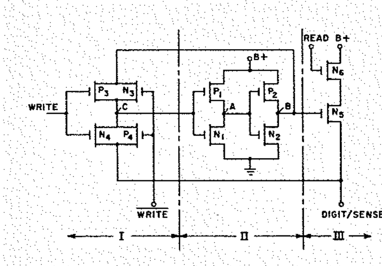
The problem in designing scratch-pad memories is to find a sufficiently fast system that can be fabricated at an acceptable cost. An integrated-circuit MOS transistor memory cell that appears quite promising for use in a scratch-pad memory has been conceived, constructed, and successfully tested at RCA Laboratories. The cell consists of six n-type and four p-type transistors (see Fig. 1) and requires an area of only 19 by 26 mils as shown in Fig. 2. Storage of a bit of information is achieved by setting the state of a 4-transistor complementary-symmetry flip-flop. The remaining six transistors in the cell are used for selection, write-in, and read-out. Because complementary symmetry is employed, the quiescent current drawn by the cell is only a few microamperes. Nondestructive read-out is accomplished in less than 2 nanoseconds and the write-in time is less than 15 nanoseconds—*considerably faster than magnetic cores.*

The excellent performance of the MOS memory cell was made possible by the development of a technique for growing single-crystal silicon layers on a sapphire substrate. If similar circuits are fabricated on bulk silicon, it is necessary to include back-biased diodes for isolation, and parasitic effects are troublesome. These problems have been eliminated by the use of the insulating sapphire substrate.

To fabricate these memory cells at an acceptable cost, the number of processing steps must be kept to a minimum. An important factor in reducing the number of processing steps is the use of a new technique wherein both the n-type and the p-type transistors are made from the same p-type silicon layer. The n-type transistors are obtained in the usual manner by creating an n-type inversion layer on the surface of the silicon, while the p-type transistors are deep-depletion transistors[1] that do not require an inversion layer. Another important factor is the use of low-capacitance silicon-film conduits for conductor crossovers. These conduits are fabricated at the same time as the transistors and are insulated from the metallic conductors by the oxide layer, so that only a single metallization step is required.

Acknowledgment: The research reported in this paper was sponsored by the Air Force Cambridge Research Laboratories, Office of Aerospace Research, under Contract No. AF19(628)-4208, and RCA Laboratories, Princeton, New Jersey. D. Cummings, A. O'Toole, J. Jaklik and R. Adams fabricated the memory cells.

Fig. 2—Photo-micrograph of silicon on sapphire memory cell ready for mounting and bonding. Bonding lands are 2 by 2 mils.



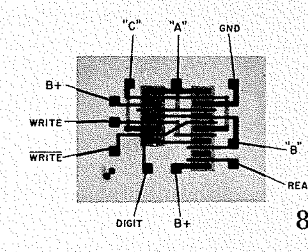


Fig. 1—Complementary-symmetry MOS scratch-pad memory cell circuit diagram.

Short-Circuit Protection for Regulated Power Supplies With Automatic Reset

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Final manuscript received July 7, 1966

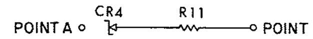


Fig. 2—Automatic start circuits No. 1. The minimum zener voltage of the diode must be greater than the maximum voltage that can exist between the collector and emitter of Q3 under all normal operating conditions.

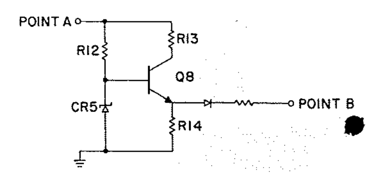


Fig. 3—Automatic start circuit No. 2. (Points A and B are referenced to Fig. 1).

This circuit prevents damage to a voltage regulator when a short circuit occurs across its output terminals, by supplying a short-circuit current that is much less than the regulator's full load current capability. The regulator automatically returns to normal operation the short circuit is removed.

Most power-supply regulator-protect circuits are of two general categories: 1) *current limiting* or 2) *voltage "crow bar"*. The current-limiting type switches to constant-current operation at some predetermined value of load current demand which is necessarily greater than the full-load current. The "crow-bar" type switches completely off under an overload condition and usually requires a manual reset switch or complicated reset circuitry for return to normal operation.

The protect-circuitry described here requires only two low-power transistors for both the protect and the automatic-reset functions, in addition to the normal regulator circuitry. The current output under short-circuit conditions is much less than full load current. For example, A 12-V 2A regulator would allow only 300mA of short-circuit current. The short circuit current is not supplied by the series pass transistor, but by a special *start circuit* that is gated into the output circuit only during a short circuited condition. During this shorted time, the series-pass transistor is completely cut off, which also reduces the power dissipation requirements (as opposed to the current-limiting type of operation) because it never has to carry more than full-load current. The short-circuit current from the start-circuit is just sufficient to restart the regulator when the shorted condition is removed from the output terminals.

Detailed Circuit Description: (Fig. 1): A normal transistor regulator circuit comprises *Q3, Q4, Q5, Q6, Q7* together with *CR2, CR3, R3, R4, R5, R6, R7, R8, R9,* and *R10.* The *CR1, Q2,* and *R1* comprise a constant current generator to furnish base current drive to the beta multiplier (*Q3* and *Q4*). The *Q5* is the normal DC error voltage amplifier of a regulator circuit controlled by the differential amplifier (*Q6* and *Q7*). The *Q1* is in series with *CR*1 of the constant current generator and controls the current drawn thru the zener diode *CR1*. When a short circuit is placed across the regulated voltage output terminals, Point *C* (collector of *Q7*) drops immediately to zero volts, cutting off *Q1* which in turn cuts off *CR1* and *Q2*. Therefore, the constant current generator is cut off, and in turn the beta multiplier *Q3* and *Q4*. The regulator will stay in this cut-off condition until a sufficient voltage at Point *C* starts *Q1* conducting and restarts the constant current generator.

Start Circuit No. 1 (Fig. 2): Under short circuited conditions, the voltage between Points *A* and *B* will be greater than the zener voltage of *CR4* and will, therefore, conduct the current being limited by *R11*. When the short circuit is removed, the current supplied through *CR4* and *R11* is enough to raise the voltage at Point *C* to restart the constant current generator.

Start Circuit No. 2: (Fig. 3): This circuit improves on circuit No. 1 when large excursions of the unregulated voltage at Point *A* might cause zener diode *CR4* to conduct under normal load conditions. In circuit No. 2, the zener diode *CR5* maintains the voltage at the emitter of *Q8* constant ($V_{ZCR5}-V_{BEQ8}$). Under normal operating conditions, Point *B* is at a higher voltage than the emitter voltage of *Q8*, thus back-biasing the gating diode *CR5* and presenting current flow through *R15* to Point *B*. Under short circuit conditions, Point *B* is at zero volts, thereby forward-biasing the gating diode *CR5* into conduction. When the short circuit is removed, the current supplied through *CR5* and *R15* is enough to raise the voltage at Point *C* to restart the constant current generator and in turn the regulator circuitry.

Unique Features of the Circuit:

1) The short circuit current which is allowed to flow is much less than the full load current capability of the regulator. This reduces the required power capability of the unregulated source.
2) The series-pass transistor is cut-off during the short circuit conditions and therefore has no power dissipation. This may allow the use of smaller heat sinks, less air flow requirements, or even a smaller power transistor.
3) When the short-circuited condition is cleared, the regulator automatically returns to normal operation (without the use of a manual reset switch) due to current supplied from the disclosed start circuitry.

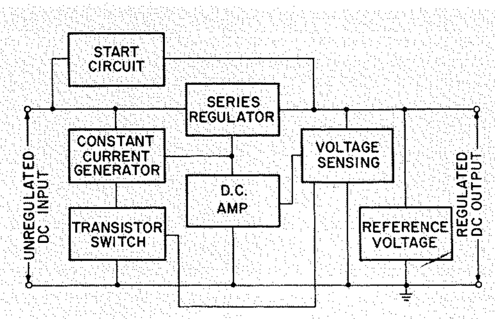
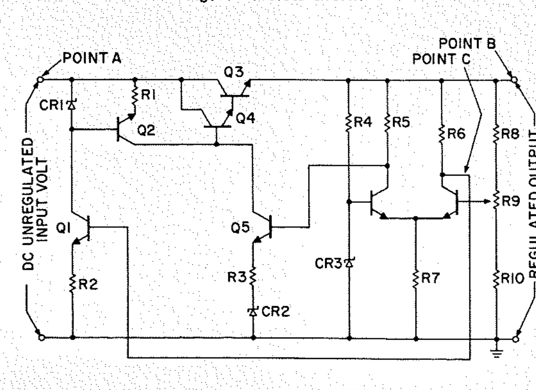
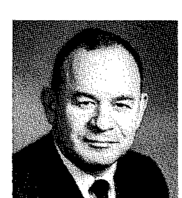


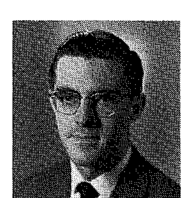
Fig. 1—Protect circuit.



Laminated Memory Construction with Discrete Ferrite Elements



I. Drukaroff



J. L. Vossen

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Final manuscript received Aug. 23, 1966

This *Note* describes a monolithic laminated-ferrite memory construction having discrete, physically-separated magnetic memory elements as shown in Fig. 1. It differs from a previously described[1,2] monolithic laminated-ferrite memory which is made by printing conductive patterns on green ferrite sheets, laminating the sheets, and firing the lamination to sinter the ferrite. In the latter approach, memory element is constituted by the ferrite surrounding each crossover of two printed conductors. Memory elements must be sufficiently spaced from each other so the magnetic flux in one memory element does not extend through the continuous magnetic material to the regions of adjacent memory elements.

In the approach shown in Fig. 1, three doctor-bladed, nonmag-

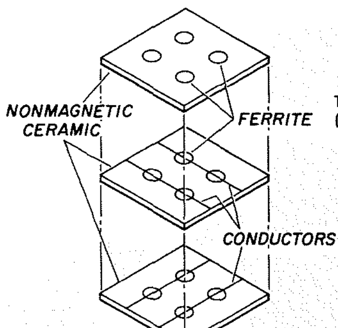

Fig. 1—Lamination arrangement. Top and bottom sheets of ceramic (not shown) may be added.

netic ceramic, green, leather-like sheets are provided with registered punched holes where memory elements are desired. The holes are filled with green ferrite in the form of leather-like discs or in the form of a paste which is allowed to solidify. Paste conductors are then printed over the ferrite spots on two surfaces of the nonmagnetic ceramic which become interior surfaces when the sheets are laminated together. The lamination may include top and bottom sheets (not shown) of nonmagnetic ceramic material. The lamination is fired to drive off binders and to sinter the ferrite and nonmagnetic ceramic into a unitary monolithic body having imbedded conductors. The ferrite surrounding each crossover of two conductors forms a discrete magnetic memory core separated from other similar magnetic cores by nonmagnetic ceramic material. Excellent magnetic isolation between ferrite elements, low power consumption, and speed of operation are features of this approach.

*Mr. Drukaroff is now with Defense Microelectronics, Somerville. Mr. Vossen is now with RCA Labs., Princeton.

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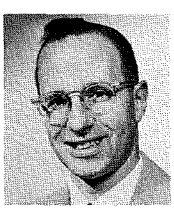
2. R. A. Shahbender, "Research on a Combined MOS-Circuit, Monolithic-Ferrite Integrated Memory," RCA ENGINEER, 11-3, Oct.-Nov. 1965.

An Approximate Equation Without Transcendentals for the Loaded-Q of a Waveguide Cavity Resonator



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Final manuscript received Aug. 15, 1966

Several equations relating the doubly-loaded Q of a waveguide cavity resonator to the normalized susceptance of the input-output inductive coupling obstacles (i.e., posts or irises) have appeared in the literature. These equations employ transcendental functions that are not convenient for many design calculations. This *Note* presents an approximate equation that does not employ transcendental functions and gives reasonably accurate results for most practical cases.

A single cavity resonator symmetrically loaded by matched waveguide terminations through inductive discontinuities can be represented by the equivalent circuit shown in Fig. 1. For a nominal half-wave resonator, resonance occurs when:

$$\tan\phi = \frac{2}{B} \qquad (1)$$

where B = normalized susceptance of inductive discontinuity, $\phi = 2\pi L/\lambda_{go}$ = cavity electrical length, L = cavity physical length, and λ_{go} = guide wavelength at resonance.

Assuming B varies linearly with λ (i.e., $B = K\lambda_g$ where K is a constant), the loaded Q of the resonant cavity has been derived by Reed:[1]

$$Q_L = \tfrac{1}{4}\left(\frac{\lambda_{go}}{\lambda_o}\right)^2\left[-B\sqrt{B^2+4}\tan^{-1}\left(\frac{2}{B}\right)+\frac{2B^2}{\sqrt{B^2+4}}\right] \qquad (2)$$

where Q_L = loaded Q of resonant cavity, and λ_o = free-space wavelength at resonance. This exact expression for Q_L takes into account the frequency sensitivity of B. (It should be noted that the center frequency is neither the geometric nor algebraic mean of the half-power frequencies.[2])

Approximate equations for loaded Q are available that neglect the frequency sensitivity of B. In Mumford's classic paper on waveguide filters,[3] the equation for loaded Q is:

$$Q_L = \tfrac{1}{4}\left(\frac{\lambda_{go}}{\lambda_o}\right)^2\left[-B\sqrt{B^2+4}\tan^{-1}\left(\frac{2}{B}\right)\right] \qquad (3)$$

This differs from Equation 2 by the absence of the factor $2B^2/\sqrt{B^2+4}$.

Riblet[4] employs an expression for Q_L as follows:

$$Q_L = \frac{(\pi-\phi)\cos\phi}{\sin^2\phi}. \qquad (4)$$

Letting $\phi = \tan^{-1} 2/B$, Equation 4 becomes identical to Equation 3.

Pritchard,[5] and Fano and Lawson[6] compute the loaded Q from the following equation:

$$Q_L = \left(\frac{1+B^2}{4}\right)\left(\frac{\lambda_{go}}{\lambda_o}\right)^2\tan^{-1}\left(\frac{2B}{B^2-1}\right). \qquad (5)$$

This equation has also appeared in Volume 9 of the *Radiation Laboratory Series*,[7] which presents the following simplified equation applicable to narrowband filters when $B \gg 10$:

$$Q_L = \frac{\pi B^2}{4}\left(\frac{\lambda_{go}}{\lambda_o}\right)^2 . \qquad (6)$$

It is hereby recommended that the following equation be used to relate doubly-loaded Q's to normalized susceptances:

$$Q_L = \frac{\pi}{4}(1+B^2)\left(\frac{\lambda_{go}}{\lambda_o}\right)^2 . \qquad (7)$$

Values of loaded Q for different normalized susceptances have been calculated using Equations 2, 3, 5, 6, and 7. These are in Table I, assuming $\lambda_{go}/\lambda_o^2 = 2.08$. It can be seen that the approximate Equation 7 provides the closest results when compared to exact Equation 2. Errors of less than 3.3% will be incurred for values of $B \geqq 2.0$. When $B \geqq 10.0$, this error is reduced to a maximum of 0.6%. Equation 7 works well because it entails compensating errors. By neglecting the factor $2B^2/\sqrt{B^2+4}$ in Equation 2, lower values of Q_L are obtained. By setting $-\tan^{-1}(2/B) = \pi$ for all values of B, higher values of Q_L are obtained.

TABLE I—Comparison of Calculated Q-Values

B	Q_L				
	Eq. 2	Eq. 3	Eq. 5	Eq. 6	Eq. 7
2	8.44	6.93	5.79	6.56	8.16
4	28.8	24.9	22.9	26.2	27.8
6	61.8	55.7	54.5	59.0	60.4
8	108	99.8	98.1	105	106.5
10	166	156.5	154	164	165
12	240	266	226	236	237

Equation 7 has not been derived from any analysis of the waveguide cavity resonator. It is recommended because it provides reasonably good accuracy and is simple to use. It should be noted that Equations 2, 3, 5, 6, and 7 are applicable to the singly-loaded waveguide cavity by replacing the factor ¼ by ½.

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2. S. B. Cohen, "Direct-Coupled Resonator Filters," Feb., 1957, *Proc. of the IRE*, Vol. 45, pp 187-196.
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4. H. Riblet, "Synthesis of Narrow Band Direct-Coupled Filters," Oct. 1952, *Proc. of the IRE*, Vol. 40, pp 1219-1223.
5. W. L. Prichard, "Quarter-Wave Coupled Waveguide Filters," Oct., 1947, *Journal Applied Physics*, Vol. 18, pp 862-872.
6. R. M. Fano & A. W. Lawson, "Microwave Filters Using Quarter-Wave Couplings," Nov., 1947, *Proc. of the IRE*, Vol. 35, pp 1318-1323.
7. G. L. Ragan, *Microwave Transmission Circuits*, McGraw Hill Book Co., Inc., New York, N.Y. pp 677-715, 1948.

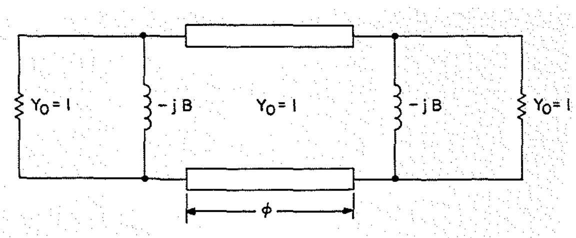

Fig. 1—Equivalent circuit of waveguide cavity resonator.

A Note On Resistor Temperature Gradients In High-Reliability Cordwood Packaging

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Final manuscript received November 8, 1966

Cordwood packaging is largely the result of the emphasis on smaller volume, lower weight electronic equipment. The decrease in volume, however, is not always accompanied by a decrease in power dissipation; quite the contrary, the trend is often toward an increase in power dissipation. The effect then is an increase in heat dissipation and corresponding thermal problems. At the same time, the reliability requirements may be increased because of the nature of the mission. Component temperatures consequently, must be more accurately predicted and controlled.

In cordwood packaging for a space (vacuum) environment, conduction and radiation are the only means of heat transfer. For the component temperatures normally encountered, radiation is negligible compared to conduction, so that the problem of heat transfer in such electronic equipment becomes one of conduction analysis.

One of the many variables in the conductive heat transfer path is the temperature drop in the component itself; by this is meant the temperature differential from the component surface hot spot to its lead. Very little information on this is available. The manufacturer usually gives only the maximum "allowable" temperature, and often this may not be valid for the particular packaging and cooling arrangement. The information desired is not readily amenable to calculation and can be practically obtained only by experiment. This *Note* reports the results of such experiments on resistors.

Fig. 1 shows a test set-up for measuring the hot-spot temperature drop (hot-spot temperature minus lead temperature) in resistors with cordwood type packaging. The resistors were mounted with one end flush (or practically so) with the circuit board, and were tested in a vacuum environment (10^{-6} mm-Hg or less).

This set-up was the same for all resistors tested. A thermocouple at the upper end of the resistor measured the hot spot temperature since the heat flow in the resistor was down toward the sink. Steady state temperatures were read at various power dissipations.

Summarized below are the tests made on typical types of resistors to determine the rate of temperature differential to heat dissipation (thermal resistance). Several resistors of each type were tested to obtain a typical range of thermal resistance (R_T) values from an upper limit to a lower limit. In two cases, resistors of the same type but from different manufacturers were used.

Resistor Type	Thermal Resistance (°F/Watt) Upper Limit	Lower Limit
RC07	596	403
RC20	384	241
RC32 (Conformal Coated)	162	150
RN55	370	280
RN55*	300	190
RNR57	317	250
RNR57*	260	250
RN70	128	106

* *Tests on resistors from another manufacturer.*

The results shown in Fig. 2 for a type RC20 resistor illustrate the considerable variations in a single manufacturer's resistors. From a design point of view, it would be dangerous to use any but the upper limit of approximately 385°F/watt for this type of resistor.

Fig. 3 illustrates the thermal resistances for two type RNR57 resistors, each made by a different manufacturer. Here again a variation is shown among the resistors made by the same manufacturer and between manufacturers. The upper limit is approximately 317°F/watt for one manufacturer and 260°F/watt for the second, with the data spread much larger at the higher value. Note that care must be taken, as a test with a single component may not be representative of that particular type and could lead to incorrect thermal analyses, especially where high dissipation is present. In a similar manner, the variation in thermal resistance for the RC07 resistor was between approximately 400°F/watt and 600°F/watt (see Fig. 4). This is obviously a considerable variation, and again the only resistance that can be safely used in a thermal analysis is the higher value.

Conclusions: The temperature drop per watt in a resistor with cordwood packaging is a variable which depends on the type of resistor and its manufacturer. Even for a given type of resistor, and specific manufacturer, this temperature drop varies from one resistor to the next. Consequently, tests made with a single resistor may not be representative of its type and could lead to incorrect conclusions from a temperature drop standpoint. The upper limit values should be used for conservative thermal design.

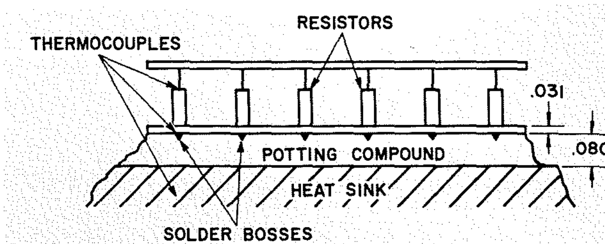


Fig. 1—Resistor module configuration.

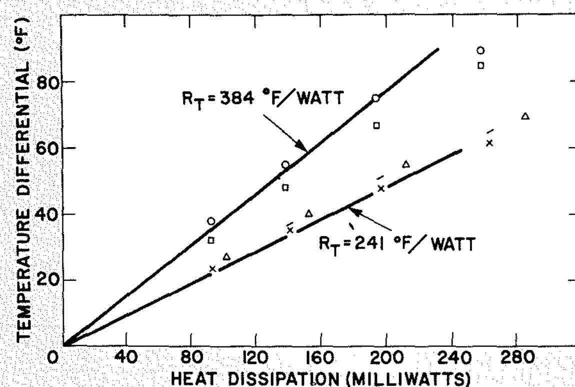


Fig. 2—Thermal resistance of RC20 resistors.

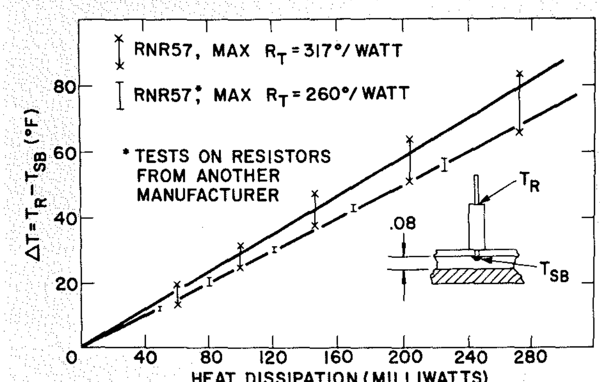


Fig. 3—Temperature rise along RNR57 resistor under high-vacuum conditions.

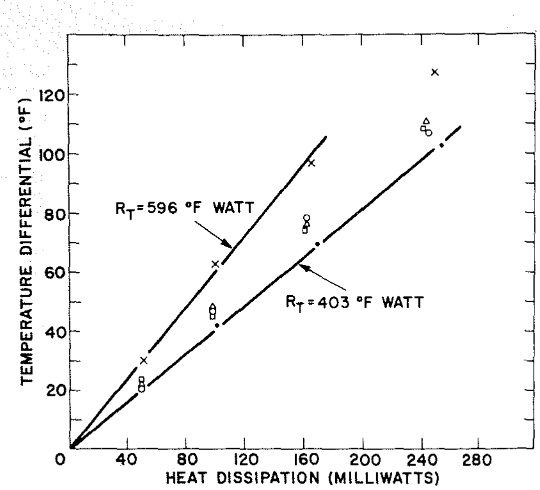
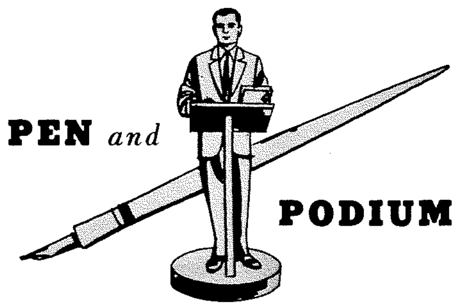


Fig. 4—Thermal resistance of RC07 resistor.

* *M. Mark is a consultant and Professor of Mechanical Engineering at Northeastern U., Boston, Mass.*



PEN and PODIUM

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Subject index categories are based upon the Thesaurus of Engineering Terms, Engineers Joint Council, N. Y., 1st Ed., May 1964.

SUBJECT INDEX

Titles of papers are permuted where necessary to bring significant keyword(s) to the left for easier scanning. Authors' division appears parenthetically after his name.

AMPLIFICATION

PARAMETRIC AMPLIFIERS, Superconducting—P. Bura (Labs, Pr) Nat'l Superconductivity Information Mtg., Brookhaven Nat'l Lab., Upton, N. Y., 11/9-11/66

ANTENNAS

CASSEGRAIN ANTENNA ILLUMINATION, A Technique for Frequency-Independent—C. E. Profera, L. H. Yorinks (MSR, Mrstn) 1966 Aerospace & Electronics Sys. Conv., Washington, D.C., 10/66; *Supplement to IEEE Trans. on Aerospace & Electronics Systems*

CASSEGRAINIAN ANTENNA, LEM Rendezvous Radar—J. B. Grabowski, W. E. Powell (MSR, Mrstn) 1966 IEEE Int'l Antenna & Propagation Symp., Palo Alto, Calif., 12/5-7/66; *Conf. Proc.*

MODERN ANTENNAS: Theory and Design of, Vol. 1—P. Foldes (RCA Ltd., Mont) Lectures for Postgraduate Students at McGill Univ., 12/66

SEARCH RADAR Antenna, An Improved LP-CP—O. M. Woodward (MSR, Mrstn) and R. W. Klopfenstein (Labs, Pr) 1966 IEEE Int'l Symp. on Antennas & Propagation, Palo Alto, Calif., 12/5-7/66; *Conf. Proc.*

CIRCUITS, INTEGRATED

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COMPLEMENTARY INTEGRATION TECHNOLOGY for Communication Systems—W. Y. Pan (CSD, NY) IEEE Chapter of Lincoln, Nebraska, 4/66

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SUPERCONDUCTIVITY (& cryoelectrics)

CRYOELECTRIC BRIDGE CELL, Characteristics of the Three-Wire—W. C. Stewart, L. S. Cosentino (Labs, Pr) 12th Ann. Conf. on Magnetism & Magnetic Matls., Washington, D.C., 11/15-18/66

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ELECTRON-MICROSCOPE IMAGES (TV-Displayed), On the Possible Use of Data Correlation Techniques for Extracting Phase-Contrast Information from—Dr. J. W. Coleman (BCD, Cam) Armed Forces Inst. of Pathology at Walter Reed Hosp., 12/1/66

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High Efficiency Light Modulation System—F. Sterzer (ECD, Pr) U.S. Pat. 3,302,028, January 31, 1967

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Method of Making Electron Gun Mount—T. J. Kelly (ECD, Hr) U.S. Pat. 3,298,083, January 17, 1967

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Electron Tubes and Methods of Operation Thereof for Energy Conversion—A. L. Eichenbaum (Labs, Pr) U.S. Pat. 3,274,404, September 20, 1966

Acoustic-Electromagnetic Device—H. S. Sommers (Labs, Pr) U.S. Pat. 3,274,406, September 20, 1966

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Microwave Phase Shifter System Providing Substantial Constant Phase Shift Over Broad Band—N. L. Korman (BCD, Cam) U.S. Pat. 3,275,952, September 27, 1966

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Method of Coating a Metal Surface with a Ferrite Composition—A. N. Schmitz (MSR, Mrstn) U.S. Pat. 3,284,236, November 8, 1966 (assigned to U.S. Gov't)

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Parallel Amplifier Circuit Having Load Equalization Means—B. J. Jones (CSD, Cambridge) U.S. Pat. 3,292,094, December 13, 1966

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Sound Signal Correction System—D. L. Richter (RecDiv, NY) U.S. Pat. 3,293,364, December 20, 1966

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Photocomposing System—H. E. Haynes (AppRes, Cam) U.S. Pat. 3,273,476, September 20, 1966

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Single Tube Vertical Deflection Circuit for a Television Receiver—R. N. Rhodes, J. B. Beck (HI, Indpls) U.S. Pat. 3,287,596, November 22, 1966

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Semiconductor Signal Translating Circuit—G. E. Theriault (HI, Pr) U.S. Pat. 3,290,613, December 6, 1966

Variable Saturable Reactor—M. W. Garlotte (HI, Indpls) U.S. Pat. 3,283,279, November, 1, 1966

Insulated Gate Field Effect Transistor Oscillator Circuits—L. A. Harwood (HI, Pr) U.S. Pat. 3,281,699, October 25, 1966

Portable Phonograph or Similar Article—B. A. Grae, R. S. Cox (RCA Sales Corp., Indpls) U.S. Pat. 205,795 (Design Patent), September 20, 1966

Pulse Forming Circuit for Horizontal Deflection Output Transistor—H. C. Goodrich (H.I., Indpls) U.S. Pat. 3,302,033, January 31, 1967

Magnet Means for Correction of Blue Beam Lateral Deflection for Color Television Receiver Tubes—E. Lemke (HI, Indpls) U.S. Pat. 3,302,049, January 31, 1967

Adjustable Deflection Yoke Mounting for Color Picture Cathode Ray Tubes—M. J. Obert, J. M. Ammerman (HI, Indpls) U.S. Pat. 3,302,050, January 31, 1967

Tape Transport Threading Mechanism Having Movable Pressure Roller—T. C. Weathers, H. Jensen (HI, Indpls) U.S. Pat. 3,298,583, January 17, 1967

Stereophonic FM Receivers Having Automatic Switching Means for Stereo Reception—J. F. Merritt (HI, Pr) U.S. Pat. 3,294,912, December 27, 1966

Meetings

MARCH 20-22, 1967: **Physical Process in the Lower Atmosphere**, U. of Michigan American Meteorological Soc. **Prog. Info.:** A. X. Wiin-Nielsen, Meteorology and Oceanography Dept., U. of Michigan, 2038 East Engineering Bldg., Ann Arbor, Mich.

MARCH 20-23, 1967: **IEEE International Convention & Exhibition**, All Groups & TAB Comms., Coliseum & N.Y. Hilton Hotel, N.Y., N.Y. **Prog. Info.:** IEEE Hdqs., 345 E. 47th St., N.Y., N.Y. 10017.

MARCH 28-30, 1967: **6th Photovoltaic Specialists Conference**, G-ED, Sheraton Cape Colony Inn, Cocoa Beach, Florida. **Prog. Info.:** W. R. Cherry, NASA, Goddard Space Flight Center, Greenbelt, Md.

APRIL 5-7, 1967: **Int'l Magnetics Conference** (INTERMAG) G-Mag, Shoreham Hotel, Wash., D.C. **Prog. Info.:** R. F. Elfant, IBM, Yorktown Heights, N.Y.

APRIL 10-14, 1967: **Conference on P.A.L. Colour Television Systems**, IEEE, U. K. & Eire Section, Univ. of Nottingham, Nottingham, Eng. **Prog. Info.:** IEEE Hdqtrs., 345 E. 47th St., N.Y., N.Y. 10017.

APRIL 11-13, 1967: **Cleveland Electronics Conference**, Cleveland Section, et al., Cleveland Engrg. Center, Cleveland, Ohio. **Prog. Info.:** Mike Lapine, Cleveland Elec. Conf. Inc., 616 Hanna Bldg., Cleveland, Ohio.

APRIL 17-19, 1967: **Region 3 Meeting**, Region 3, Heidelberg Hotel, Jackson, Miss. **Prog. Info.:** J. E. May, 1120 Auburn Dr., Jackson, Miss.

APRIL 17-19, 1967: **Thermal Balance of Spacecraft**, American Inst. of Aeronautics and Astronautics, Nat'l Bureau of Standards, Air Force. **Prog. Info.:** Y. S. Rouloukian, Thermophysical Properties Research Center, Purdue U., 2595 Yeager Rd., Lafayette, Ind.

APRIL 18-19, 1967: **Electronics & Instrumentation Conf. & Exhibition**, IEEE Cincinnati Sec., ISA, Carousel Inn, Cincinnati Gardens, Cincinnati, Ohio. **Prog. Info.:** IEEE Headquarters, 345 E. 47th St., N.Y., N.Y. 10017.

APRIL 18-20, 1967: **Spring Joint Computer Conference**, IEEE-AFIPS, Chalfonte-Haddon Hall, Atlantic City, N.J. **Prog. Info.:** M. P. Chinitz, 326 Township Line Rd., Norristown, Pa. 19403.

APRIL 19-20, 1967: **Electronics & Instrumentation Conference & Exhibition**, Cincinnati Section, ISA, Carousel Inn, Cincinnati Gardens, Cincinnati, Ohio. **Prog. Info.:** R. H. Englemann, Univ. of Cincinnati, Cincinnati, Ohio.

APRIL 19-21, 1967: **Southwestern IEEE Conf. & Exhibition** (SWIEEECO), Region 5, Dallas Memorial Auditorium, Dallas, Texas. **Prog. Info.:** A. A. Dougal, Univ. of Texas, Engrg.-Science Bldg., Austin, Texas 78712.

APRIL 19-22, 1967: **Semiconductor Device Research Conf.**, IEEE Region 8 et al., Bad Nauheim, Germany. **Prog. Info.:** Prof. W. J. Kleen, 8 Munchen 8 (F.R. Germany) Blanastr. 73.

APRIL 24-MAY 5, 1967: **Conference on Integrated Circuits**, Region 8, U. K. & Eire Section, et al., London, England. **Prog. Info.:** IEEE Headquarters, 345 E. 47th St., N.Y., N.Y. 10017.

MAY 1-2, 1967: **4th Annual Rocky Mountain Bioengineering Symp.**, Denver section, Univ. of Colorado Medical School, Denver, Colo. **Prog. Info.:** IEEE Headquarters, 345 E. 47th St., New York, N.Y. 10017.

PROFESSIONAL MEETINGS

DATES *and* DEADLINES

Be sure deadlines are met—consult your Technical Publications Administrator or your Editorial Representative for the lead time necessary to obtain RCA approvals (and government approvals, if applicable). Remember, abstracts and manuscripts must be so approved BEFORE sending them to the meeting committee.

MAY 2-3, 1967: **Insulation Coordination Forum**, G-P, Little Rock Section, Marion Hotel, Little Rock, Arkansas. **Prog. Info.:** J. O. Noel, Illinois Power Co., 500 South 27th St., Decatur, Illinois.

MAY 2-4, 1967: **Conference on Integrated Circuits**, U. K. & Eire Section, et al., Eastbourne, Eng. **Prog. Info.:** IEEE Headquarters, 345 E. 47th St., New York, N.Y. 10017.

MAY 3-5, 1967: **Joint Parts, Materials & Packaging Conf.** (formerly the Electronic Components Conf.), IEEE G-PMP, EIA, Marriott Motor Hotel, Washington, D.C. **Prog. Info.:** C. K. Morehouse, Globe Union Inc., Box 591, Milwaukee, Wisc. 53201.

MAY 3-5, 1967: **8th Conference on Human Factors in Electronics**, G-HFE, Cabana Motor Hotel, Palo Alto, Calif. **Prog. Info.:** J. C. Bliss, Stanford Univ., Stanford, Calif. 94305.

MAY 3-6, 1967: **Rare Earths**, Oak Ridge Nat'l Lab, Air Ford Office of Scientific Research. **Prog. Info.:** W. C. Koehler, ORNL, Oak Ridge, Tenn.

MAY 8-10, 1967: **G-MTT Int'l Microwave Symposium**, G-MTT, Hilton Hotel & New England Life Hall, Boston, Mass. **Prog. Info.:** T. Saad, Sage Labs., 3 Huron Drive, Natick, Mass.

MAY 9-11, 1967: **Packaging Industry Technical Conference**, G-IGA, Holiday Inn, New York, N.Y. **Prog. Info.:** IEEE Headquarters, 345 E. 47th St., New York, N.Y. 10017.

MAY 9-11, 1967: **Region 6 Conference**, Region 6, Western Skies & Holiday Inn, Albuquerque, N.M. **Prog. Info.:** O. M. Stuetzer, Sandia Corp., Albuquerque, N.M.

MAY 9-11, 1967: **9th Annual Symposium on Electron, Ion and Laser Beam Technology**, G-ED, Univ. of Calif., Berkeley, Calif. **Prog. Info.:** IEEE Headquarters, 345 E. 47th St., New York, N.Y. 10017.

MAY 15-17, 1967: **IEEE Aerospace Elec. Conv.** (NAECON), G-AES, Dayton Section, Dayton, Ohio. **Prog. Info.:** IEEE Dayton Office, 124 E. Monument Ave., Dayton, Ohio 45402.

MAY 15-17, 1967: **Power Industry Computer Application Conf.** (PICA), G-P, Pittsburgh Hilton, Pittsburgh, Pa. **Prog. Info.:** L. W. Coombe, Detroit Edison Co., 2000 Second Ave., Detroit, Mich. 48226.

MAY 16-18, 1967: **Nat'l Telemetering Conf.**, IEEE, AIAA, ISA, San Francisco, Hilton Hotel, San Francisco, Calif. **Prog. Info.:** M. A. Lowy, Gen'l Elec. Co., P.O. Box 8048, Phila., Pa.

MAY 18-19, 1967: **10th Midwest Symposium on Circuit Theory**, G-CT & Purdue University, Purdue University, Lafayette, Indiana. **Prog. Info.:** B. J. Leon, Purdue Univ., Lafayette, Indiana 47907.

MAY 22-24, 1967: **Frequency Generation & Control for Radio Systems**, IEE, U. K. Eire Section, Savoy Place, London, England. **Prog. Info.:** R. G. Cox, IEE, Savoy Place, London, W.C. 2, England.

MAY 22-25, 1967: **Industrial & Commercial Power Systems Tech. Conference**, G-IGA, Cleveland Section, Statler Hilton Hotel, Cleveland, Ohio. **Prog. Info.:** A. B. Gipe, Albert B. Gipe Assoc., 1226 N. Charles St., Baltimore, Md.

MAY 23-26, 1967: **Joint Canada-USA URSI Meeting**, URSI-IEEE, National Research Council, Ottawa, Ontario, Canada. **Prog. Info.:** George Sinclair, Galbraith Bldg., Univ. of Toronto, Toronto 5, Canada.

Calls for Papers

JUNE 19-21, 1967: **San Diego Symposium for Biomedical Engineering**, IEEE, U.S. Naval Hospital, et al., San Diego, Calif. **Deadline Info.: 4/12/67** (Abst.) TO: D. L. Franklin, Scripps Clinic & Res. Foundation, La Jolla, Calif.

JULY 9-14, 1967: **Summer Power Mtg.**, IEEE, G-P, Portland Hilton Hotel, Portland, Oregon. **Deadline** (Papers) **4/12/67** TO: E. C. Day, IEEE, 345 E. 47th St., New York, N.Y. 10017.

JULY 10-14, 1967: **Nuclear & Space Radiation Effects Conference**, G-NS, Ohio State Univ., Columbus, Ohio. **Deadline Info.:** J. L. Wirth, Sandia Corp., Albuquerque, New Mexico.

JULY 18-20, 1967: **9th Electromagnetic Compatibility Symposium**, IEEE, G-EMC, Shoreham Hotel, Washington, D.C. **Deadline Info.:** F. T. Mitchell, Atlantic Res. Corp., Shirley Hwy. & Edsall Rd., Alexandria, Va.

AUG. 13-17, 1967: **2nd Intersociety Energy Conversion Engineering Conf.**, ASME, IEEE, AIChE, ANS, SAE, AIAA, Fountainebleu Hotel, Miami Beach, Fla. **Deadline Info.:** IEEE Headquarters, 345 E. 47th St., New York, N.Y. 10017.

AUG. 22-25, 1967: **Western Electronic Show & Convention** (WESCON), IEEE-WEMA, Cow Palace, San Francisco, Calif. **Deadline** (Abst.) **5/15/67**, TO: WESCON, 3600 Wilshire Blvd., Los Angeles, Calif.

SEPT. 4-8, 1967: **Solid-State Devices Conference**, U. K. & Eire Section, et al., Univ. of Manchester Institute of Science and Technology, Manchester, Eng. **Deadline Info.:** L. Lawrence, Inst. of Physics and Physical Society, 47 Belgrave Sq., London, S.W. 1, England.

SEPT. 6-8, 1967: **First Computer Conference**, G-C, N.W. Univ., Chicago Section, Edgewater Beach Hotel, Chicago, Illinois. **Deadline Info.:** IEEE Headquarters, 345 E. 47th St., New York, N.Y. 10017.

SEPT. 11-15, 1967: **Int'l Symposium on Information Theory**, G-IT, Kings Palace Hotel, Athens, Greece. **Deadline:** (Ms.) **5/1/67** TO: IEEE Headquarters, 345 E. 47th St., New York, N.Y. 10017.

SEPT. 24-28, 1967: **Joint Power Generation Conference**, G-P, ASME, et al., Statler Hilton Hotel, Detroit, Michigan. **Deadline: 6/23/67** TO: Henry Wallace, Jr., 110 S. Orange, Livingston, N.J.

SEPT. 26-28, 1967: **Conference on Magnetic Materials and Their Applications**, U. K. & Eire Section, et al., London, England. **Deadline Info.:** IEEE Hdqts., 345 E. 47th St., New York, N.Y. 10017.

SEPT. 25-27, 1967: **Int'l Electronics Conf. & Exposition of the Canadian Region**, Canadian Region, Toronto Section, Automotive Bldg. in Exhibition Pk., Toronto, Ontario, Canada. **Deadline Info.:** Rudolph G. deBuda, 1819 Yonge St., Toronto, Ontario, Canada.

OCT. 9-10, 1967: **Joint Engineering Management Conference**, ASME-IEEE, et al., Jack Tar Hotel, San Francisco, Calif. **Deadline:** (Papers) **4/1/67** TO: B. B. Winer, Westinghouse Elec. Corp., East Pittsburgh, Pa.

OCT. 11-13, 1967: **System Science & Cybernetics Conference**, G-SSC, Statler Hilton Hotel, Boston, Mass. **Deadline Info.:** David Smith, Moore School of Electrical Engrg., Univ. of Penna., Phila., Penna.

OCT. 16-18, 1967: **Aerospace & Electronic Systems Convention**, G-AES, Sheraton Park Hotel, Washington, D.C. **Deadline Info.:** Donald Hagner, Bellcomm Inc., 1100 17th St. N.W., Washington, D.C.

OCT. 16-18, 1967: **Fall URSI-IEEE Meeting**, URSI-IEEE, Rackham Bldg., Univ. of Michigan, Ann Arbor, Mich. **Deadline Info.:** T. B. A. Senior, Radiation Lab., 201 Catherine St., Ann Arbor, Michigan.

OCT. 17-19, 1967: **Int'l Symposium on Antennas & Propagation**, G-AP, Rackham Bldg., Univ. of Michigan, Ann Arbor, Mich. **Deadline Info.:** T. B. A. Senior, Radiation Lab., 201 Catherine St., Ann Arbor, Michigan.

OCT. 18-20, 1967: **Electron Devices Meeting**, G-ED, Sheraton-Park Hotel, Washington, D.C. **Deadline Info.:** IEEE Headquarters, 345 E. 47th St., New York, N.Y. 10017.

OCT. 23-25, 1967: **6th Symposium on Adaptive Processes**, G-AC, G-IT, G-SSC, McCormick Place, Chicago, Illinois. **Deadline Info.:** IEEE Headquarters, 345 E. 47th St., New York, N.Y. 10017.

OCT. 23-25, 1967: **Nat'l Electronics Conf.**, IEEE, et al., McCormick Place, Chicago, Illinois. **Deadline Info.:** Nat'l Electronics Conf., 228 N. LaSalle St., Chicago, Illinois.

OCT. 1967: **Electron Devices Meeting**, G-ED, Washington, D.C. **Deadline Info.:** IEEE Headquarters, 345 E. 47th St., New York, N.Y. 10017.

OCT. 30-NOV. 2, 1967: **14th Nuclear Science Symposium**, G-NS, Statler Hilton Hotel, Los Angeles, California. **Deadline Info.:** R. C. Maninger, Lawrence Radiation Lab., P.O. Box 808, Livermore, Calif.

NOV. 1-3, 1967: **Northeast Res. & Engrg. Meeting** (NEREM), New England Section, Boston-Sheraton Hotel, Boston, Mass. **Deadline:** (Abst.) **6/15/67** TO: IEEE Boston Office, 31 Channing St., Newton, Mass.

M&SR NAMES COTTLER AS CHIEF ENGINEER

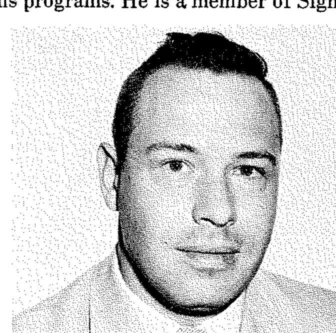
J. H. Sidebottom, Division Vice President and General Manager, Missile and Surface Radar Division has announced the appointment of: **D. M. Cottler** as Chief Engineer; **W. V. Goodwin,** Manager, Marketing Department; and **E. W. Petrillo,** Manager, Program Management. The new organization, Mr. Sidebottom said, "will simplify operations and shorten lines of communications, particularly in the acquisition of new business.

Mr. Cottler received a BChE from CCNY in 1942. He did graduate study in electrical engineering at the Polytechnic Institute of Brooklyn. Mr. Cottler joined RCA Moorestown after eight years at the Signal Corps White Sands Proving Ground in New Mexico. As Chief of the Engineering Division there he was responsible for all radar range instrumentation. His initial RCA responsibility was Project Leader for tracking-radar development and field evaluation in the Talos Land-Based Missile Program. Mr. Cottler later became Manager, General Information Processing, in the M&SR Engineering Department. Immediately prior to his new assignment, Mr. Cottler was Chief Engineer of the SAM-D Program. He was a charter member of the Inter-Range Instrumentation Working Group on Electronics Trajectory Measurements and is a member of the American Institute of Aeronautics and Astronautics.

As Manager, Programs Management, **Mr. Petrillo** has the responsibility for the management of all contracts. In addition, he will support the Marketing Department in the acquisition of new business for established product lines and extensions thereof. Mr. Petrillo received BSEE from Yale University in 1939. He is a member of the American Ordnance Association and Tau Beta Pi. Mr. Petrillo joined the Signal Corps Engineering Laboratories (SCEL) in 1943 and served ten years as Assistant Director of Engineering for Communications and Chief of Plans and Programs. In 1953, he was appointed Director of the SCEL Radar Division, and spent three years in management supervision of all Signal Corps R&D Programs in the radar guided missile electronics areas. He joined RCA's Missile and Surface Radar Division in 1956 as Manager, Associated Systems Projects Engineering for range instrumentation projects. Since that time, he has held many responsible engineering management positions. Mr. Petrillo was Deputy Program Manager of Range and Re-Entry programs prior to his new assignment.

As Marketing Manager, **Mr. Goodwin** has the responsibility for all marketing including planning for new business, sales, and contract administration for all M&SR programs. Mr. Goodwin received his BEE in 1949 from Rensselaer Polytechnic Institute and an MSEE in 1955 from the University of Pennsylvania. Upon joining RCA in 1949, he worked on Radar Systems Development and was responsible for the final development and evaluation of the TERRIER monopulse tracking radar. Prior to his new assignment, Mr. Goodwin served as Deputy Program Manager for Navy Weapons Systems programs. He is a member of Sigma Xi.

E. W. Petrillo



W. V. Goodwin

EDISON MEDAL FOR DR. BROWN

Dr. George H. Brown, Executive Vice President, Research and Engineering, in March will receive the Edison Medal at the IEEE 1967 International Convention and Exhibition in New York. He is being honored "for a meritorious career distinguished by significant engineering contributions to antenna development, electromagnetic propagation, the broadcast industry, the art of radio frequency heating, and color television."

DR. SIMON LARACH NAMED FELLOW OF RCA LABORATORIES

In recognition of his outstanding scientific contributions to the field of luminescence, **Dr. Simon Larach** has been appointed Fellow, Technical Staff, RCA Laboratories. The RCA Laboratories designation of Fellow is comparable to the same title used by universities and technical societies. It is given in recognition of a record of sustained technical contribution in the past and of anticipated continued contribution in the future.

HENSMAN CHIEF ENGINEER OF NEW RCA DIVISION

Harry G. Hensman is the Chief Engineer of the newly formed RCA Magnetic Products Division in Indianapolis. He reports to **J. Stefan,** Division Vice President and General Manager. Prior to the formation of the division on January 1, Mr. Hensman had been Manager, Magnetic Tape Engineering, in the RCA Victor Record Division.

In 1953, Mr. Hensman received BS degrees in Mechanical and Chemical Engineering from the University of Detroit. He then worked at the E. I. duPont Yerkes Research Laboratories with his activities centering on chemical reactions and the development of processes for industrial films, such as Mylar, Teflon, and others. He also worked for the Ampex Corporation before joining the RCA Victor Record Division in 1964.

CHANGES IN INTERNATIONAL OPERATIONS

President **Robert W. Sarnoff** has announced the following changes regarding RCA international responsibilities:

1. Responsibility for negotiation and administration of foreign patent licensing, and technical aid agreements is transferred to the Vice President, Licensing;
2. Responsibility for Record Matrix Licensing is transferred to the RCA Victor Record Division;
3. Responsibility for foreign purchasing is transferred to the Executive Vice President, Manufacturing Services and Materials;
4. Responsibility for export marketing of their products is transferred to the respective domestic organization;
5. Responsibility for foreign product manufacturing "feeder" operations for domestic requirements is transferred to the respective domestic organization.

The RCA International Division will continue to be responsible for managing existing foreign manufacturing subsidiary companies, other than feeder operations.

The RCA International Division will also have the responsibility for providing specialized international regional services and international functional services, such as credit and collection, financial, marketing services, personnel, and distribution and commercial relataions, to all product and service divisions.

RCA PROFITS AND SALES SET RECORDS IN 1966

RCA in 1966 achieved the largest sales and profit increase for a single year in its 47-year history, **Elmer W. Engstrom,** Chairman of the Executive Committee and **Robert W. Sarnoff,** President, have announced.

The year will be the fifth in succession in which the company's sales and earnings have reached new peaks, the two executives said, and they added that "we have every expectation that this momentum will continue in 1967."

"Barring major changes in the economic climate," they continued, "we believe that RCA will emerge at the end of 1967 with significant further increases in both sales and profits, an even more diversified base of operations, and an organization that is increasingly international in outlook and technology."

Subject to final confirmation, RCA's sales for 1966 will surpass $2.5 billion and profits will exceed $130 million. The comparable figures for 1965 were $2 billion and $102 million.

Dr. Engstrom and Mr. Sarnoff said that nearly every major operating unit of RCA contributed to the overall progress of the corporation, and that nine of the company's divisions and subsidiaries advanced to new all-time high profit levels. Among these, they cited RCA's home instruments, electronic components and devices, service, and communications operations, and the National Broadcasting Company and Random House, Inc.

They pointed out that RCA undertook in 1966 the largest capital expenditure program in its history, amounting to a more than $200 million domestic investment in a variety of activities. During the year, the company obtained or undertook construction of eleven new manufacturing plants in the United States and abroad, and major expansions at fifteen existing plant locations.

The executives said that color television "in its two principal aspects—manufacturing and broadcasting—made the largest contribution to the company's overall progress."

"RCA leads the industry in color TV set production and sales, and we expect to maintain our leadership in the growing market during 1967," they said. "RCA's color set manufacturing capacity in 1967 will be three times that of 1965, and we anticipate that every bit of it will be needed to meet the continuing public demand for color."

They added that sales of RCA home instruments in 1967 are expected to exceed $1 billion, representing a doubling of the company's sales volume of these products in only two years.

PRODUCTION STARTED AT NEW SCRANTON PLANT

The first color picture tube came off the production line on Jan. 18 at RCA's new $26 million plant in Scranton, Pa., marking the start of manufacturing operations. The rectangular tube, with a viewing area of 227-square inches, completed tests with "flying colors," according to **Harry Seelen,** Division Vice President and General Manager, RCA Picture Tube Division. Mr. Seelen joined **Joseph H. Colgrove,** Scranton Plant Manager, and members of the latter's staff in observing the event.

CIRCULARLY POLARIZED FM BROADCAST ANTENNA

The first practical FM broadcast antenna to employ circular polarization for improved signal reception by automobile radios has been announced by the RCA Broadcast and Communications Products Division.

With the growing popularity of FM car radios, broadcast stations have been adding a separate vertically-polarized transmitting antenna for better reception by the automobile's upright "buggy-whip" aerial. Normally, the FM signal is horizontally polarized.

The new FM transmitting antenna reduces by 50 percent the weight and wind load over a combination of separate horizontally and vertically polarized antennas needed for an equivalent broadcasting job.

The new RCA antenna uses a radiating element made of two circular dipoles to produce circular polarization with a relatively simple radiator. An advantage of circular polarization is that transmitter power can be doubled without exceeding the licensed maximum in the horizontal plane since the additional power is radiated in other polarizations.

FIRST HIGH-SPEED WEATHER CIRCUIT BETWEEN U.S. AND GERMANY

RCA Communications, Inc. has announced the opening of the first high-speed weather communications link between the United States and Eastern Europe. The circuit was officially placed into operation during a ceremony at the Weather Bureau's National Meteorological Center in Suitland, Maryland. The new high-speed coaxial cable circuit links the U.S. Weather Bureau center in Maryland directly with the National Weather Service center at Offenbach (Frankfurt), Germany. The circuit is routed via the newest transatlantic cable (TAT-4), directly connecting the United States with Continental Europe.

The new circuit is expected to revolutionize the exchange of weather information between the two continents. Capable of voice, data, or pictorial transmission, the circuit replaces a link which carried only teletypewriter messages.

Meteorological data will be transmitted at the speed of 1050 words per minute, more than 10 times the 100-word-per-minute capability of the earlier circuit. Weather maps which were previously not exchanged will now be transmitted in both directions at a speed of 120 rpm. Voice Communications will be utilized for cue and control, maintenance and other functions.

135-POUND COLOR TV CAMERA

A 135-pound color TV camera designed for on-the-spot coverage of news, sports and other events away from the studio has been announced by the RCA Broadcast and Communications Products Division. The TK-44 field camera will be available by mid-1968 in time for the Democratic and Republican National Conventions.

Weighing approximately 135 pounds, less detachable viewfinder and lens, the camera combines ease of handling and operation with the superior performance of RCA's four-tube pickup system. The new camera follows RCA's design concept of a high-quality separate luminance channel. It employs a three-inch image orthicon type tube, with improved signal-to-noise ratio, for the luminance signal. The red, green and blue channels will use one-inch electrostatic focus tubes of the vidicon type.

E C & D REALIGNMENT

A reorganization of RCA's electronic components operational units and management staff has been announced by **John B. Farese,** Vice President, RCA Electronic Components and Devices. Mr. Farese said that the move was made to keep pace with advances in technology, expanding world markets, and to maintain RCA's strong position in the highly competitive electronic components business. "A major objective of the reorganization," he stated, "is to consolidate all solid-state devices and receiving tube activities into a single division. Formerly, these activities were spread over three divisions."

The following new appointments have been made:

C. E. Burnett, appointed Division Vice President and General Manager, Solid-State and Receiving Tube Division. This newly-formed division will include: the Special Electronic Components Department (integrated circuits, direct energy conversion devices and superconductive products); Industrial Semiconductor Operations Department; and the Commercial Receiving Tube and Semiconductor Operations and Marketing Departments.

William H. Painter, Division Vice President, Electronic Components and Devices International Operations.

Joseph T. Cimorelli, General Manager, Memory Products Division.

Gene W. Duckworth, General Manager, Industrial Tube Division. This newly created division will be responsible for the engineering, production and marketing of power tubes, conversion tubes and microwave devices.

Joseph A. Haimes, Manager, RCA Distributor Products, responsible for the overall sales and marketing of components and devices in the distributor market.

Harry R. Seelen continues as Division Vice President and General Manager, Television Picture Tube Division.

Mr. Farese also announced these new appointments to his management staff: **Harold F. Bersche,** Division Vice President, Distributor Marketing Relations.

Michael J. Carroll, Equipment Marketing Relations. **Howard C. Enders,** Manager, News and Information.

The following members of Mr. Farese's staff continue in their present positions: **G. C. Brewster,** Manager, Operations Planning and Support; **William C. Dove,** Purchasing Agent; **Alan M. Glover,** Division Vice President, Technical Programs; **Lawrence A. Kameen,** Manager, Personnel; and **Julius Koppelman,** Controller, Finance.

All members of the management staff, except Mr. Duckworth, have headquarters located at the home office of RCA Electronic Components and Devices in Harrison, N.J. Mr. Duckworth's office is at Lancaster, Pa.

JOHNSON ON ELECTRIC-CAR PANEL

Edward O. Johnson, Manager, Engineering, EC&D Technical Programs, has been named to a 16-man panel established by the U. S. Secretary of Commerce to investigate the possibilities of developing an electrically powered automobile. The panel, which includes experts on transportation, technology, economics, and associated fields, such as air pollution, is headed by Dr. Richard S. Morse of the MIT Sloan School of Management.

The group's assignments include: Survey the current electric car technology. Determine the technical and economic feasibility of electric cars. Compare their performance and effects of electric vehicles. Recommend a role for the Federal Government in the research and development of such cars.

STAFF ANNOUNCEMENTS

Lawrence M. Isaacs has been named Vice President and Controller and **Earl S. Kauffman,** Staff Vice President, Management Information Systems. They report to **Howard L. Letts,** Executive Vice President, Finance.

Robert J. Anglis appointed Vice President, Sales, RCA Communications, Inc., reporting to **Howard R. Hawkins,** President.

E. J. Derenthal named Director, Transportation, Warehousing, and Physical Distribution, by **G. A. Fadler,** Staff Vice President, Materials.

Delbert L. Mills, Executive Vice President, Consumer Products, has been elected to the RCA Board of Directors.

Robert L. Werner named Executive Vice President and General Counsel and **E. M. Tuft** elected Executive Vice President, Personnel, both reporting to President **Robert W. Sarnoff.**

Frederick M. Hoar appointed Director, Publications and Exhibits, and **R. Kenyon Kilbon** named Director, Editorial Services. Both report to **Alexander S. Rylander,** Staff Vice President, News and Information.

Kenneth L. Snover named Manager, Manufacturing Operations, RCA Electronic Data Processing, by **James R. Bradburn,** Vice President and General Manager.

Norman Racusin appointed Division Vice President and General Manager, RCA Victor Record Division, report to **Charles M. Odorizzi,** Group Executive Vice President.

G. B. Herzog, appointed Director, Process Research and Development Laboratory, by **J. Hillier,** Vice President, RCA Laboratories.

K. K. Miller named Manager, Defense Planning and **R. A. Newell,** Manager, Special Plans and Programs, reporting to **H. I. Miller,** Division Vice President, DEP Defense Planning and Regional Offices.

M. H. Glauberman appointed Manager, Plans and Programs by **A. L. Malcarney,** Executive Vice President, Manufacturing Services and Materials.

K. Hesdoerffer named Manager, Manufacturing Operations and Product Assurance, reporting to **S. W. Cochran,** Division Vice President and General Manager, Graphic Systems Division.

H. F. Kazanowski, appointed Manager, Business Analysis, by **F. H. Erdman,** Division Vice President, New Business Programs.

J. Stefan named Division Vice President and General Manager, Magnetic Products Division, reporting to **C. M. Odorizzi,** Group Executive Vice President.

J. J. Benavie, Staff Vice President, Domestic Licensing, and **S. S. Barone,** Staff Vice President, International Licensing, both reporting to **M. E. Karns,** Vice President, Licensing.

M. R. Amsler, appointed Manager, Programs Management, reporting to **J. M. Hertzberg,** Division Vice President and General Manager, CSD.

E. Noel Luddy named Manager, Broadcast and Communications Consultant Relations in Washington, D.C. He reports to **E. C. Tracy,** Division Vice President, Broadcast Sales Department, B&CP.

A. D. Beard, EDP, Chief Engineer, announces his staff as: **C. N. Breeder,** Manager, Administration and Control; **L. Iby,** Manager, Design Integration; **H. Kleinberg,** Manager, Engineering—Palm Beach; **J. L. Maddox,** Manager, Systems Engineering; **J. N. Marshall;** Manager, Computer and Communications Product Engineering; **H. J. Martell,** Manager, Peripheral Equipment Program; and **G. D. Smoliar,** Staff Engineer.

J. P. Veatch, Director, RCA Frequency Bureau, announces his staff as: **J. F. Eagan, Jr.,** Manager, Camden Office; **W. Mason,** Manager, Maritime Projects; and **R. E. Simonds,** Manager, New York Office.

C. H. Colledge, Division Vice President and General Manager, BC&P, announces his staff as: **A. L. Hammerschmidt,** Manager, Electronic Recording Products and Scientific Instruments Department, **E. J. Hart,** Manager, Communications Products Department, **A. F. Inglis,** Division Vice President, Engineering and Merchandising Department, **A. M. Miller,** Division Vice President, Instructional Electronics Department, **J. P. Taylor,** Manager, Marketing Services, and **E. C. Tracy,** Division Vice President, Broadcast Sales Department.

In the Engineering and Merchandising Department, **A. F. Inglis,** Division Vice President, announces his staff as follows: **J. H. Cassidy,** Manager, Sales Support and Services, **T. M. Gluyas,** Manager, Broadcast Audio and Transmitter Engineering, **H. N. Kozanowski,** Manager, TV Advanced Development, **A. H. Lind,** Manager, Studio Equipment Engineering, **R. L. Rocamora,** Manager, Antenna Engineering and Merchandising, **W. B. Varnum,** Manager, Studio Equipment Merchandising, **C. A. Wallack,** Manager, Broadcast Audio and Transmitter Merchandising, **H. S. Wilson,** Manager, Microwave Engineering and Instructional TV Merchandising and Engineering, **J. E. Young,** Manager, Systems Engineering and Administration.

... PROMOTIONS ...

to Engineering Leader & Manager

As reported by your Personnel Activity during the past two months. Location and new supervisor appear in parentheses.

Electronic Components and Devices

E. L. Batz: from Sr. Liaison Eng. to *Ldr. Liaison Engr.* (J. Wright, Bloomington)

R. J. Shedlak: from Engr. Manu. to *Mgr. Tube Parts* (M. B. Still, Marion)

M. K. Brown: from Sr. Eng. to *Eng. Ldr. Prod. Devel.* (Mgr. Color Devel., Lancaster)

F. Donovan: from Sr. Mbr. Tech. Staff to *Ldr. Tech. Staff* (L. Wood, Needham)

W. Goulder: Engr., Equip. Devel. to *Ldr. Tech. Staff* (L. Wood, Needham)

D. A. Campbell: from Sr. Mbr. Tech. Staff to *Ldr. Tech. Staff* (E. A. Schwabe, Needham)

P. C. March: from Sr. Mbr. Tech. Staff to *Ldr. Tech. Staff* (E. A. Schwabe, Needham)

J. J. Cosgrove: from Sr. Mbr. Tech. Staff to *Ldr. Tech. Staff* (E. A. Schwabe, Needham)

Broadcast and Communications Division

L. V. Hedlund: from Eng. D&D to *Ldr. D&D Eng.* (R. N. Hurst, Camden)

J. R. West: from Eng. D&D to *Ldr. D&D Eng.* (R. N. Hurst, Camden)

Missile and Surface Radar Division

B. C. Stephens: from Cl. "A" Eng. to *Ldr. D&D Eng.* (W. S. Perecinic, Mrstn.)

W. S. Perecinic: from Ldr. D&D Eng. to *Mgr. Advanced Design Engrs.* (H. Eigner, Mrstn.)

Communications Systems Division

E. VanKeuren: from Eng. "A" to *Ldr. Systems Projects* (J. Santoro, Camden)

E. D. Menkes: from "A" Engr. to *Ldr. D&D Eng.* (D. H. Westwood, Camden)

D. Hampel: from Sr. Proj. Mbr. Tech. Staff to *Ldr. Tech. Staff* (D. P. Goodwin, Camden)

C. J. Moore: from "AA" Eng. to *Ldr. D&D* (T. L. Genetta, Camden)

E. J. Westcott: from Ldr. to *Mgr., Systems Assurance* (C. G. Arnold, Camden)

M. Raphelson: from Mgr., Systems Assurance to *Mgr., Secure Communications* (J. M. Osborne, Camden)

West Coast Division

R. Norwalt: from Sr. Mbr. D&D to *Ldr. D&D Eng.* (G. Turner, Van Nuys)

A. Virnig: from Sr. Mbr. D&D to *Ldr. D&D Eng.* (G. Grondin, Van Nuys)

Astro-Electronics Division

D. Williams: from Eng. to *Ldr. Engrs.* (M. Shepetin, Princeton)

Electronic Data Processing

R. L. Cox: from Engr. to *Ldr. D&D Eng.* (G. Waas, Camden)

J. E. Linnell: from Ldr. D&D to *Mgr. Disc File Eng.* (H. J. Martell, Camden)

R. H. Yen: from Ldr. D&D to *Mgr. Computer Logic Design* (J. Marshall, Camden)

T. A. Franks: from Mgr. Emulator Design to *Mgr. Computer Projects* (J. Maddox, Camden)

PROFESSIONAL ACTIVITIES

R&E Product Engineering, Camden, N.J.: **G. A. Keissling,** Manager, Product Engineering Professional Development, has been reappointed to the IEEE Professional Relations Committee for 1967.

ASD, Burlington, Mass.: **Oliver T. Carver** has been renamed to the Technical Committee on Support Systems of the American Institute of Aeronautics and Astronautics for 1967.—*D. Dobson*

Dr. Alfred N. Goldsmith, Honorary Vice President, RCA, has been named a Director Emeritus of the IEEE for 1967.

ASD, Burlington, Mass.: **M. C. Kidd** and **B. T. Joyce** participated in the IEEE, EIA, ME, Computer Aided Design Committee in San Francisco, last November. **M. Anderson** took part in the AGARD/NATO Lecture Series on Application of Microelectronics to Aerospace Equipment in Washington last October. **L. C. Drew** participated in the NATO Microelectronic Lectures in Washington in October. **Ed Kornstein** has been elected Chairman of the Boston section of the Society of Motion Picture and Television Engineers.—*D. Dobson*

RCA Laboratories, Princeton, N.J.: **Robert S. Hopkins** received his MSEE from Rutgers—The State University in January.

RCA INSTITUTES TO MOVE

RCA Institutes, Inc., one of this country's oldest and largest schools devoted to electronic technology, has signed a 20-year lease for occupancy of the four-story building at 320 West Thirty-first St., New York. The school, has had its headquarters at 350 West Fourth St. since 1949. The move, scheduled to be completed by Jan. 1, 1968, will provide easier access and enhanced facilities for Resident School and Preparatory Department Students.

TECHNICAL EXCELLENCE AWARD WINNERS NAMED BY M&SR

Technical Excellence Award winners for the Third Quarter of 1966 have been announced by the DEP Missile and Surface Radar Division, Moorestown, N.J. They are: **C. M. Brindley,** for his work as technical director of the Manual Space Objective Identification Technique Improvement Program; **C. P. Clasen**—for contributions to the SAM-D radar effort; **W. O. Koch**—for his contribution of an inexpensive solution to an RFI problem in the Real Time Telemetry Data System; **J. Luber**—for his performance as the lead designer on the Lunar Module Tilt Mechanism; **W. T. Patton**—for his technical leadership and contribution on the Radar Electronic Scanning Program; **R. J. Pschunder**—for his performance as senior engineer in the Mechanical Engineering Skill Center; **B. C. Stephens**—for contributions as the key project engineer on the ARIS Data Playback and Digitizing Equipment; and **W. Yanovitch**—for his work as a drafting-illustrator.

SELF-CONTAINED DISPLAY DEVICE

A new self-contained video data terminal that combines third generation circuitry with advanced computer-communications techniques has been announced by RCA Electronic Data Processing. Resembling a portable television set attached to a typewriter keyboard, the Spectra 70/752 is a compact, low-cost communications terminal that houses the video screen keyboard, controls and power supply in a single unit.

The device, the newest member of RCA's Spectra 70 family, employs third generation integrated circuits in its logic elements similar to those pioneered in the Spectra 70/35, 45 and 55 systems. Tiny silicon chips will enable the new terminal to operate with greater reliability, fewer connections, less maintenance and at lower costs.

Coupled with Spectra 70 computers and mass storage devices, the 70/752 will provide fast, economical and direct means of accessing centralized files from local or remote points for routine daily work or time-sharing. The result will mean rapid reply to inquiries, faster customer service and better business control.

The manually-controlled input-output device can display up to 1080 characters—letters, numerals or other symbols—on the 12-inch screen of a cathode ray tube at the rate of 120 characters per second. The screen can accommodate twenty 54-character lines, with clear, sharp focus maintained over the screen's entire surface.

A conventional four-row keyboard contains all necessary controls. Inquiries and transactions may be composed and visually verified. If necessary, corrections can be made by retyping before transmission. A moveable cursor, displayed as an underscore, makes positioning easy and automatic, and guards against erasure of needed information.

C. W. Fields, the Communications Systems Division Editorial Representative deserves the thanks of the Editors and readers of the RCA ENGINEER for his work in stimulating and coordinating many of the technical papers concerned with advanced communications in this issue.

NEW DATA GATHERING SYSTEM

A new third generation data gathering system for use in factories, offices, libraries, hospitals and other institutions, has been announced by RCA Electronic Data Processing. The RCA Spectra 70/630 Data Gathering System, or DGS, makes possible the transmission of data from a work area to a computer and provides management with vital information on inventory changes, down-time on machines and other production facts. The system includes various input terminals, line concentrators and buffers.

The input terminal, utilizing integrated circuits, features modular design, ease of operation and high transmission speeds.

The modular design permits the customer to mold the input station to fit his needs. A DGS station can be configured with card readers, badge readers and variable data readers, or various combinations of these.

Designed for use with the RCA Spectra 70/35, 45 or 55 computers, the DGS input station can be located as far as 30 miles from the computer without additional equipment. By using a telephone network, a manufacturer can collect data from plants or warehouses throughout the world.

The data gathering system is capable of transmitting information at a rate of 120 characters per second, and as many as 384 terminals can be connected to each Communications Control Multichannel unit (CCM) of the computer.

RCA TO OPERATE INDIAN TRAINING CENTER

The U.S. Bureau of Indian Affairs has awarded the RCA Service Company a contract to establish, manage, and operate a training center on the Choctaw Indian Reservation at Philadelphia, Miss. The training program has been especially developed to enhance the Indian's adaptability to urban living. It will employ advanced teaching and learning techniques integrated into a systems approach which has been custom designed for this requirement. Unlike most training programs, this one will upgrade the basic education of both the men and women in the Indian family while the husbands are learning skills that will help them get and hold a job, their wives will be mastering the social and homemaking skills associated with living in an urban society.

MIT RESEARCH DIRECTORY AVAILABLE IN RCA LIBRARIES

As a participant in the MIT Industrial Liaison Program, RCA has available, in its libraries, copies of the *1967 MIT Directory of Current Research.* The *Directory* lists more than 1300 projects (all nonclassified research programs) including, for example: "Conduction in Ionized Gases," "Chemiluminescence," "Engineering Education," "Computer Aided Design," "Synthesis of Switching Networks," "Plasma Physics," "Thermonuclear Systems," "Simulation of Systems," "Array Radar Techniques," "Programming Tools for Time-Sharing," "Group Factors and R&D Problem Solving," "New Product Decisions," "Thermal Interactions in Integrated Circuits," "Energy Gap of Niobium for Superconductivity," and "Analysis of Stochastic Systems."

The research projects in the *Directory* usually result in publications available to RCA engineers and scientists. Requests for information concerning the research program and publications, the MIT Summer Session Program, and a number of private symposiums to be offered for the benefit of specialists should be forwarded through the librarian in your RCA activity.

The Industrial Liaison Program also makes available to RCA a valuable service by welcoming visits by engineering and research personnel to the campus for discussions with faculty members involved in the research areas described in the *Directory.* For information concerning the campus visitation program, to arrange for a visit, or for further information concerning any aspect of the MIT Industrial Liaison Program, contact G. A. Kiessling, Manager, Product Engineering Professional Development, Camden 2-8, PC-5650.

VIDEOCOMP SYSTEM FOR TYPESETTER

An RCA electronic typesetter system capable of producing text at the rate of 600 characters per second, is scheduled for installation in mid-1967 by the Michigan Typesetting Company in Detroit.

Michigan Typesetting will use RCA's Videocomp 70/820, introduced last June by the RCA Graphic Systems Division, Princeton, N. J., as the first commercially available typesetter to employ all-electronic character generation. It can produce the entire text for a newspaper page in just 2 minutes using video and computer techniques.

The complete $1 million installation at Michigan Typesetting includes an RCA Spectra 70 computer.

Michigan Typesetting numbers among its customers the nation's four leading automobile manufacturers. Last minute changes on new models make it especially important to print large volumes of material in a short period of time.

RCA ETV SYSTEM TO LINK MEDICAL FACILITIES

A $100,000 contract to link five medical facilities in the Atlanta, Georgia, area in the nation's first 2,500 MHz television system for exchanging medical instructional programs has been announced by the RCA Instructional Electronics Department, Broadcast and Communications Products Division.

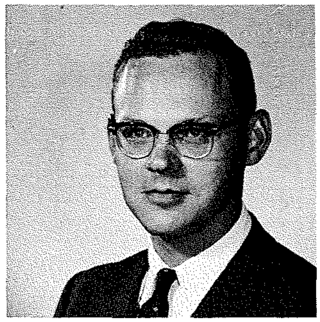
Two of the five locations—Grady Memorial Hospital and the Public Health Service Audiovisual Facility, U. S. Department of Health, Education and Welfare—will be equipped with transmitters for originating programs, and all five stations will have receiving systems.

Other participants in the network are the Atlanta Veterans Administration Hospital, Emory University School of Medicine and Hospital, and the Georgia State Department of Health with its Mental Health Institute. Roof-top antennas and down converters—devices that change the 2,500-MHz signals to a frequency that can be picked up by a standard TV set—will provide program reception at all locations.

Plans for the system, expected to be in operation next spring, call for broadcasts of medical seminars, consultations and demonstration techniques, among other material, for use in teaching medical students, nurses, and other health professionals.

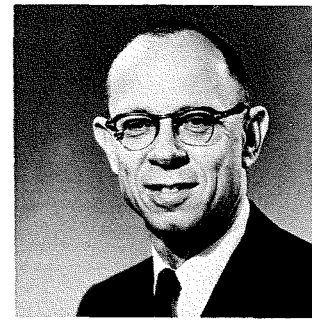
Erratum: In Fig. 9, page 36 (Vol. 12-4) of the article by Dr. Clorfeine and Dr. Taylor, the lower trace was accidentally clipped. Those wishing to see the complete traces are invited to contact the authors.—*The Editors.*

J. R. Hendrickson



G. R. Kornfeld

K. C. Shaver



G. Smoliar

J. R. HENDRICKSON NAMED ED REP FOR DEP CENTRAL ENGINEERING

John R. Hendrickson, Sr., has been appointed an RCA ENGINEER editorial representative for DEP Central Engineering, Camden. He will serve on **F. Whitmore's** DEP Editorial Board.

Mr. Hendrickson received his BS and MS degrees in Chemical Engineering from the University of Washington in 1932 and 1933, respectively. He is presently working for an MS in Systems Engineering and Operations Research at the University of Pennsylvania. He also did graduate work at the University of Washington and the University of Maryland.

Since coming to work for RCA in October 1956 as a Consulting Chemical and Nuclear Engineer, he has worked on a wide variety of space and military projects.

He is a member of the Nuclear Engineering Division of the American Institute of Chemical Engineering, the Aerospace System Safety Society, the Air Force Association, the American Nuclear Society, the American Institute of Aeronautics and Astronautics, the IEEE, the American Management Association, and the American Ordnance Association. He has published more than 50 reports in the atomic energy field. Before coming to RCA he had worked in the Army's Nuclear Defense Laboratory, the Manhattan Project, Oak Ridge, Tenn., and the Lawrence Radiation Laboratory at the University of California in Berkeley.

Mr. Hendrickson is a Registered Professional Engineer in the State of Pennsylvania and he's listed in Who's Who in Engineering, Who's Who in Industry and Technology, Who's Who in Chemistry, American Men of Science, and Who's Who in the East.

G. R. KORNFELD NAMED ED REP FOR ECD MEMORY PRODUCTS

George R. Kornfeld has been appointed to serve as RCA ENGINEER editorial representative for ECD's Memory Products Division. Mr. Kornfeld replaces L. Thomas in this capacity and will serve on **C. A. Meyer's** Editorial Board.

George R. Kornfeld graduated from Temple University in 1950 with a B.A. degree in English. After working for several months as librarian at the University, he was drafted, and was trained by the Army as a radar technician. He joined the RCA Service Company, Government Service Division, Technical Publications, as a technical writer in 1953. From 1953 to 1959 he worked as writer, editor, and group leader of field technical writing groups, preparing technical reports on the development of proximity fuses for Frankford Arsenal and Diamond Ordnance Fuse Laboratories, and on the development of ballistic aircraft escape systems for Frankford Arsenal. Mr. Kornfeld then worked for a short time preparing instruction books for the C-Stellarator at Princeton, and later was transferred to the Bedford Communications Projects Office (Massachusetts) to write evaluation reports on Time Division Data Link systems. He left the Service Company in 1960 to join the Memory Products Operation at Needham, where he was in charge of Engineering Systems and Procedures. In Spring 1961, he taught technical writing at the Cambridge Center for Adult Education. In 1964 he left RCA to work at Massachusetts Institute of Technology Instrumentation Laboratory as an Administrative Assistant on the Apollo program. He then returned to RCA in December 1965 and has been working in the design of shipping packages and engineering standards since then.

K. C. SHAVER NEW ED REP FOR B&CP MICROWAVE ENGINEERING

Karl C. Shaver has been named RCA ENGINEER editorial representative for the Microwave Engineering, Broadcast and Communications Product Division in Camden. He will serve on **D. R. Pratt's** B&CP Editorial Board.

Mr. Shaver received his BSEE from the University of Illinois in 1950 and joined Public Service Company of Northern Illinois. His work, then in the Communications Department, was concerned primarily with VHF mobile radio and telephone system. Mr. Shaver later spent two years on active duty with the Signal Corps at Fort Monmouth, N. J. in electronic countermeasures. He joined RCA in 1953 with the Systems Engineering Group of the Engineering Products Division. This work involved the preparation of technical proposals and special equipment design for microwave and VHF mobile systems. He left RCA in 1957 to work with the Lenkurt Electric Company. Mr. Shaver rejoined RCA in 1960 with the Microwave Section of Industrial Electronics Product Division. His assignments included microwave station assembly design, system test procedures, multiplex unit design and a return to systems engineering. Mr. Shaver was promoted to Leader, Systems Engineering Group, in the Broadcast and Communications Products Division in 1966.

EDP NAMES G. SMOLIAR AS TECH. PUBL. ADMINISTRATOR

Gerald Smoliar has been appointed to serve as Technical Publications Administrator for the Electronic Data Processing Division, Camden, N.J. He is a Staff Engineer, reporting to **A. D. Beard,** EDP Chief Engineer. In this capacity, Mr. Smoliar is responsible for review and approval of all technical papers and presentations by his division. He also will serve as Consulting Editor on the RCA ENGINEER; he will assist in editorial activities in cooperation with the Editors and Editorial Representatives.

Mr. Smoliar received his BSEE in 1937 from CCNY. After eight years in civil service in various positions he entered the field of digital computers in 1947 as a design engineer in the company that eventually became the Univac division of Sperry Rand. For the past 19 years he has worked continuously in this field as a designer and as a supervisor of engineers. He joined RCA in 1962 as a manager of computer product design. Mr. Smoliar is a Registered Professional Engineer in the state of Pennsylvania and holds two patents on computer components. He is the author of several articles pertaining to digital equipment and teaches logic design to engineers in the RCA training program.

Clip out and Mail to Editor, RCA ENGINEER, #2-8, Camden

RCA ENGINEER

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