Consumer information systems

Consumer Information Systems — entirely new electronic home services — have been demonstrated in the research environment. Six of the articles in this issue are therefore a timely introduction to RCA technical work related to Consumer Information Systems.

But, before such developments can become a new business, much more is required than technological prowess. In the words of RCA President, Anthony L. Conrad, "Research must be clearly aware of the markets at which the company aims. Engineering must proceed from realistic concepts of competitive costs, consumer needs, and quality demands. Marketing must be the radar — scanning the horizon of the future and relaying its findings back to the scientists and engineers. We must operate as an integrated system, with built-in feedback involving all elements of the company."

Technically, Consumer Information Systems are combinations of broadband transmission techniques, linking data facilities with consumer terminals, existing home tv receivers, and new home sensors. This is a challenging combination of many technologies requiring component and product development and systems optimization.

Commercially, this can mean many new services and sales of equipment for interactive merchandising, home surveillance, and new forms of entertainment. This could become a large new business not only for RCA but for owners of transmission facilities and service organizations.

But none of these services are now commercial realities. To determine how they can be implemented profitably for a mass market, RCA is calling on the know-how of the Laboratories, the solid state, broadcast equipment, consumer electronics and electronic industrial engineering organizations — all working with corporate marketing in an integrated effort to address technological challenges, product requirements, and market economics.

Fortunately, RCA has never lacked the financial resources and the management vision for leadership when the opportunity was ripe. But on the way to Consumer Information Systems as a profitable new RCA business, we also see some cautions. We must not only create the winning product combinations but also choose the precise timing in terms of market readiness. This is not necessarily the earliest possible timing based on technology alone.

In the years to come, technology, market research, and business planning must feed back on each other to achieve RCA leadership in Consumer Information Systems. Readers of the *RCA Engineer* can expect to hear much more of, and undoubtedly will contribute greatly to, the exciting and evolving world of Consumer Information Systems.

Richard W Sonnenfeldt

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Staff Vice President
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Our cover

... shows Bert Arnold, Manager of Distribution Systems for the Electronic Industrial Engineering Division, surrounded by two generations of CATV equipment which he developed. Pictured on the left are Series-30 amplifiers, one of the industry's first bi-directional systems. Upper right is the recently developed Series-100 amplifier which incorporates hybrid microcircuitry in amplifier modules. The original Series-30 amplifiers have been used in bi-directional systems throughout the United States since their introduction in a 1971 experiment in two-way cable television in Overland Park, Kansas.

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RCA Engineer

• 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.

Contents

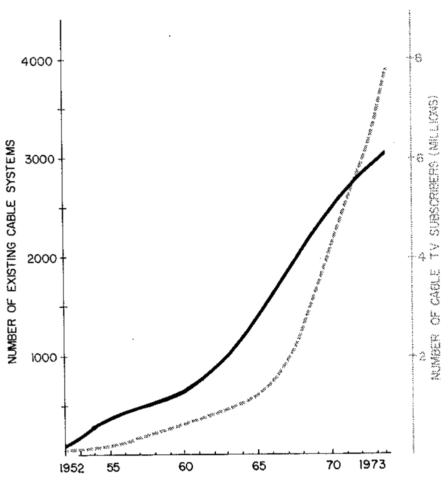
Emphasis — Broadband Communications

editorial input

CATV

By 1948, television had become a national phenomenon. More than a million television receivers were carrying Uncle Milty, Howdy Doody, Roller Derby, and Kukla, Fran, and Ollie, to many more million viewers. Yet, in those early days, television reception was a reality only for those living in the populous areas close to transmitter sites. Small towns surrounded by mountains, or located some distance from a transmitting antenna, were left out.

However, it wasn't long before individual entrepreneurs in these outlying communities started mounting antennas on the tops of mountains or tall buildings and stringing wires through tree tops, along makeshift poles, to their appliance stores and to the homes of those customers who purchased television receivers. These were the beginnings of Community Antenna Television (CATV). Small independent enterprises, usually owned and operated by local radio dealers, brought television to community groups of less than a hundred customers.



Growth of cable television over past two decades. As of July 1, 1973, there were 3032 cable systems serving 7.8 million subscribers. Note that the number of systems (black line) is starting to level off, but the number of subscribers (gray) is increasing rapidly, leading to an increase in the average size of a cable system. (Source: *Television Digest,* **Vol. 13, No. 2).**

As cable systems grew, government regulation, economic realities, and political pressures worked as conservative influences. Yet today in the U.S., more than 12% of the homes receiving television signals get those signals via cable. The cable has grown significantly and steadily (see graph), but not into the national communications system that many feel is inevitable, and not without problems. And many problems are still not resolved — *e.g.*, remuneration for the copyright owner of desirable broadcast material that is picked up and redistributed via cable.

Problems notwithstanding, most predictions point to about 30% of the nation's television households being serviced by cable by 1980. This predicted growth is no longer based on the ability of a cable system to carry a good number of television broadcasts. The future is tied to the two-way communications capabilities of cable systems: these imply a broad range of exciting new services for cable customers.

Such two-way capabilities allow a person, with a suitable terminal device used in conjunction with his television receiver, to communicate information "upstream" to other points along the cable. This could make the promise of a national cable network awesome indeed, provided the cable industry emerges from some of its current financial problems.

New services that could be offered to home subscribers include opinion polling, direct merchandising, narrowcasting, home security, interactive education, and many more; a complete list could fill pages.

Mr. Sonnenfeldt — in his cover message on Consumer Information Systems — points to the technical feasibility of such services and to some of the near-term obstacles that must be overcome before these services can be a reality. Six of the articles that lead off this issue address some of technical implications of such broadband communications services. Significantly, two of these papers serve to introduce one of RCA's newest divisions—Electronic Industrial Engineering of Hollywood, California—for many years, a respected supplier of two-way cable television systems and equipment.

— J.C.P.

Future issues

The next issue of the *RCA Engineer* will feature Command, Control, and Communications systems and equipment. Some of the topics to be covered are:

Remotely piloted vehicles
Human factors
Computers for command and control
Performance monitoring
AEGIS command and control system
Hardware/software tradeoffs
System simulation

Discussions of the following themes are planned for future issues:

RCA Ltd., Canada
International activities
Consumer electronics
Automatic testing
Parts and Accessories
SelectaVision
Videovoice
Advanced communications
ElectroOptics

Two-way data communication for ancillary services on cable tv systems

B.J. Lechner | C. M. Wine

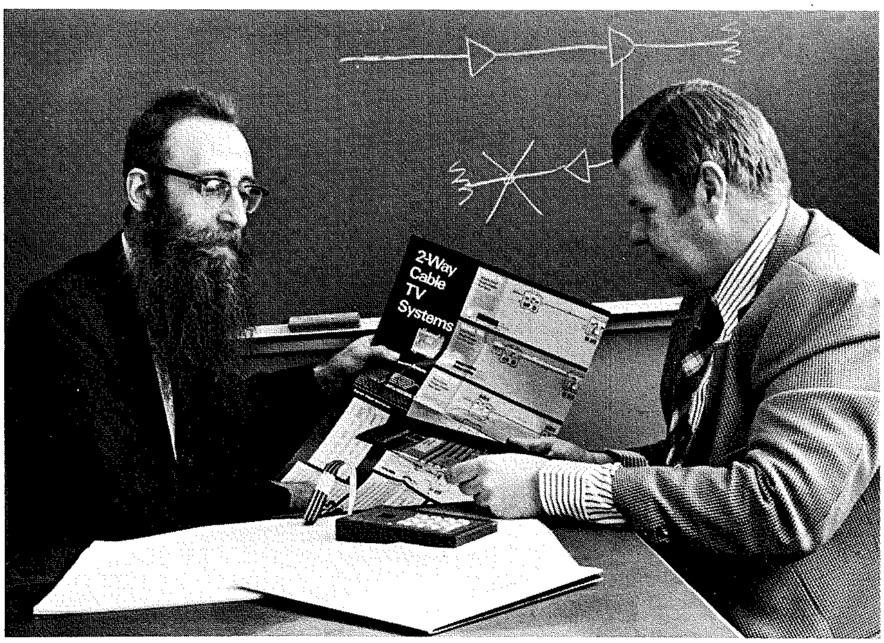
A system has been developed and demonstrated to provide the data communication required for ancillary services on two-way cable tv systems. The system will work on a variety of possible cable tv distribution systems, requires only modest terminal hardware at the home-end, can accommodate both simple and sophisticated services efficiently, and retains the flexibility to add additional services in modular fashion.

MANY NEW SERVICES, ancillary to the distribution of entertainment tv, have been suggested for modern cable tv systems having two-way transmission capability. These services include narrowcasting, home security system monitoring, interactive shopping, opinion polling, interactive education, remote banking, and electronic mail. All such services will greatly enhance the scope of communications available to the consumer in his home. To implement these new services, a two-way data communications facility linking each subscriber with the system headend and/or other central points, *e.g.* schools, banks, *etc.*, must be added to the basic television signal transmission capability of the cable tv system.

Two-way cable tv system

The layout of a typical modern two-way cable tv system is shown in Fig. 1. The headend provides a source of tv signals, including those picked up "off-air" with antennas and those locally generated with tape or film equipment. In some cases there may be a studio equipped with live cameras for local origination. These tv signals are distributed over the coaxial cable on vhf carriers. The cable tv system is a tree-like structure with a trunk and many distribution arms. Repeater amplifiers are interspersed periodically and taps on the distribution arms serve approximately 10,000 subscribers. Each distribution arm will have a maximum cable run of about 10 miles from headend to furthest subscriber.

Authors Wine (left) and Lechner.



The Engineer and the Corporation

Charles M. Wine, Communications Research Laboratory, RCA Laboratories, Princeton, N.J., received the BEE degree, cum laude, from the City College of New York in 1959 and joined the technical staff at RCA Laboratories. He has done some graduate work at Princeton University. At RCA Laboratories he has been involved in work in the following fields: psycho-acoustic subjective testing, the application of tunnel diodes to home instruments, ultra-high-speed tunnel diode logic and memory circuits and systems, high-speed cryogenic logic elements and memory systems, read-only memories, character generators, and low cost displays for time sharing systems. Most recently he has been working on both hardware and software aspects of computer peripheral equipment and systems. Mr. Wine is the holder or co-holder of seventeen issued U.S. patents. He is a senior member of IEEE, a member of AAAS, ACM, and also a member of Eta Kappa Nu and Tau Beta Pi.

Bernard J. Lechner, Head, Community Information Systems Research, Communications Research Laboratory, RCA Laboratories, Princeton, N.J., received the BSEE from Columbia University in 1957 and has done graduate work in Electrical Engineering at Princeton University. While with the U.S. Army he served as an Instructor in Electronics. In June of 1957, he joined RCA Laboratories as a Member of the Technical Staff. At RCA Laboratories Mr. Lechner has done research on video tape recorders and on kilomegacycle computer circuits. In 1962 he received an RCA Laboratories Achievement Award, and was a recipient of the 1962 David Sarnoff Team Award in Science for his contributions to high-speed computer circuitry. In 1962 he was appointed head of a group engaged in research on digitally controlled visual displays. He received an RCA Laboratories Achievement Award in 1967 for his work in the display field. From 1966 until late 1971 Mr. Lechner was Head of the Peripheral Equipment Research Group of the Data Processing Applied Research Laboratory at RCA Laboratories and conducted research on electroluminescent displays, liquid crystal displays, color displays, computer terminal displays, storage tube displays, keyboards, high-speed printers, and other computer peripherals. He was appointed to his present position in 1971. Mr. Lechner holds seven US patents and has applications pending for several more. He has numerous publications relating to his work and received a "Best Paper Award" at the 1965 International Solid-State Circuits Conference. Mr. Lechner is a Fellow of the SID, a Senior Member of IEEE, and a member of Tau Beta Pi, Eta Kappa Nu, Sigma Xi, and ARRL. He is past Chairman of the Princeton Section, IEEE; and served as a member of the Honorary Editorial Advisory Board of *Solid-State Electronics.* He is presently a member of the advisory commission for electrical engineering at Mercer County Community College. Mr. Lechner was Chairman of the Program Committee for the 11th Symposium of the Society for Information Display, and was the General Chairman for the 12th Symposium in 1971. Mr. Lechner was the first recipient of the Frances Rice Darne Memorial Award in 1971 for his outstanding contributions to the display field. In 1972 he was elected to the SID Board of Directors and in 1973 was elected Treasurer of SID.

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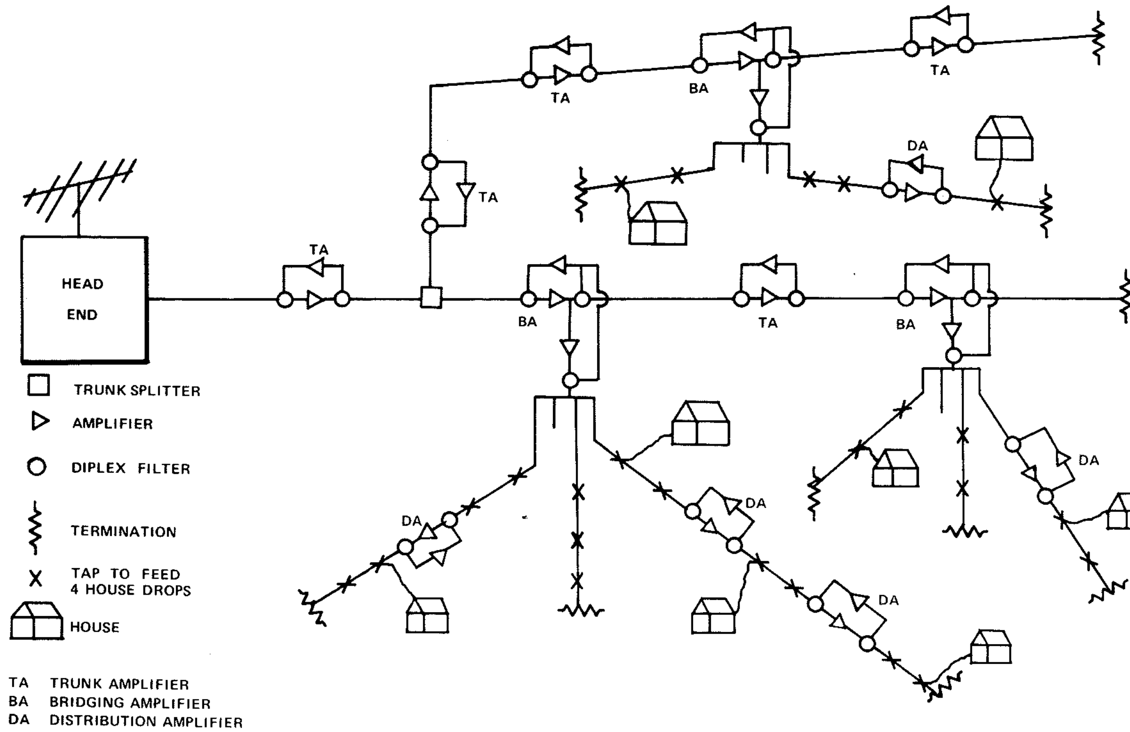

Fig. 1 — A typical two-way cable tv system configuration.

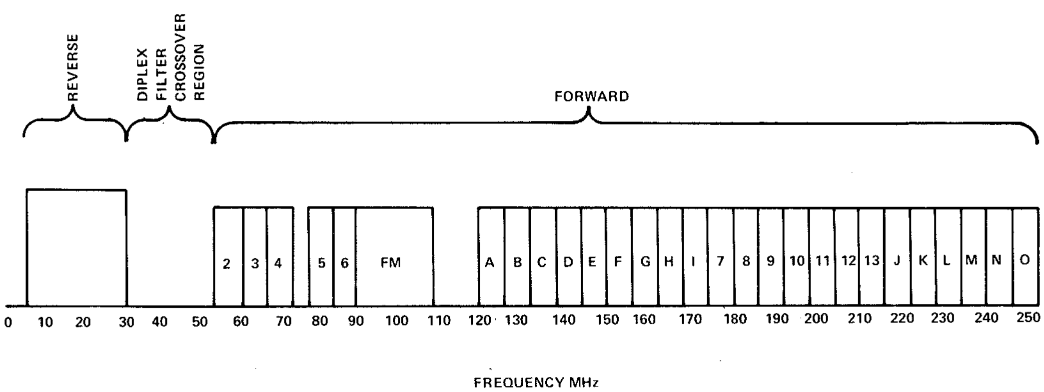

Fig. 2 — A typical spectrum partition for a two-way cable tv system.

By placing diplex filters at input and output of the amplifiers and including reverse amplifiers, the spectrum may be split so that signals may be sent in both directions on the cable. A typical partition of the spectrum for two-way operation is shown in Fig. 2. From 50 to 250 MHz the cable carries downstream signals. These include the standard vhf tv channels, 2-13, and, on many systems, the fm band from 88-108 MHz. The mid-band between 108 and 174 MHz and the super-band above 216 MHz may be used to carry additional tv signals. A typical assignment placing 9 channels (A-I) between 120 and 174 MHz and 6 channels (J-O) between 216 and 252 MHz is shown in Fig. 2.

With the diplex filters crossing over at about 40 MHz, the spectrum below 30 MHz is available for reverse communication. The spectrum below 5 MHz is customarily not used for communication since 60-Hz ac power is usually distributed on the cable to power the amplifiers and since the taps and other passives are not ordinarily designed to pass signals below 5 MHz. The reverse bandwidth is capable of carrying up to four tv channels and is used in some systems to transmit signals to the head-end from points located along the cable system. This may allow local origination of programs, such as sports, or town meetings. They are then up-converted to a forward channel and distributed to the subscribers.

It should be noted that the particular single-cable sub-band split system described above is merely illustrative. The data communications methods described here can be applied to virtually any cable system configuration including mid-band split systems, dual-cable systems, etc. All that is necessary is that the system be two way.

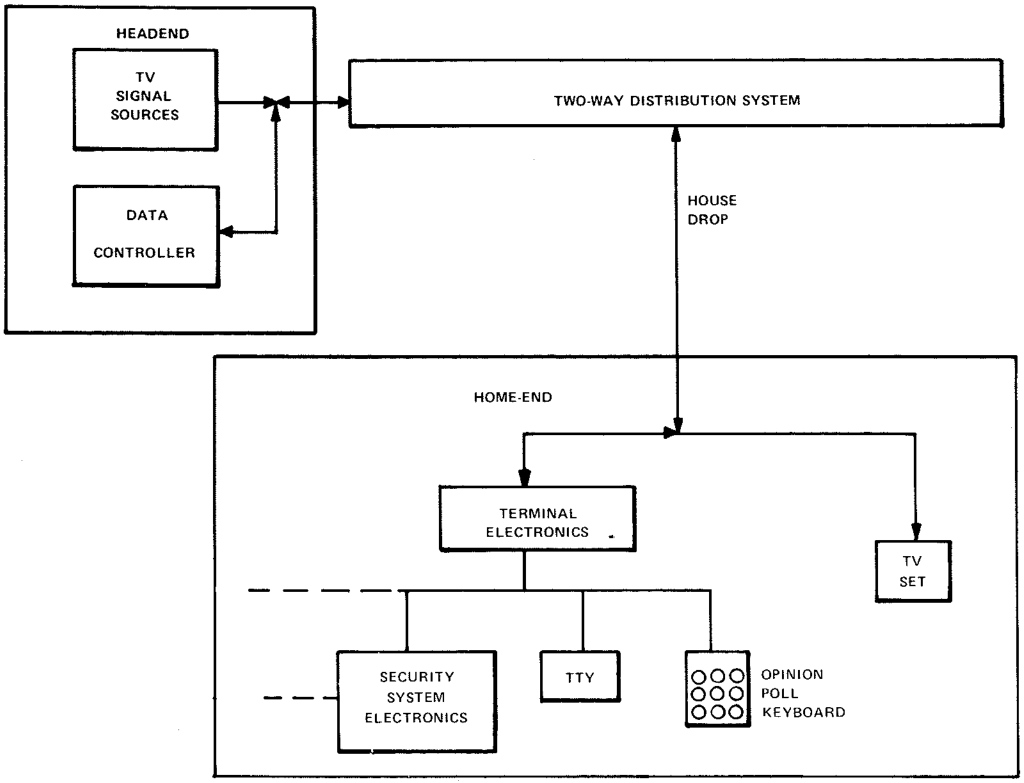

Fig. 3 — Two-way cable system with terminal and data controller.

Data communications

By reserving a portion of the reverse bandwidth, *e.g.* from 10 to 16 MHz and a portion of the forward bandwidth, *e.g.* from 110 to 116 MHz for data transmission, the two-way data communications link needed for the ancillary services can be implemented. Given the data communications capability, appropriate terminal electronics may now be added at the home-end and a data communications controller at the headend as shown in Fig. 3. The terminal electronics permit a variety of devices to be interfaced to the system at the home-end, as required by the services implemented. The headend data controller may be a simple hardware box or a minicomputer depending on the requirements of the services provided.

There are many ways in which the data communications may be accomplished but considering the requirements of the anticipated services and the need to keep the home-end hardware as simple and inexpensive as possible, a polled time-division-multiplex system is the most logical choice. A specific polled time-division-multiplex communications discipline, especially well-suited to the requirements imposed by the new services for cable tv, has been developed. The system combines efficiency and flexibility in a modular fashion and permits simple services requiring only a low-data-rate to coexist on the same communications link with more sophisticated high-data-rate services. Central control is accomplished with an inexpensive hardwired controller for simple services or with a minicomputer in the case of more sophisticated services.

The system is based on polling the individual subscriber terminals in appropriate sequence from the headend, thus maintaining control of all communications at the headend. Downstream messages consist of a string of ASCII bytes sent asynchronously at a data rate of 1.25 Mb/s. Each byte consists of a start bit, 7 data bits, a parity bit, and a stop bit, as shown in Fig. 4. At a data rate of 1.25 Mb/s each byte has a duration of 8 μs.

The downstream message has the following format (refer to Fig. 5): SYN (the ASCII asynchronous idle character: 0110100), TID1, TID2, DID, an arbitrary number of data bytes, SYN. The initial and terminal SYN characters indicate to the terminal the start and end of the message. TID1 and TID2 identify the particular terminal being addressed; this allows approximately 16,000 terminals on the system. DID identifies which one of 127 possible service-related devices is being specified, *e.g.* the home security system, an opinion poll keyboard, a teletype terminal, or a credit-card reader, etc.

The data bytes, if required, are then routed to that device. Upstream messages also consist of a string of ASCII bytes sent asynchronously at 1.25 Mb/s. The upstream transmission commences immediately after TID2 (as soon as the particular terminal has been identified) and has the following format: STATUS, DR, a number of data bytes equal to the number received. The scheme is that one byte is sent for each byte received following TID2 up to and including the final SYN. The STATUS byte reports terminal status and provides a means for initiating requests to the headend for various kinds of service. DR is the response byte from the particular device that has been addressed. The short gap between TID2 and DID is provided to allow the terminal transmitter to come on and stabilize before modulation commences.

The basic message exchange explicitly

showing the effects of propagation delay on the cable is illustrated in Fig. 6. In the simplest implementation, to insure that return messages never overlap, the head-end waits for a reply to a transmission before initiating the next transmission or times out (after waiting for a time equal to the worst-case round-trip delay) so that the system will not hang up in the event a terminal fails to respond due to a transmission error or hardware failure.

For a simple service, *e.g.* a home security system, requiring only that a device response byte be retrieved from the terminal, the downstream message is SYN, TID1, TID2, DID, SYN and the upstream message is STATUS, DR. At 1.25 Mb/s, the message time is 40 μs. Assuming a cable 10 miles long, the average round-trip propagation delay will be ~60 μs. This indicates that, on average, 100 μs are required to address one subscriber's home security system. Thus 10,000 subscribers can be polled in 1 second.

For a more sophisticated service, *e.g.* a teletype, the downstream message is SYN, TID1, TID2, DID, DATA, SYN and the upstream message is STATUS, DR, DATA. One data character is thus delivered to the teletype printer and one character is retrieved from the keyboard. This takes 110 μs on average, and to keep up with the 10-character/s rate of a teletype it must be polled 10 times/s. This requires 1100 μs/s, or 0.11% of the total communication capacity. Thus approximately 900 teletypes would fully load the system. Buffered terminals can, of course, accommodate longer messages and need be addressed less frequently, increasing the overall communications efficiency. In case the message lengths upstream and downstream are inherently unequal, *e.g.* with a card reader, dummy bytes are sent to equalize the lengths.

In a typical application there will be a mix of high and low data rate services and appropriate priority scheduling will be accomplished at the headend in hardware, or if a computer is used, in software. Note that an increase in data rate would not gain much because of the 60-μs average round-trip delay. In fact, a data rate that makes a typical message time approximately equal to the average round-trip delay is a near-optimum choice. However, use of an optimum polling sequence to permit overlapping transmissions without interference is possible and will improve communications efficiency, and then some increase in data rate would be more effective in further improving the capability of the system. However, a higher data rate will result in more expensive terminal hardware.

Although the system conceptually segments the services and treats each service as a separate device with its own DID, it is of course possible to combine services and define, for example, basic service "A" as the home security system and the opinion poll keyboard. Such combination services make sense if a majority of subscribers have the com-

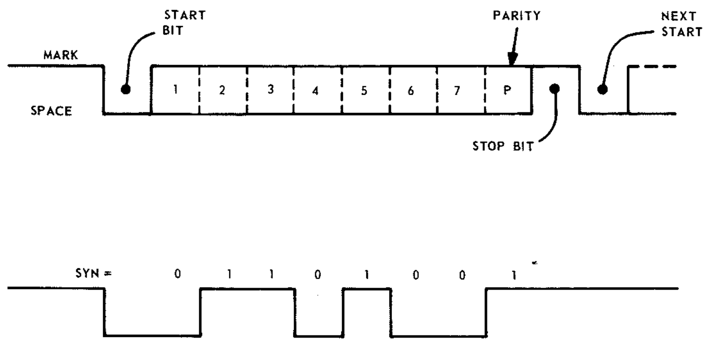

Fig. 4 — Asynchronous byte format.

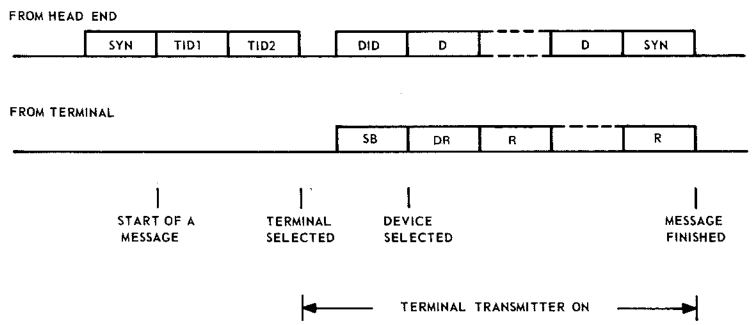

Fig. 5 — Basic communication procedure.

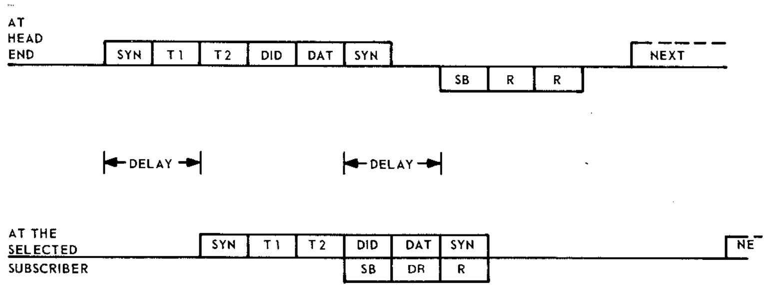

Fig. 6 — Basic message exchange

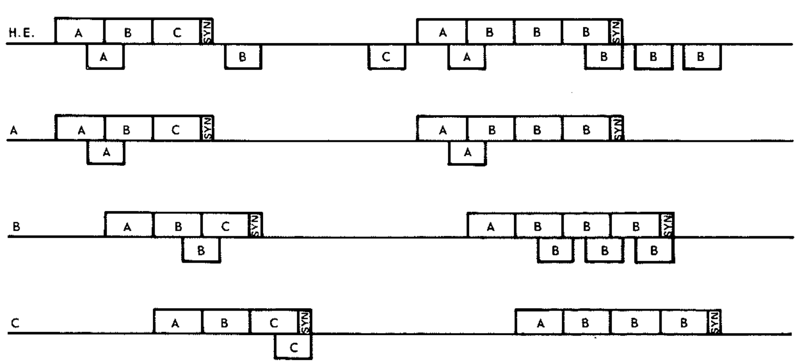

Fig. 7 — Overlapped and blocked message transmission.

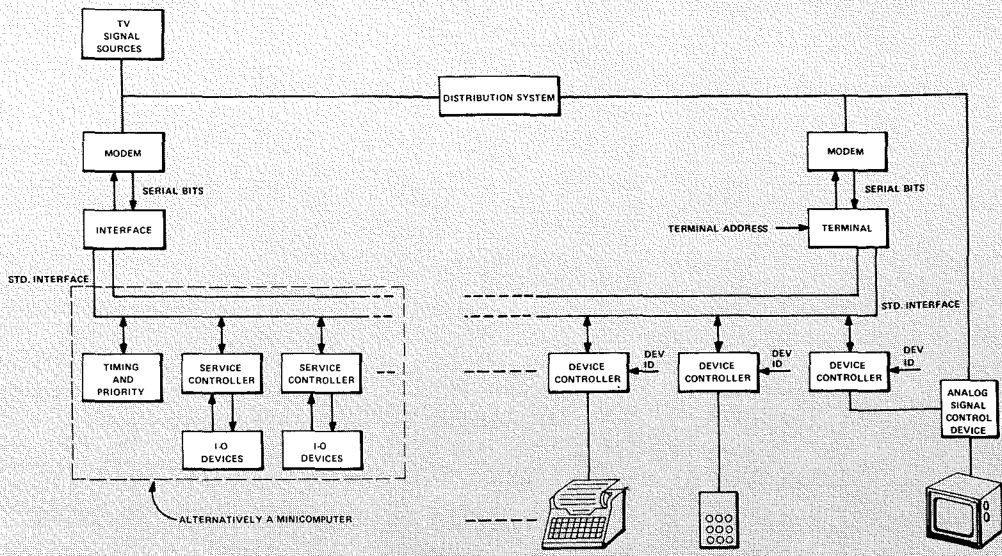

Fig. 8 — The experimental system.

bination and if the combined services require polling at approximately the same rate. It would not make sense, for example, to combine security and teletype service.

A means of improving communications efficiency by blocking and overlapping messages is shown in Fig. 7. Given three terminals A, B, and C at different distances from the headend, the headend need not wait for the reply from A before transmitting to B, if B is further from the headend than A. Thus transmissions can be sent back-to-back when the terminals being addressed are ordered in the polling list such that each terminal in the list is at least as far from the headend as the preceding terminal in the list. Also if we have several messages for the same terminal, they may be sent back-to-back without danger of the replies overlapping. When operating in this mode, we may use a single SYN between messages to serve as both the end-of-message character for one message and the start-of-message character for the next message, sending a final SYN after the last message in the block.

Experimental system

An experimental version of the system has been implemented at RCA Laboratories. It operates on a test-bed vhf cable tv system using EiE two-way cable tv amplifier and headend hardware (Fig. 8 is a diagram of the experimental system). Data is sent using FSK modems at 112-114 MHz downstream and 12.3-13.8 MHz upstream. The modem data interface is a conventional bit-serial logic-level interface and is independent of carrier frequency and modulation method. Thus the same data communications approach could be used in virtually any cable tv environment including an hf switched cable tv system. The basic headend and terminal hardware includes a byte-serial interface permitting the various service-related devices to "plug-in" independently at the subscriber terminal and their controllers (hardware or software) to similarly "plug-in" at the headend. At the headend the basic system includes timing generation and the priority scheduler; at the terminal end the basic hardware includes circuitry for timing recovery and circuitry to recognize SYN, TID1, and TID2. The device controllers at the terminal end include circuitry to recognize DID and to accommodate the input-output requirements of the particular device.

Broadcast services, such as billboard information for an alphanumeric display, can also be accommodated within the system. Since all received data is available at the byte serial interface in the terminal, a particular TIDl, TID2 can be reserved for a broadcast service and be recognized by a "broadcast receiver" which plugs in just like any other device. In this case there is no return message, of course.

As shown in Fig. 8, some services, such as narrowcasting, require interaction betweeen the digital terminal and the analog signal path to the subscriber's tv set. For example, a filter or other device may be used to block a particular channel from the tv set unless an appropriate command is sent via the data communications system to bypass the filter for a subscriber authorized to receive the narrowcast program. In similar fashion, signals may be directed to a frame-freezer used by a subscriber in conjunction with his tv set to obtain still-frame pictures from a central information file at the headend. The digital signal would inform the frame-freezer exactly which frame to store for that particular subscriber.

The experimental system at the Laboratories has been operated both with a hardware controller at the headend and more recently with a minicomputer. The services implemented include, among others, narrowcasting, opinion polling, full-duplex teletype-to-teletype communication, remote control of a tv camera, and transmission of billboard information to an alphanumeric display.

Acknowledgment

The authors wish to acknowledge the contributions of their many colleagues in the Community Information Systems research group at RCA Laboratories to the development of this system and the implementation of the experimental test system. We also wish to thank our colleagues at EiE for their cooperation.

Growth of consumer information systems

E.S. Rogers

There are many prognostications on the future of CATV and its application to Consumer Information Systems. This paper describes the present state of the technology, outlines some of the system limitations, and describes some of the future systems that could be used to provide two-way broadband communication links.

THE WIRING of our cities with reliable broadband communication networks has lông been an elusive dream. Although the transmission of broadband signals has been technically feasible for many years, the cost of the service has been prohibitive. For example, picturephone service, which was talked about more than two generations ago and was actually in operation on a limited scale in Germany in the late 1930's, is still economically beyond the reach of the average consumer. However, it appears now that broadband systems for consumers are on the threshold of a new era. Today, a number of forward looking businessmen are examining broadband communiation systems that can provide many new services for industrial, educational, and home consumer markets. It is envisioned that these new systems will eventually culminate into vast two-way cable communication networks throughout the nation.

The vehicle for this new enterprise was introduced some 25 years ago in remote areas surrounding broadcast tv transmitters where the off-the-air recep-

Edward S. Rogers, Communications Research Laboratory, RCA Laboratories, Princeton, N.J. received the BA in Mathematics and Physics from Susquehanna University in 1942 and the MS in Physics from Case Institute of Technology in 1943. After two years as a Member of the Technology Staff at Columbia University, Division of War Research, NDRC, and the USN Underwater Sound Reference Laboratory, he joined RCA Laboratories as a Member of the Technical Staff in 1945. His work at RCA Laboratories has included a wide variety of assignments in research on ultrasonic devices, solid state electromechanical transducers, noise reduction systems, room acoustics and underwater sound devices. Several years were spent on work oriented toward man-machine interaction with digital computers involving machine recognition and synthesis of speech. More recently his work has tended toward system analysis and design. For the past two years he has been investigating the system aspects of broadband information systems covering two-way CATV systems. He has published several technical papers on his research and has five issued patents. He has received RCA Achievement Awards for outstanding work in 1949, 1961, and 1968. He is a member of the IEEE, the Acoustical Society of America, and Sigma Xi.

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tion was marginal or very poor. TV reception was greatly improved through the use of a single, well-located community antenna and a cable system that could distribute the vhf signals received at the antenna site to a group of subscribers. These early systems, designated by the acronym CATV (community antenna television) were technically and economically successful because of the public demand for some form of entertainment television. Initially, only one or two channels were available, providing a picture of mediocre quality. This service was acceptable because it provided tv reception where there was none before.

The CATV industry has had phenomenal growth in areas of poor reception brought about by the insistent demand to provide adequate tv reception to the people who were unable to receive the signals directly. This infant industry has grown over the years to the point where approximately 10% of the tv homes in the USA are cable connected. [*source*: *TV Factbook*, 1972.] A number of systems are now providing 12 or more channels and, in some cases, have already added local program origination of community activities. At present, CATV systems are being designed with 20 or more channels and two-way communication facilities to provide services to subscribers who now have adequate reception of three or more off-the-air tv channels.

This possibility of many more channels and two-way capability has sparked the imagination of the communication industry. Cable television has the potential of providing vast and diverse channels of communication for all kinds of new services at a price the average homeowner can afford. Of course, the mere fact that many of these new services are technically possible does not mean that the public is willing to pay for all of them now. But it is anticipated that CATV systems will grow, in some modular fashion, into large interconnected networks with desirable services being added as they become economically feasible. It is envisioned that many new services, initially introduced as luxuries, will eventually become necessities. Because of the magnitude of the capital outlays required to provide signal distribution to millions of subscribers in densely populated areas, it is essential that careful planning be undertaken in the design of future systems to ensure that they will be able to handle increased services without becoming obsolete or impossible to maintain.

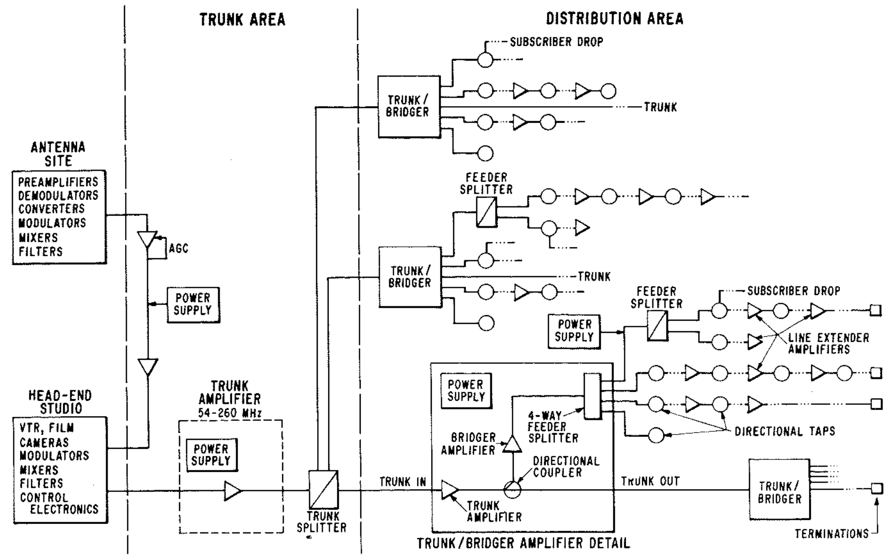

Fig. 1 — Typical contemporary single-cable one-way CATV system.

The problems of growth of the new industry are not entirely technical. As is well known, the social, political, and economic forces that have shaped the industry thus far will continue to play a major role in determining the form of the total system. Perhaps the most important factor will be economic in nature: that of providing the future services in the top 100 market areas of the United States, which will create the same kind of insistent demand that initiated CATV. These market areas have been defined by their percentage of television households and represent the major metropolitan areas of the country. The cost tradeoffs in the design of these new communication systems are tightly bound to the technical specifications and services provided by the systems.

Present state of the art of CATV in the US

The average CATV system today has fewer than 2500 subscribers and provides less than 12 channels of entertainment tv. Most existing systems serve rural areas which account for less than 15% of the potential subscribers. Some urban systems, such as Manhattan and San Diego, exist because of special situations. In Manhattan, the direct signal is poor because of multiple reflections from tall buildings. In San Diego, a natural line-of-sight barrier between San Diego and Los Angeles has made a cable system marketable.

The existing systems and most of those now being designed deliver signals at vhf to commercial tv receivers that can select the 12 vhf channels. The distribution of uhf channels on cable systems is not practical because of the high attenuation of these frequencies on coaxial cables. The choice of a vhf/fdm (frequency division mulitplex) signal was natural on early systems because it fitted directly with available tv receivers.

The quality of equipment used on many of the older systems is not high and generally the picture quality is only fair. However, the signals received at many antenna sites are weak and show multipath effects so that a slight deterioration of the picture quality due to system defects often goes unnoticed.

A typical contemporary single-cable one-way CATV system is illustrated in Fig. 1. It consists of a headend and/or antenna site and a cable distribution system. The antenna site can be a remote unattended facility comprising antenna arrays and suitable electronic amplifiers and converters to process the incoming signals to the desired frequency and amplitude for the distribution system. It contains vhf and uhf television antennas and radio antennas, and in more advanced systems it may contain microwave terminals. The headend is the control center of the system. In small systems, it is normally located at the antenna site; in larger systems, it may be remote from the antenna site and include a studio for local program origination.

Signals from the headend or studio are

carried to the subscriber's home by a cable distribution system that consists of a network of trunk and feeder lines. [The current terminology for the cable network varies considerably with the industry; in some cases the term *distribution system* is used to pertain only to the feeder lines. In this paper, the term *distribution system* is used to define the entire cable network.] Fig. 1 portrays a medium-size system containing a separate antenna site connected to a studio by a trunk cable. The trunk system is comprised of a main trunk and secondary or subtrunk lines.The diameter of the cables is a design parameter determined by system size and channel capacity. The outer conductor diameter of the main trunk cables is typically 0.750 inch and that of subtrunks is 0.500 or 0.412 inch. Feeder lines used to couple between trunk lines and the directional taps for subscribers are also normally coaxial cables with 0.412 inch diameter.

The connection between the directional tap and the subscriber's terminal, frequently called a "subscriber drop", is made with RG-59U coaxial cable. The trunk and feeder cables usually have a copper-clad-aluminum center conductor and a solid-aluminum outer conductor. Typical values of cable attenuation in dB/100 ft for channel 13 are shown in Table I. Cable attenuation is directly proportional to the square root of frequency with the result that the loss at channel 13 is about 6 dB greater than for channel 2.

Cable loss in the system is compensated for at regular intervals by amplification. There are basically three types of amplifiers used in the distribution system: trunk amplifiers, bridger amplifiers, and line-extender amplifiers. In present practice, the signals on the trunk line are carried at relatively low levels to minimize nonlinear distortion. The bridger and line extender amplifiers are operated at higher levels to provide adequate signal levels at the terminals of the subscriber after passing through decoupling networks. The frequency dependence of cable loss is compensated with equalizing networks at the amplifier stations throughout the network.

Table I — Typical values of cable attenuation in dB/100 feet. [f = 216 MHz (Channel 13); T = 70°F.]

Cable function	*Outer conductor diameter (inches)*	*Maximum loss/100 feet*	
		Poly Foam (dB)	*Super Foam* (dB)
Trunk	0.750	0.96	0.72
Trunk/feeder	0.500	1.35	1.07
Trunk/feeder	0.412	1.65	1.35
Subscriber drop	RG-59	5.20	4.20

The trunk amplifiers serve only to maintain the signal level on the trunk lines. There are no subscriber drops attached directly to the trunk line. Amplifiers are spaced at intervals to restore about 20 dB gain in the cable system. Bridger amplifiers are used to interface the feeder system to the trunk line. These amplifiers may be included in the trunk amplifier housing in which case they are called trunk/bridgers. In other situations, it may be desirable to locate bridgers between trunk amplifiers. These are called mid-span bridgers. The bridger serves to increase the signal level from the trunk line to the level required for the feeder cables. If the feeder lines from a bridger are long or if they supply signals to an area where the density of subscribers is high, it may be necessary to provide amplification along the feeder line by line extender amplifiers. To keep intermodulation distortion within limits, no more than two line extenders are used in cascade following a bridger. Gain stabilization in the trunk network is accomplished through the use of AGC circuits. AGC circuitry is added at least at every third amplifier, but more frequent control is highly desirable in large systems. Each amplifier has its own dc power supply. The ac power (30 or 60 V rms) is supplied via the coaxial cables; it is inserted into the coaxial cables at various points within the system.

A number of passive devices make up the remainder of the distribution network components. They consist of line splitters in the trunking system, and subscriber taps in the feeder lines. Each tap contains a decoupling device to prevent interfering signals from entering the distribution system, and a splitter to provide outputs to several subscribers.

If more than 12 channels are carried on the system, a converter will be required at each subscriber location to receive the additional channels. An alternate contemporary system utilizes dual cables and can provide up to 24 channels without the use of a converter. In these cases, an *A-B* switch is required to select one of the two cables.

Systems limitations

The technical problems encountered in distributing many tv channels on a single cable over wide areas point up several system limitations. In a communications system involving the cascading of a series of amplifiers, signal degradation occurs at each component throughout the system. As CATV systems enter the top 100 metropolitan markets, which presently enjoy good over-the-air reception of tv broadcast signals, CATV systems must be able to provide adequate picture quality to attract subscribers. This will require the setting of standards to define picture quality on a quantitative basis. Primary factors affecting the design of cable systems are:

- Signal-to-noise ratio
- Response frequency characteristic over a channel
- Nonlinear distortion
  - Crossmodulation
  - Intermodulation
- Envelope delay distortion
- Reflections usually termed echo rating or return loss

The noise and nonlinear distortions originate in the amplifiers. Amplifier noise, normally white Gaussian (flat within a 6-MHz channel) accumulates in the system on a power basis. In a chain of identical amplifiers, for instance, the noise power would increase by 3 dB every time the number of amplifiers in the cascade is doubled. As shown in Table II, a typical amplifier has a noise figure of 11 dB. Consequently, a cascade of 20 amplifiers will cause an increase in noise level of about 13 dB which sets the minimum signal level of the system. However, as the signal level is increased, crossmodulation and intermodulation distortion caused by nonlinearities in the amplifier will limit the quality of the picture. Generally, the effects of crossmodulation and intermodulation are different and, thus, are considered separately by the CATV engineers. Crossmodulation produces interfering pictures that drift through the desired picture. Intermodulation produces beats and generates a "herringbone" pattern in the picture.

Crossmodulation and intermodulation products in a chain of amplifiers accumulate on a vectorial basis. Their amplitude depends on the output level of the amplifier. For a so-called well-behaved amplifier in which third-order

Table II — Typical operating conditions of modern vhf/fdm CATV amplifiers for a 12-channel system.

Performance per amplifier	*Trunk amp.*	*Bridger amp.*	*Line-ext. amp.*
Gain (dB)	20	38	25
Input level (dBmV)[1]	+ 10	—	+18
Output level (dBmV)	+ 30	+43	+43
Noise figure (dB)	11	11	14
CrossMod (dB)[2]	− 89	−65	−57
Intermodulation dB			
2nd order	− 84	−77	—
3rd order	−104	−82	—

[1]CATV frequently uses the term dBmV.
0 dBm = 1mW
0 dBmV = 1mV across a 75-ohm load
Thus, −35 dBm = +14 dBmV
[2]For definition, refer Ken Simons, *Technical Handbook for CATV Systems* (Jerrold Electronics Corp., Hatboro, Pa.).

distortion predominates, an increase of x dB of the output level results in an increase of $2x$ dB of the third-order intermodulation and crossmodulation products. The number of nonlinear distortion products is also a function of the number of channels transmitted. With only a few channels and few distortion products, a single strong intermodulation component within a tv channel appears as easily visible beats in the picture. As the number of channels transmitted by the CATV system increases, the number of intermodulation products increases very rapidly. In systems carrying many tv channels, therefore, one finds many intermodulation products within the 6-MHz band of a channel, but their combined effect closely resembles the effect of random noise. The result of the accumulation of noise and nonlinear distortion in amplifier cascades reduces the useful dynamic range of the system by approximately 6 dB for each doubling of the cascade length.

Envelope-delay distortion is usually caused by filters associated with amplifiers. This is particularly a problem with two-way systems using diplexer filters to separate the signals in each direction. Envelope-delay distortion also accumulates as the length of the cascade increases. Nonuniform envelope delay within a tv channel can manifest itself as a misregistration of the color information relative to the luminance information. This gives the impression the colors are misprinted, a condition termed the *funny paper* effect. Nonuniform envelope delay can also result in poor transient response.

Echoes can occur at the input or output of any active or passive device not perfectly matched to the connecting cables. Echoes can also originate within the cable itself due to impedance variations of low-quality cable or from mechanical defects. The subjective effect of echoes depends very much on the echo delay. Delays of one microsecond or longer result in clearly visible second images. Shorter delays cause loss of picture resolution and "fuzziness" in picture detail. Additional factors that affect the quality of the tv picture received over a cable system are adjacent channel interference and direct off-the-air reception of co-channel pickup due to strong local stations. These are problems related to the design of the tv receivers.

In addition to the number of distortion products produced, there are a number of problems arising as the number of channels is increased. The introduction of additional channels increases the signal level that each amplifier must carry. As the number of channels is increased, the total power carried increases on a power basis. Increasing the number of channels from one to thirty-two would require an increase of 15 dB in power-handling capacity. For a given amplifier, this has the effect of reducing the length of the cascade.

The present assignment of broadcast tv channels is arranged so that adjacent channels are not available in the same metropolitan area. Therefore, tv receivers are not required to separate adjacent channels from the same antenna. When 12 channels are carried on a single cable, many existing receivers do not provide sufficient adjacent-channel filtering to adequately prevent adjacent-channel interference. One solution the CATV industry used to solve this problem has been to provide converters with a narrow pass band to provide the additional filtering. However, this additional filter reduces the resolution of the picture and introduces other objectionable effects such as multiple fine tuning and thermal drift.

A number of CATV systems provide more than the standard twelve channels on a single cable. One plan for providing additional channels allocates eight or nine channels in the "midband" between standard channels six and seven, and several more channels in the "superband" above channel 13 (216 MHz). Assignment of channels in the midband requires the use of a converter at each tv set. As originally conceived, the channel allocation for broadcast tv excluded the midband to eliminate the effect of second-harmonic distortion. The use of this band will place more stringent requirements on the amplifiers in the distribution system. In addition to second harmonic distortion, the increased number of channels on the cable also increases the number of other intermodulation distortion products. The introduction of channels in the mid-band frequency region caused additional constraints on the tv receiver design. It will be necessary to provide improved image rejection and prevention of the local oscillator from radiating into the mid-band.

With the introduction of two-way systems, the accumulation of noise in the return channel becomes a problem. In a one-way system, the noise received at the subscriber terminal is contributed only by the amplifiers through which the signal passes between the headend and the subscriber. In a two-way system, however, the headend receives noise on the return channels from all return amplifiers in the entire system. A number of other types of interference can enter the return system from poorly shielded conditions at the subscriber terminals. Solutions to the noise problem are available, but as yet, none has been universally accepted by the CATV industry.

Along a cable route from the antenna site to the most distant subscriber, the effects of picture degradation caused by individual components of the cable distribution system gradually accumulate. Obviously, each component will contribute its share to the overall picture impairment. Therefore, for a given quality of components, only a finite number of devices can be cascaded before picture impairment reaches the established minimum quality standards.

In designing a CATV system, device parameters such as amplifier gains, signal levels, *etc.*, are chosen so as to minimize impairment effects. For example, because relatively high signal levels are used in the feeder part of the distribution system and the noise level is high from prior amplification in the trunk system, most of the tolerable amount of nonlinear distortion is allotted to the bridger and line-extender amplifiers. Trunk amplifiers, on the other hand, are operated with much lower input and output levels; consequently, they contribute most of the noise power permitted to accumulate in the system. With present equipment and depending upon the desired picture quality and number of channels, the maximum length of a cascade of amplifiers will be somewhere between 5 and 10 miles for a one-way cable system. The amplifier limitations of noise figure and overload characteristics are limited by the present state of the art of transistor technology. It is anticipated that improvements in transistor parameters will permit the design of amplifiers with better linearity and lower noise figures. A listing of typical operating conditions for a contemporary set of amplifiers is contained in Table II.

Over the years of CATV development in this country, a set of unofficial standards or practices has evolved for vhf/fdm systems. Many of these practices have resulted from a compromise between technical desires and economic considerations. For a number of years, the FCC has been following the development of CATV and periodically has issued rules governing the operation of CATV systems. On February 3, 1972, the FCC released the *Cable Television Report and Order*. This document establishes certain standards of minimum technical performance of signals arriving at any subscriber terminal. These standards are concerned with cable channels which deliver standard broadcast television signals picked up off the air or received by microwave networks and do not cover channels used for other home services. As cable systems develop, additional standards will be required. A number of committees are at work to aid in the development of a complete set of standards for future CATV systems.

The Future of CATV

The CATV industry has substantially met the demand to provide broadcast tv signals to communities that have poor off-the-air reception. However, this represents less than 15% of the potential tv homes. About 85% of the existing tv homes are located within the top 100 metropolitan markets of the US. These market areas are usually defined to lie within the grade-B contour of broadcast stations in these areas. How can CATV serve the top 100 market areas that typically have at least three broadcast stations and provide reasonably good off-the-air reception?

The answer is said to be through providing additional services to meet public demand. A formidable question is: "What services will be in demand?" A multitude of services, features, and functions have been suggested as desirable. Many of the proposed concepts will not be in demand for decades, some will not be viable until large cable networks are available, but some services may become economically realistic in the very near future. It would not be feasible to attempt to list all of the possible services one can foresee for CATV networks. A number of categories and typical services are listed in Table III as examples of the types of services a large-scale system could provide. This list gives some of the major categories that have been suggested and specific examples of the types of services that could be offered. Audio and video functions, simultaneous or separate, could be available for each of the service categories.

A time scale for the realization of these new services has not been established. In fact, there is very little concurrence on the demand for the various services or in the order in which they will be made available. A number of experimental systems are now being implemented to test the technical feasibility of the system designs and to permit market studies of the demand for services which are technically possible. It is clear that many additional channels and two-way communication facilities will soon be available, since the FCC has decreed that all new franchises shall be able to carry at least 20 channels and have two-way capability. With this added capacity available, it would seem logical that tv programs for special entertainment and educational purposes could be offered at an early date. Some of these may incur additional charges. As the two-way systems grow, two-way business services such as data transmission or video conferencing would be possible. More local origination programs of community concerns should follow with the availability of unused channels.

Within a decade, it is conceivable that a large proportion of the top 100 markets would be wired for extended two-way service. This would introduce the practicality of security systems, remote banking with the possibility of checkless transactions, random-access image files, and special limited-access controlled channels. Further into the future, one can visualize high resolution video with large screens; many more channels on very wideband cables; very large direct-access data files; and home printing of newspapers, mail, and specially selected material.

The problem for the system designer is how to plan a modular system that can provide the quality and reliability of service that will be demanded at each stage in the evolution of this new communication business. Each part of the system—antenna, headend, distribution-system, and home terminal—affects the overall performance of the total system. The design of the distribution system for this complex network can take many forms. Some factors that must be considered are outlined in Table IV. The tradeoffs possible for the system design are dependent upon these and other technical factors. Socio-economic and political factors will also affect the acceptance of the final system design.

With the potential source for growth in the top 100 market areas, the new systems must be capable of servicing these markets. Experience has shown that market penetration will probably be small at first, and then will grow to 50% or more within 10 years. As pointed out in the previous section, the radius of 35 miles defining a market area is well beyond the maximum cascadability of existing equipment for 12 channels. The requirements of more channels and high picture quality will further reduce the maximum range of vhf/fdm systems. Therefore, with added channels and improved system reserve, it will be necessary to develop a hub-type system made up of many smaller systems that could take the form of a cable supertrunk or a microwave link.

The services offered will certainly require a large number of channels, from 20-channel capability now to 48 or more channels within the next decade. Two-way communication will be a must and it would appear desirable to have a flexible arrangement whereby the number of "upstream" and "downstream" channels could be under the control of the system manager to meet varying traffic demands. Local program origination may require facilities to pick up programs anywhere within the system which can be transmitted to all subscribers or to certain groups of subscribers with an access-control system to maintain system securi-

Table III — Typical services available in future CATV systems.

Entertainment
Additional channels (40 to 60)
Interactive programs with audience participation
Preference polling
Opinion polling

Premium tv (additional charge)
New movies
Broadway shows
Special sports events

Educational
Special education, elementary, high school, college
Interactive systems
—between student and teacher
—between student and computer
General research files
video library
image files

Commercial
Banking and credit
Access to large scale computer
Meter reading: water, gas, electric
Catalog shopping from home
Utility maintenance

Medical
Limited access channels
—teaching
—diagnosis
Research files

Security
Fire, burglar, gas, emergency
System monitoring and maintenance

Table IV — Factors affecting the design of a CATV distribution system.

1. *System size*
Potential subscriber density
Actual penetration
Maximum amplifier cascade

2. *Services offered*
Number of tv channels
Two-way communication
Local origination
Premium tv — Channels with limited or controlled access
Community information services

3. *Service quality*
Picture clarity
Reliability
Maintenance costs

4. *Equipment specifications*
Cable quality
Electronics
Terminal equipment
—tv receivers
—converters
—cable switches (multiple cable systems)
—special tv receivers

ty. Similarly, access-control and billing facilities will be required for the implementation of a successful premium tv service. The Community Information Systems (CIS) facilities will vary with the service, but two-way capability will be required as well as many new terminal devices that can satisfactorily interface with the distribution system.

Specifications of service quality have not yet been established, but it will be assumed that a noise and interference-free picture will be required to be salable in the top 100 markets. This establishment of standards will no doubt require subjective measurements to establish the criterion to satisfy a majority of the viewers. The reliability and maintenance experience for the present state of the art equipment leaves much to be desired. If future CIS features, particularly security services, are to be successful, the mean time between failures will have to be increased substantially and some sort of diversity system will be needed to provide backup in the event of component failure. If steps are not taken to improve the quality of the distribution system, maintenance of very large systems with thousands of amplifiers will be virtually impossible.

CATV distribution systems capable of providing two-way services can assume many forms some of which are in existence at the present. The signal frequency can be assigned at hf or vhf. If hf signals are used, then only a few channels, possibly four or less, can be carried on a single cable; this implies a multiple cable system. Some cables could be used for downstream signals and other cables would be used for upstream signals. No diplexer filters would be required. If vhf signals are used, then many more channels can be carried on a single cable with concomitant problems discussed. Up to 30 channels are proposed for frequency division muliplexing on a single cable. In addition, three reverse channels can be carried in the hf range using diplexers and two-way amplifiers. Another option offered at the present time is vhf/fdm dual-cable system. Some of these systems have two-way on both cables, other systems use two-way on only one cable.

Another vhf system that is being installed uses three cables: two for downstream signals and one for upstream signals. In combination, various hybrid systems have been conceived in which the trunking system operates in one mode and the feeder lines are operated in a different mode. The signal interface, switching, and translation take place in a subscriber exchange.

Table V — Types of systems proposed for future two-way services.

VHF/FDM systems
Single cable
—Sub-band Split
—Midband Split

Dual Cable
—One cable for each direction
—Two way both cables
Sub-band split
Mid-band split
—One way Cable A, two-way Cable B
Mid-band split
Multicable

HF systems
Multicable switched systems

Hybrid systems
Distribution systems
Transportation systems
—Super trunk
—Microwave
—Satellites

It would be impossible to detail all of the variations that have been proposed for future two-way CIS systems. A number of systems presently being considered are summarized in Table V. The system aspects of the various approaches are described to provide an overview of possible future configurations.

Single cable

The most obvious choice of a two-way system would be frequency division multiplexing on a single cable. This is the most commonly used technique used in existing CATV systems. The natural trend is to add more channels in the vhf region and use the frequencies below 54 MHz for return or upstream channels. This is termed a single-cable, subsplit system. A typical single trunk, single feeder system is illustrated on Fig. 2. This is a variation of the contemporary one-way system shown in Fig. 1. The cable is capable of carrying signals in both directions simultaneously; however, upstream and downstream signals must be separated by diplexer filters. Some commercial amplifiers are available with passbands up to 300 MHz, providing the possibility of 30 or more channels on a single cable. The upstream spectrum is normally limited to 5 to 35 MHz with a 19-MHz guard band between the two groups of channels. This guard band

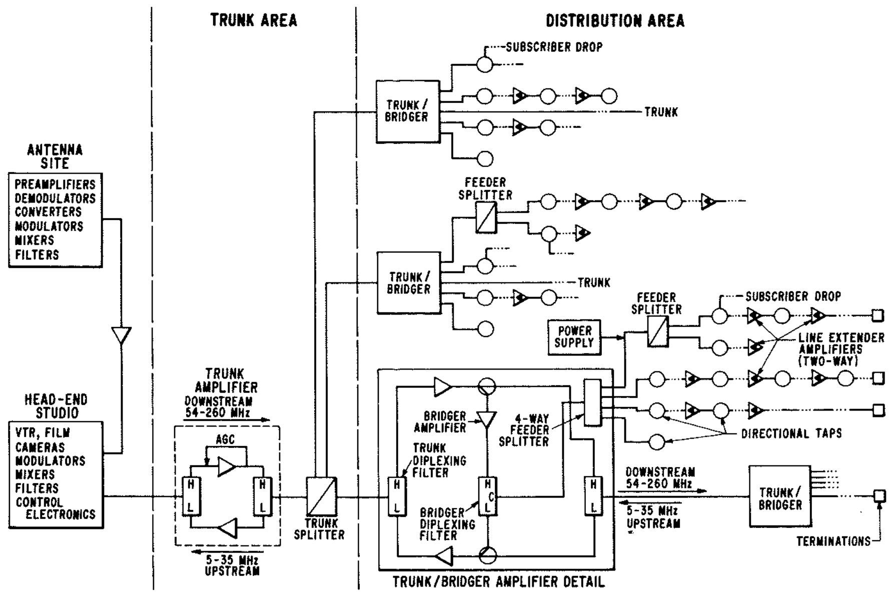

Fig. 2 — Single-cable single-feeder two-way CATV system.

relaxes the design problems of the diplexer filters with regard to passband flatness, insertion loss, return loss, and group delay distortion. The cable losses at the lower frequencies are much less than those at vhf, so fewer amplifiers are required in the upstream channels. Details of each repeater station are shown in the figure. Each repeater station requires diplexer filters regardless of whether the upstream channel is active or passive.

If more return channels are required, another version is the single-cable mid-band split. This system would probably be limited to special purposes since only vhf channels above channel seven would be available for downstream distribution.

Dual-cable system

The next most obvious step to achieve more channels both forward and reverse is the use of more cables. If more than 12 channels are sent to the present tv receivers on a single cable, a converter is required to convert the additional channels to one of standard broadcast channels. A dual-cable system can supply 24 channels that can be received on a conventional receiver without a converter. All that is required is a simple *A-B* switch at the set to select the desired cable. The *A-B* switch is much less expensive than a converter and it does not reduce resolution nor introduce secondary tuning problems. An illustration of a dual cable two-way CATV system is shown in Fig. 3.

This is essentially a parallel duplication of the single-cable two-way system. There are several options that may be used for the return channel. First, a subsplit system may be used with return channels on both cables. This requires full duplication of diplexer filters and return-channel amplifiers. Second, cable A can be used as a one-way downstream cable with up to 30 channels if desired. This eliminates the need for band splitting filters and the resulting problems in cable A. Cable B can be used as a mid-band split cable providing 14 upstream tv channels plus 19 MHz for return data. One proponent of this system uses cable A for downstream only and divides the upstream spectrum into two bands. The 5-to-35-MHz region is used by the *A*-feeder return signals which are transferred over to the *B*-cable by a high-low diplexer filter. This portion of the spectrum is allotted to the normal subscriber. The 35-to-100-MHz portion of the *B* return and the downstream *B* are used for restricted access. They may be assigned to commercial interests. This option is shown in Fig. 4.

Other variations of the dual-cable system are proposed using *A-B* switches and converters for additional channels. With the addition of more cables, a great many more variations are possible. As mentioned earlier, one Multiple System Operator (MSO) has announced the use of three cables: two downstream and one upstream.

Switched systems

As the number of cables is increased, ultimately the point is reached where each cable carries only one program. Systems of this type, called hf systems, have been in use for some years in Europe and recently were introduced in this country. Since hf systems operate at relatively low frequencies, the cable attenuation per mile is much lower and, hence these systems can cover quite large distances. At present, there are two basic forms of this system. In one case, the signals are

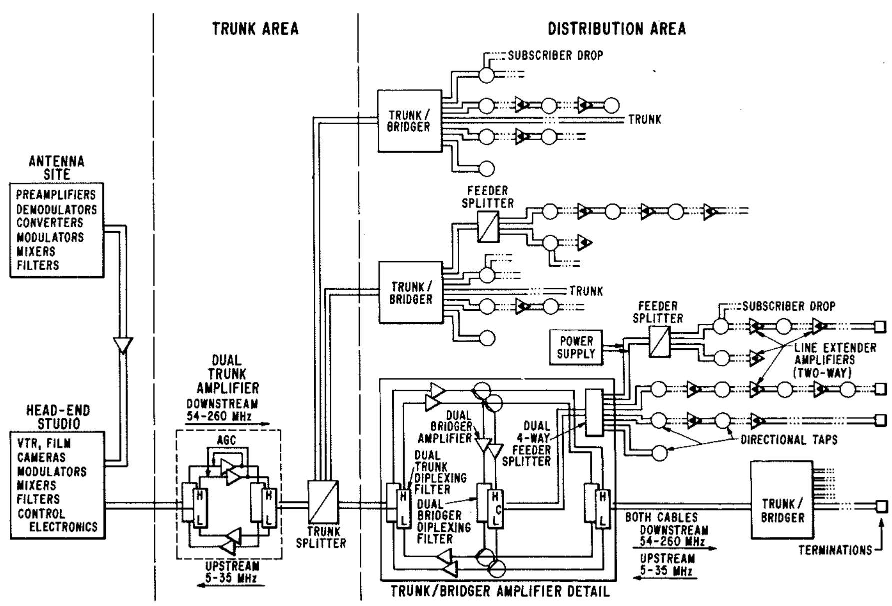

Fig. 3 — Dual-cable two-way CATV system.

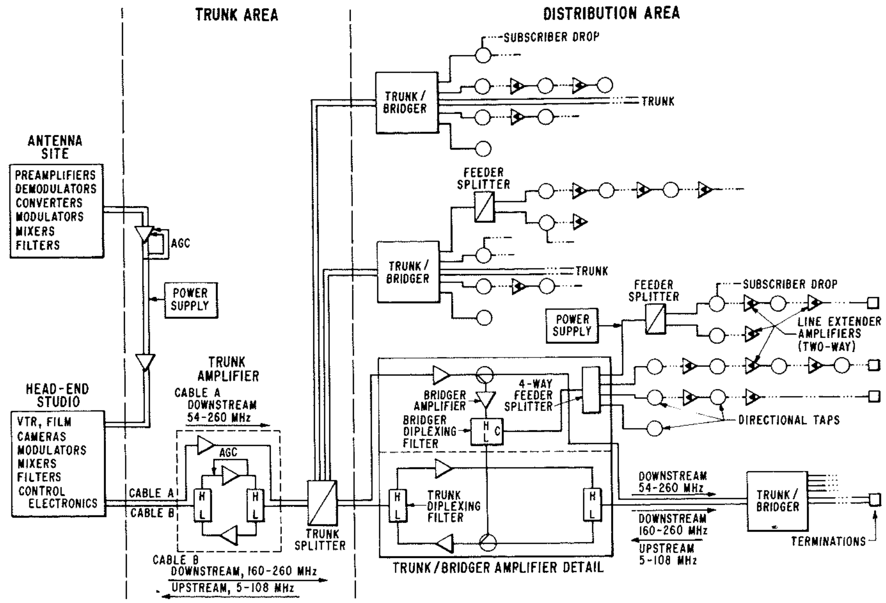

Fig. 4 — Dual-trunk single-feeder two-way CATV system.

trunked over large distances at hf and then switched to the subscriber at an i.f. frequency at a subscriber exchange. In the second case, a multi-cable trunk supplies signals to a subscriber exchange as shown in Fig. 5. Each subscriber is connected to the subscriber exchange by two wire pairs. One pair carries program material in either direction while the other cable carries control signals. At the exchange, each subscriber has a 36-position selector switch providing a maximum of 36 connections to program sources. The selector switch in the exchange is positioned by pulses from a telephone-type dial selector in the home. The dc pulses from the dial are sent to the subscriber exchange on the control cable. Program material is sent to the subscriber on the other pair of cables.

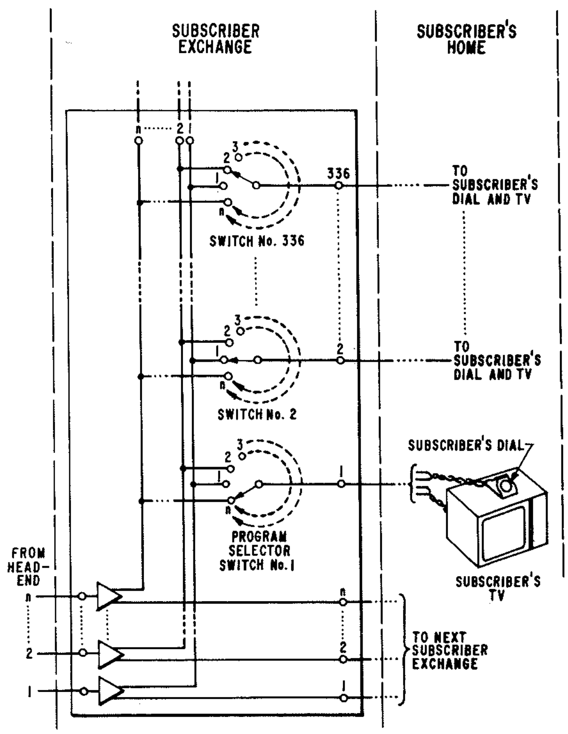

Fig. 5 — Details of subscriber exchange in an hf distribution system.

This system is extremely flexible and does not require coaxial cables. Simple hf twisted-pair wires are capable of carrying hf signals about 1800 feet from the exchange. Present designs provide for 336 subscribers at each subscriber exchange. Distortion and intermodulation is not a problem in this system since only one program is carried on a pair of wires. One drawback to this system is the number of cable pairs required from a subscriber exchange. It does appear that this method of distribution will be valuable in highly populated areas.

Hybrid systems

It is anticipated that under certain circumstances some systems will be more desirable than others. It may well be desirable to combine hf and vhf techniques into hybrid systems and use the most economical method for each problem. For example, it may become obvious that hf systems are ideally suited for large apartment complexes, but would not be economical in residential areas with private homes. By using these two types of distribution techniques where they are best suited, they can be used to complement each other.

Transportation systems

In conventional vhf/fdm CATV systems, the accumulative effects of noise and intermodulation distortion in a chain of trunk amplifiers limits in practice the number of amplifiers that can be cascaded. Depending on the desired picture quality and the number of tv channels multiplexed on the cable, the length of a trunk line is typically limited to approximately 5 to 10 miles. As the market in the top 100 metropolitan areas develops, an increasing demand will require means to transmit groups of tv channels over greater distances without undue distortion. The need, for instance, arises to extend an already existing CATV system to serve additional suburbs or communities, but the distances involved rule out vhf trunk connections. One can also envisage the need to interconnect different headends or distribution hubs to spread the cost of local program origination or premium tv over a larger number of subscribers.

One solution for the transportation of tv signals over larger distances is the use of microwave links. The frequency band

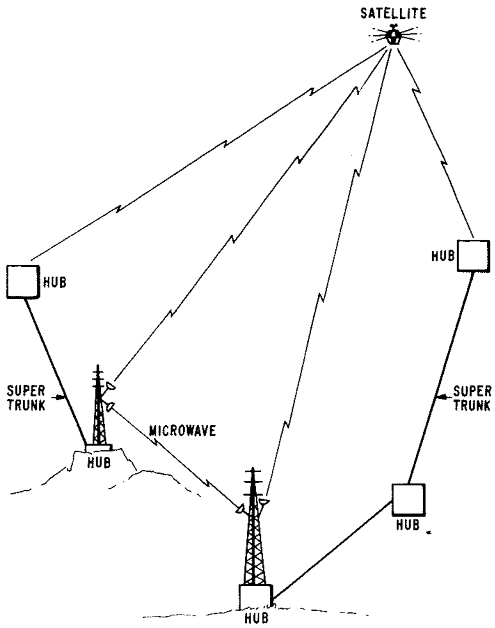

Fig. 6 — Future large-area CATV system using various transportation techniques.

12.70 to 12.95 GHz has been provided by the FCC for this Community Antenna Relay Service (CARS). An alternative solution makes use of multiple coaxial cables to transmit the tv signals at frequencies in the hf (instead of vhf) band to take advantage of the lower cable attenuation at the low frequencies. This approach will be referred to as an "hf supertrunk". A possible future large-area CATV system is illustrated in Fig. 6. This indicates the possibility of using domestic satellites as another eventual alternative.

HF trunking by coaxial cable has been pioneered since the middle or late 1960's in the US and in Europe. The early systems started by transmitting a single tv channel per coaxial cable of small (approximately 1/4 inch) diameter. Subsequently, systems were developed with two tv channels per cable. Very recently, a version of an hf supertrunk, transmitting four (and possibly up to seven) tv channels over conventional 3/4 inch CATV cables at frequencies below 50 MHz has been announced. No such system has been installed at this time.

Both microwave links and hf cable trunking are well established techniques for commercial wideband communications systems; however, it is extremely difficult at present to obtain a precise evaluation of their relative merits for CATV applications. This is particularly true with regard to the cost and reliability of multi-channel microwave links. More experience with all systems will be required to establish their value.

Conclusions

The CATV industry has grown from small rural systems distributing a few vhf channels to large businesses operating several large distribution networks. The business in the past has grown because it satisfied a need; namely, to provide adequate tv signals to subscribers who could not receive a satisfactory off-the-air picture.

The business is now on the verge of a new era. It is technically capable of providing a vast array of new services throughout the nation. This is a big step from the original purpose and objectives of the systems. The new services will require bidirectional operation and many additional channels. The form of these two-way systems has yet to be established.

A number of exploratory systems are being designed and implemented using some of the system configurations described. Feasibility tests of various prototype models will further demonstrate the advantages and disadvantages of different concepts from both technical and economic points of view. Market studies using these trial systems will be valuable in determining those services which will be in demand. Experiments with the new systems may uncover problems as yet unforeseen.

When the future markets for various CIS services have been identified, it is conceivable that there will be applications for more than one type of proposed system in a given market area. For example, in regions of high population density where many services may be economically viable, a switched system with signals distributed at hf on balanced twisted-pair may prove most effective. On the other hand, in sparsely populated regions of a metropolitan area, it may not be economically sound to provide more than a few special services. In these regions, a single-cable two-way system using vhf /fdm may be satisfactory. Whatever the form these new systems take and whatever services become in demand, it will be necessary to provide a quality product at a reasonable cost to attract subscribers. Despite the technical, financial, and marketing problems to surmount, there is little doubt that CATV will grow and prosper, bringing a new era to consumer information systems.

Acknowledgment

Many discussions and meetings with various members of the Laboratories have been valuable in forming the ideas expressed in this paper. I am particularly grateful to Dr. H. G. Schwarz for his kind and tolerant patience discussing detailed aspects of the future of cable, and to Dr. J.J. Gibson for his valuable constructive comments.

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Design aspects of bidirectional cable tv systems

R. L. Schoenbeck

Bidirectional (two-way) cable tv systems require additional design consideration over and above that required for normal downstream (forward) cable distribution systems. Cascade length based on overload and signal-to-noise ratio in the downstream portion of the cable system may prove to be unsatisfactory for high quality transmission of video signals in the up-stream (reverse) direction.

ONE ASPECT of the upstream cable system that is quite different from the performance of the downstream system is the manner in which the noise increases or builds up.

In the downstream system, noise increases 3 dB each time the number of trunk amplifiers is doubled in a given leg or branch of a cascade. In the upstream system, noise increases 3 dB each time the total number of trunk amplifiers is doubled in a section operated from a common trunk line. The prime reason for the difference in noise buildup in the upstream cable system is the behavior of power splitters used in the trunk line. In the downstream direction, the splitter provides the same loss to the noise and signal so the signal-to-noise ratio does not change. In the upstream direction, the splitter is operated as a combiner. The signal is attenuated 3 dB plus any dissipative losses in the splitter, but the noise is attenuated only by the dissipative losses in the splitter since noise is entering on two ports. The result is a 3-dB degradation in signal-to-noise ratio when two cables with equal noise are combined in a splitter (see Fig. 1).

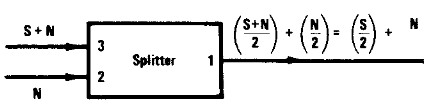

Fig. 1 — Combining characteristic of a splitter in a bidirectional system.

This noise-adding effect produces a noise level at the headend equal to a level that would be generated if all the trunk line amplifiers in the system were connected in one long cascade. The size of a single trunk line system must therefore be limited to minimize the effect of the noise adding. In practice, the maximum number of trunk amplifiers that should be operated from a single trunk cable is about 100. This number of amplifiers will produce a 46-dB signal-to-noise ratio at the headend with typical reverse amplifier noise figures and operating levels (see Fig. 2).

Limiting the number of amplifiers in a single trunk line to a maximum of 100 may seem like an extreme restriciton on a cable system; however, 100 trunk-bridger amplifier locations can provide signals to as many as 10,000 subscribers. There is no restriction to the number of these separate trunk lines that branch out from the headend.

The question then arises — how many distribution amplifiers can be used and still maintain a given signal-to-noise ratio in the upstream direction of the cable system? It can be shown with typical amplifiers that four distribution amplifiers contribute the same amount of noise in the upstream system as one trunk-bridger amplifier. That means the number of trunk-bridger amplifiers must be reduced by one (from the recommended 100 maximum) whenever four additional distribution amplifiers are added to the system. The total number of potential subscribers remains at 10,000 for a 46- dB signal-to-noise ratio as distribution amplifiers are traded for trunk-bridger amplifiers, because four distribution amplifiers serve the same number of subscribers as one trunk-bridger amplifier.

−59dBmV	Reference noise level
10 dB	Typical amplifier noise figure
−49 dBmV	Noise level at amplifier input
20 dB	Noise increase from 100 amplifiers
−29 dBmV	Noise level at headend
+17 dBmV	Typical signal level at headend

(+17 dBm) — (−29 dBmV) = 46 dB signal-to-noise ratio

Fig. 2 — Typical upstream signal-to-noise ratio for a bidirectional cable system.

A few calculations will show a correlation between signal-to-noise ratio and potential number of subscribers and vice versa. Fig. 3 shows the results of these calculations. The horizontal axis on the graph indicates the number of trunk-bridger amplifiers, used in a given area, which are fed from a common trunk line. The vertical axis indicates the number of distribution amplifiers which are fed from the bridging outputs of the above trunk-bridger amplifiers. The sloping dashed line on the left side of the graph represents a recommended limit of two distribution amplifiers on each bridger output. The limit of two distribution amplifiers is based on downstream system performance, but a restriction in either direction affects total system design.

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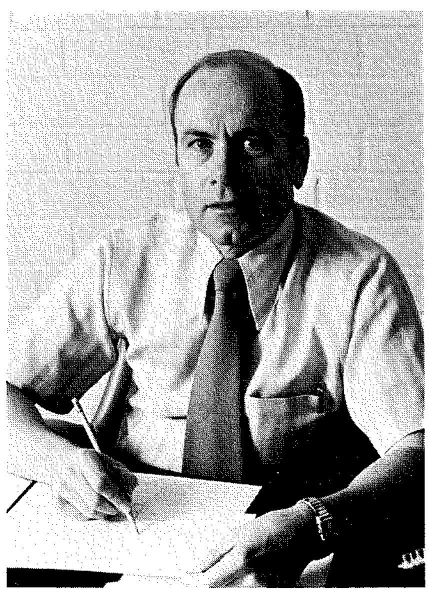

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To use Fig. 3 to predict upstream signal-to-noise ratio from a system layout, count the number of trunk-bridger amplifiers used in the area of concern, find this number on the horizontal axis of the graph and draw a vertical line through that number. Next, count the number of distribution amplifiers operated from the previously mentioned trunk-bridger amplifiers, find this number on the vertical axis, and draw a horizontal line through that number. The intersection of the horizontal and vertical lines will indicate the signal-to-noise ratio in the upstream portion of the system. An example, shown in Fig. 3, of a trunk system having 62 trunk-bridger amplifiers and 108 distribution amplifiers yields a 46.5-dB signal-to-noise ratio. The family of curves from Fig. 3 shows that upstream signal-to-noise ratio is definitely a function of system density rather than cascade length.

Although the family of curves shown in Fig. 3 allows the system designer to predict upstream signal-to-noise ratio when using EiE equipment, it is not necessarily valid for all manufacturer's equipment. A similar type graph would have to be generated for each line of equipment.

Radio frequency interference

Radio Frequency Interference (RFI) is another design parameter that requires special consideration for two-way cable systems. The downstream cable system using the vhf spectrum has had to contend with the RFI caused by consistently controlled signals of off-air tv stations.

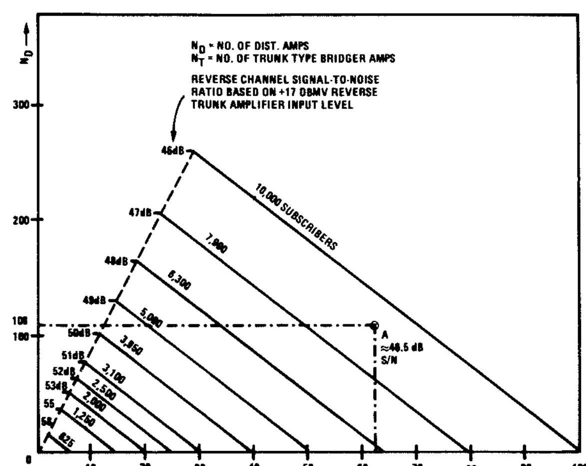

Fig. 3 — Preferred region of operation for EiE distribution system.

In many cases, this interference can be avoided by operating the cable signals off-channel. When operating on-channel with strong off-air signals, the RFI integrity of the downstream cable system becomes very important.

The upstream cable system can be faced with a hostile RFI environment, since many different frequency allotments and services are provided in the 10-MHz to 30-MHz frequency range. Some of the common sources of RFI leakage or pickup in the upstream cable system are poor grounding of connectors to the cable sheath, poorly shielded house-drop cables, and exposed twin-lead at the subscriber's tv set. Poor shielding or lack of RFI gaskets in taps and amplifier housings can add to the leakage problem.

The connectors used in a two-way cable system should be chosen carefully with attention given to the technique used to ground the connector to the cable sheath. An effective RFI-proof connector will have a large contact surface area to clamp to the sheath of the cable. A sleeve will be used to insert under the cable sheath to restrain the sheath from taking a permanent set as the connector is tightened. An RFI-type O-ring moisture seal should also be used. Such connectors are available.

Subscriber drop cables can act as unwanted antennas when their shielding is ineffective. Solid-sheath or double-braid house-drop cables have proven to be effective for RFI performance.

RFI pickup from the tv set, or from the set twin-lead acting as an antenna, can best be eliminated by using a highpass filter at the set end of the drop cable.

Hub system

Because of the previously mentioned conditions and potential problems associated with two-way cable systems, it is desirable to separate or isolate different sections of the cable system as much as possible. A hub-type cable distribution system, with as many separate trunk lines emerging from the hub as possible, can maintain the upstream signal-to-noise ratio at an acceptable level (see Fig. 4).

The isolation obtained from the hub system is also very useful for troubleshooting any RFI leakage problems that can occur in the system. For example, a

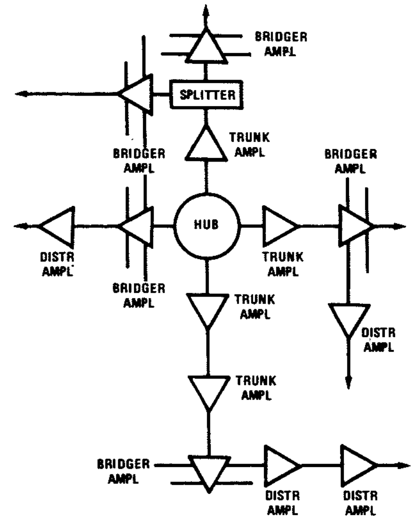

Fig. 4 — Hub system.

ten-trunk hub split would isolate an RFI leakage problem to within 10% of that hub system. Final isolation would require field testing with proper test equipment. A spectrum analyzer which covers the 10-MHz to 30-MHz frequency range is a valuable asset for RFI leakage testing.

The location of the hub, or hubs if a large system is under consideration, is an imporatant part of the system design. The optimum hub location for the downstream cable system may not be the same as the optimum hub location for the upstream portion of the cable system. Fig. 5 shows an example of this condition.

In Fig. 5, a situation exists where the subscriber density is concentrated on one side of a system (area 2, 3, 4, and 5) with a small number of subscribers located at the opposite side of the system (area 1). If the hub is located for optimum downstream system performance (in this case 15 cascade maximum) as is shown in Fig. 5a, the upstream signal-to-noise ratios on the two trunk lines at the hub do not balance. In fact, the signal-to-noise ratio is 7 dB lower on the cable coming from the high density part of the system than the signal-to-noise ratio on the cable from the less dense part of the system.

This upstream signal-to-noise performance can be improved by locating the hub near the high density part of the system as is shown in Fig. 5b. In this configuration, each area is fed from a separate trunk line which means the upstream noise build-up will be less on the individual trunk cables at the hub. A

6-dB upstream signal-to-noise improvement is obtained by this particular hub location compared to the previous hub location. The penalty for this 6 dB improvement is an increase of the trunk cascade length to area 1, from 15 to 20, which would degrade the downstream signal-to-noise ratio approximately 1 dB.

In the case of Fig. 5, a compromise location would be required for optimum system operation. Physical and economic consideration may also influence the location of the hub.

Level control

Various methods of controlling the levels of the upstream signals are possible. Some of them are:

1) Thermal compensation.
2) Composite AGC.
3) Programmed AGC from the downstream AGC circuits.
4) Pilot-carrier AGC.

Thermal compensation is much less complex than the other types listed; tests in environmental chambers and in the field indicate it to be more than adequate. When you consider that a 20 dB cable spacing changes less than 1 dB in the sub-low frequency band for a 120°F temperature change, an elaborate AGC system is not necessary or justifiable.

Composite AGC suffers from the problem of maintaining a signal in all upstream amplifiers in the cable system. When a signal is removed from one of the amplifiers, its AGC circuit will turn the amplifier gain to a maximum which will increase the upstream system noise level.

Programmed AGC from the downstream AGC circuits can be made to work as well as thermal compensation but it is much more complex than thermal programming (thermal compensation).

Pilot carrier AGC requires pilot carrier generators at the downstream end of each trunk cascade, which can produce interference problems due to many pilot generators combining at the main trunk lines.

Channel capacity

Channel capacity is another factor to consider when designing a two-way cable system. The 10-MHz to 30-MHz sub-low frequency band allows three video channels to be sent simultaneously in the upstream direction on each trunk cable which originates from the headend or hub. A hub-type system can therefore increase the upstream channel capacity. For example, a ten-trunk hub split with three upstream channels on each trunk cable would provide a total of thirty channels in the upstream system. A four-trunk-line hub split would provide twelve simultaneous upstream channels.

The recommended placement of the video signals in the 10-MHz to 30-MHz spectrum is on channels T8, T9, and T10; however, operation with inverted or reverse spectrum is recommended (picture carrier frequency higher than the sound carrier frequency) (see Fig. 6). Inverted carrier operation places the three upstream picture carriers at frequencies that are less than one octave apart, thus eliminating any possibility of second-order distortion.

Adding to the desirability of the inverted spectrum transmission in the upstream cable system is that it allows a single conversion signal processor to restore the sub-low channel to a standard vhf channel. A single conversion processor will have more dynamic range than a double conversion processor and thus less degradation in picture quality. A typical application of the sub-low upstream channels is shown in Fig. 7. The sub-low modulator, signal processor and two-way cable system hardware are all available for present day system needs.

Conclusion

It is important, in two-way cable systems, to put proper emphasis on the upstream portion of the system. The time to consider upstream noise buildup, RFI leakage, hub locations, and sub-low headend equipment is at the same time the downstream system is designed. Building a one-way cable system which is "expandable to two-way" could be a disappointing experience when an attempt is made to build and operate the upstream portion of the system at a later date.

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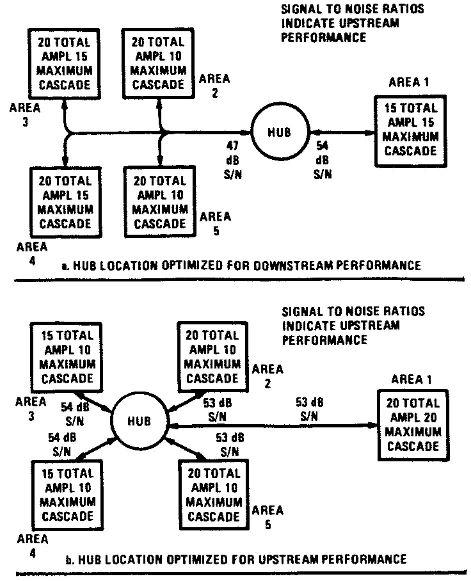


Fig. 5 — Hub locations.

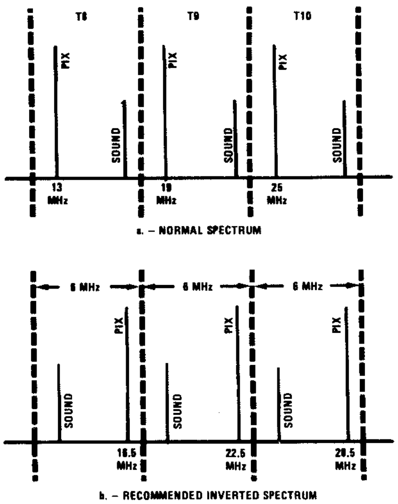


Fig. 6 — Sub-low channels.

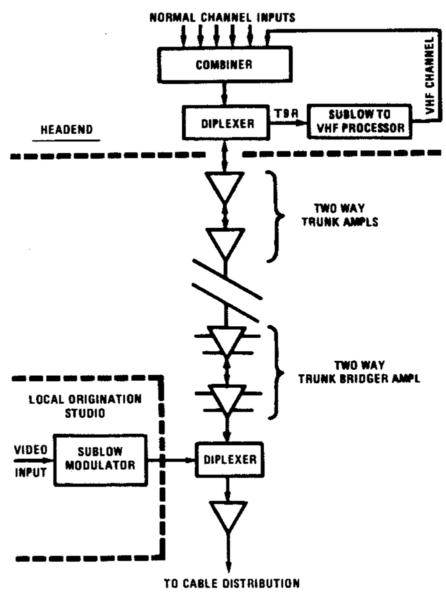
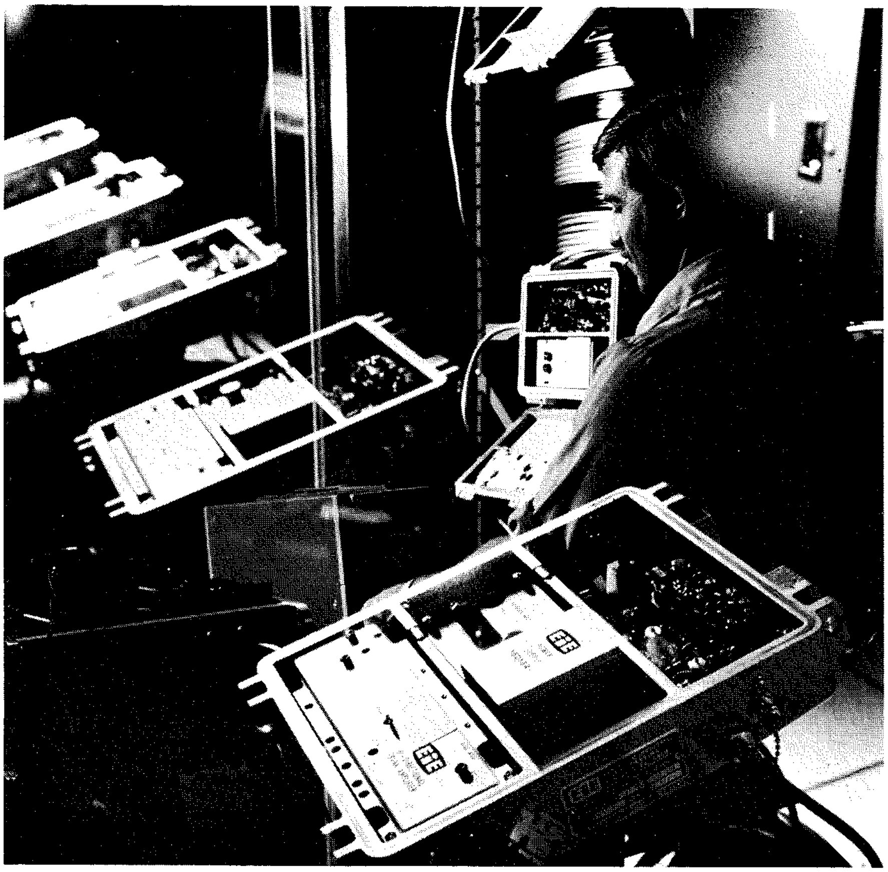


Fig. 7 — Typical upstream channel application.



Required system triple-beat performance

B. Arnold

A curve showing the triple-beat level required in a system for high quality performance versus the number of triple beats per channel is given in this paper. The curve is the result of subjective tests conducted at EiE in which the threshold of perceptibility of the third-order intermodulation (triple beat) is observed on a tv receiver. The paper discusses the development of the curve and substantiates it by probability theory. A table is also given which lists from 1 to 30 channels and the number of triple beats that would be generated from these channels on the worst channel. This gives the system designer the necessary triple-beat performance required to design a multichannel system.

THIRD-ORDER INTERMODULATION products (triple beat) have been much discussed (and cussed) over the past couple of years in the CATV industry. Most manufacturers now include third-order intermodulation (triple beat) in their specification, many CATV operators are considering their effects in proposal requests, and some manufacturers are offering coherent headends to reduce the effects of all intermodulation distortion. Up to now, the industry has depended on educated guesses with regard to the visibility of third-order intermodulation products (triple beats) in a tv channel. This paper presents the results of subjective tests conducted at EiE which allow the system designer to accurately determine triple-beat specifications that will provide the protection needed to avoid picture impairment.

Number of spurious beats

A computer run[2] has been made giving the exact number of beats on each channel. Table I summarizes the worst case for various numbers of channels. The computer run[2] showed that the center channel in a group of channels will be the worst case for the number of spurious beats. The two types of third-order spurious signals given in the Table are the signals resulting from two carriers ($2f_1 + f_2$ type) and the signals resulting from three carriers or triple beat ($f_1 + f_2 \pm f_3$ type). The spurious signals resulting from the two carriers are one-fourth (−6 dB) the level of the spurious signals resulting from the three carriers.[1] The effective or equivalent total number of triple beats would therefore be the sum of this number of triple beats plus the sum of one-fourth the number of two-channel beats.

Intermodulation threshold

A limited amount of subjective testing has been done in determining the threshold of perceptibility of third-order

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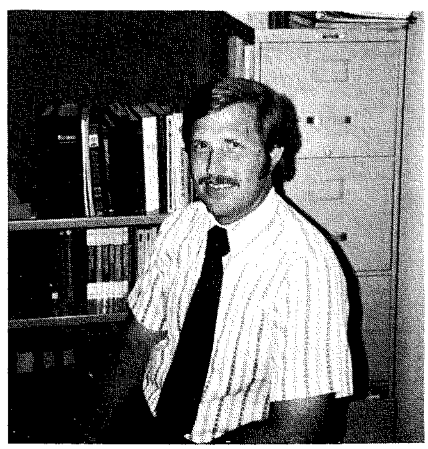

intermodulation distortion, and the data that is presented in this paper are based on experiments performed at EiE. The test setup is shown in Appendix A. Channel 10 was modulated with staircase modulation (grey scale) and other channels were unmodulated. The unmodulated carrier is a worst-case condition for viewing third-order intermodulation distortion, but it provides for more consistent results just as synchronous modulation does for cross-modulation testing. The unmodulated carriers would represent the synchronous tip power in the worst-case condition where all channels were synchronously modulated. If the carriers were modulated, there would be an improvement in the threshold, but the amount of improvement would depend upon the characteristics of the modulation. The threshold given in Fig. 1, therefore, provides some safety factor for an actual system just as synchronous cross modulation does. The threshold of perceptibility is not necessarily an acceptable level that the average viewer would tolerate, but is the worst-case condition. Also, no attempt was made to space the channels so that the spurious signals fall within the null points around the carriers. The interfering effects of intermodulation on the television screen are reduced when the spurious signals are offset at frequency intervals about the carriers at approximately the half-line scanning fre-

Glossary

Beats — Sum and difference frequencies produced from the product of two or more frequencies.

Coherent headend — A headend in which an identical frequency spacing exists between the various picture carriers of the various channels.

Intermodulation — In a nonlinear transducer element, the production of frequencies corresponding to sums and differences of the fundamentals and harmonics of two or more frequencies transmitted through the transducer.

Carrier to intermodulation ratio — The ratio between the carrier level and the level of the intermodulation.

Intermodulation distortion — The impaired fidelity resulting from the production of new frequencies that are the sum and the differences between frequencies contained in the applied waveform.

Intermodulation products — The frequencies produced by intermodulation.

Third-order intermodulation — Intermodulation resulting from the cubic, X^3, characteristics of a nonlinear transducer element. Includes two-frequency beat ($2f_1 \pm f_2$) and triple beat ($f_1 + f_2 \pm f_3$), and third harmonic ($3f$).

Third-order spurious signals — Unwanted signals resulting from the cubic, X^3, characteristic of a nonlinear transducer element. Third-order intermodulation.

Triple beat — Sum and difference frequencies produced from the product of three frequencies. Third-order intermodulation resulting from three frequencies, $f_1 + f_2 \pm f_3$.

Equivalent triple beat — The total number of triple beats plus one-fourth the total number of two-frequency beats.

Individual triple beat — One of many triple beats produced from the product of three frequencies.

Threshold of perceptibility — The level at which an effect (i.e., intermodulation distortion) is first observed on a tv receiver.

NO. OF CHANNELS	CHANNELS	CENTER CHANNEL	MAX. NO. ON CENTER CHANNEL $2f_1 \pm f_2$	$f_1+f_2 \pm f_3$	TOTAL EQUIVALENT TRIPLE BEAT
0	0	0	0	0	0
1	13	13	0	0	0
2	12-13	13	0	0	0
3	11-13	12	0	1	1
4	10-13	12	1	2	2.25
5	9-13	11	2	4	4.5
6	8-13	11	2	7	7.5
7	7-13	10	2	11	11.5
8	I-13	10	3	15	15.75
9	H-13	9	4	20	21
10	G-13	9	4	26	27
11	F-13	8	5	33	34.25
12	E-13	8	5	40	41.25
12 STD	2-13	10	2	19	19.5
13	D-13	7	6	47	48.5
14	C-13	7	6	56	57.5
15	B-13	7	7	65	66.75
16	A-13	I	7	77	78.75
17	A-J	I	8	88	90
18	A-K	I	8	100	102
19	A-L	I	9	112	114.25
20	A-M	7	9	125	127.25
21	A-N	7	10	139	141.25
22	A-O	8	10	157	158.25
23	A-P	8	11	170	172.75
24	A-Q	9	11	187	189.75
25	A-R	9	11	204	206.75
26	6-R	10	12	204	207
27	5-R	10	12	206	209
28	4-R	9	12	212	215
29	3-R	9	12	219	222
30	2-R	8	13	226	229.25

Table I — Third-order spurious.

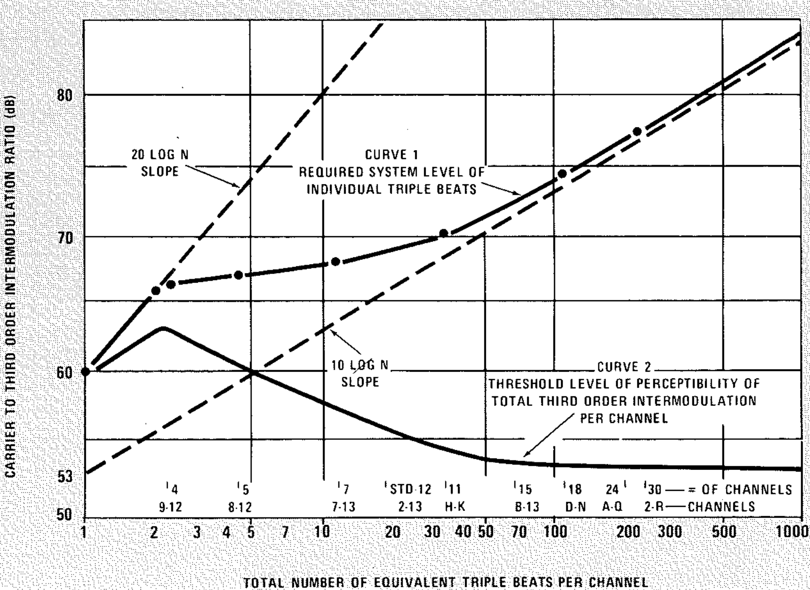

Fig. 1 — Individual triple-beat level and total third-order intermodulation level versus total number of equivalent triple beats per channel.

quency, or 8 kHz. The channels and frequencies used in the test are given in Appendix B.

Fig. 1 is a plot of two curves, curve 1 being the required system level of an individual triple beat as a function of the total number of beats per channel. The measured points on curve 1 were found by observing the threshold of perceptibility on a tv receiver of the total third-order intermodulation on one channel and then measuring the level of an individual triple beat with the spectrum analyzer.

Curve 2 is a plot of the actual threshold level of perceptibility of the total third-order intermodulation per channel as a function of the total number of equivalent triple beats per channel. Curve 2 is derived from curve 1 by summing levels of the third-order intermodulation products that fall on one channel using the formula:[3]

$$V_{rms} = [\sum v_i^2]^{1/2}$$

Curve 2 is given for reference since it is very difficult to measure the summation of the third-order intermodulation products in a system due to the level of system noise. The bandwidth cannot be reduced enough to eliminate the system noise and still sum the intermodulation products; therefore, curve 1 or the individual triple beat should be used when specifying or measuring third-order performance in a system.

The unusual shape of curve 1 in Fig. 1 requires some explanation. When one spurious signal was beating with the viewed channel, the threshold of perceptibility was a carrier-to-intermodulation ratio of 60 dB. This value agrees with most of the testing that has been done in the past with a single beat frequency. Increasing the number of random spurious signals to two, the worst-case threshold increases to 66 dB due to the periodic summing of the peaks of the two spurious signals. As the number of spurious signals is doubled, the theoretical threshold level would increase 6 dB due to the periodic summing of the peaks of the spurious signals, but subjective viewing has shown that the periodic summing of more than two spurious signals becomes quite random. The curve makes a sharp transition after two spurious signals and crosses over to another line which is a power addition (10 log N) of the spurious signals.

The shape of the curve in Fig. 1 is substantiated by probability theory. The central-limit theorem[4,5] in probability states that as the number of independent random variables approaches infinity, the density approaches the normal density curve (Gaussian distribution) which is the density for white noise. This would explain why, after a large number of random signals (which the triple beats are), the signals add on a power basis (or in the same way as random noise).

Several authors[4,5] of probability theory have pointed out that since the central-limit theorem involves a limit of infinity, one might feel that the number of random variables must be large before an approximation of the normal distribution can be made.

However, the convergence for many of the ordinary density functions is surprisingly fast. In fact, the normal distribution curve is closely approximated by just three random variables, and four random variables is an "extremely good" approximation.[4] This explains the sharp transition in Fig. 1 after just two spurious signals.

Since the large number of spurious signals are all clustered about the signal carrier within ±20 kHz, the effect is narrowband random noise as compared to wideband (4 MHz) random noise. Extrapolating the 10 log N slope of the curve in Fig. 1 back to the vertical axis, one finds that the threshold for narrowband random noise would be 53 dB. To compare directly the visual effects of noises having different power spectra, a weighting factor[6] must be used. The weighting factor of a specific noise spectrum is obtained by integrating the noise spectrum multiplied by the weighting function, over the video bandwidth to be considered. The weighting factor from narrowband flat noise to wideband flat noise is −6.1 dB. Therefore, the random noise threshold in Fig. 1 would be approximately 47 dB for wideband (4 MHz) random noise.

Intermodulation reduction

Some of the methods that can be used to reduce the third-order intermodulation distortion in a system were discussed in the earlier papers[1], and will be mentioned again for clarity. Device manufacturers are constantly being urged to develop transistors with greater linearity to reduce intermodulation distortion. Until such linear transistors are developed, the power will have to be limited to maintain good quality pictures. The power can be reduced by operating the amplifier with a tilted output, that is, the low channels operating at progressively lower levels than the higher channels to equalize for the cable attenuation.

Another obvious method to reduce the power is to reduce the output level of all amplifiers, but in order to maintain the signal-to-noise ratio, the spacing must be reduced an equal amount. Fig. 1 shows that there is an approximate 8-dB degradation in the threshold level when changing from standard 12 channels to 30 channels, which means that for 30 channels the output level of each amplifier should be reduced 4 dB for equal performance to a standard 12-channel system.

The beat between the third-order products and the signal carrier could be eliminated by utilizing a coherent headend, but this would not eliminate the sidebands resulting from the modulation of the spurious signals. The improvment in picture quality in a system with a coherent headend will depend upon the relative threshold of intermodulation and cross-modulation distortion. A problem with the coherent headend would be in the case where it was desired to phase lock to two or more "off air" channels (vhf broadcast) in a high-signal-level area, only one channel could be used. The most practical way to deal with the intermodulation problem is to consider it when designing a system to insure that the picture quality is not degraded by the intermodulation.

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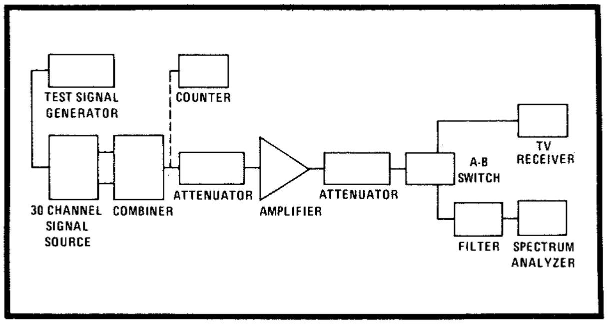
Appendix A — Test setup for triple-beat measurement.

EQUIPMENT	MANUFACTURER	MODEL
30 Modulators	EiE	CTMI
1 NTSC Test Signal Generator	Tektronix	R140
1 Counter	Eldorado	1450
2 Combiners	EiE	151125-1
2 Attenuators	Texscan	SA-78
1 Test Amplifier (Hybrid Module)	TRW	CA613
1 A-B Switches	EiE	AB5-75
1 TV Receiver	RCA	XL-100
1 Set of Filters	Hamlin	BPF10
1 Spectrum Analyzer	H.P.	8554L/8552A



Appendix B — Test-channel frequencies.

CHANNEL	FREQUENCY MHz	Δf khz
2	55.2515	+1.5
3	61.2469	−3.1
4	67.2449	−5.1
5	77.2480	−2.0
6	83.2493	−0.7
A	121.2469	−3.1
B	127.2492	−0.8
C	133.2423	−7.7
D	139.2461	−3.9
E	145.2504	+0.4
F	151.2496	−0.4
G	157.2504	+0.4
H	163.2503	+0.3
I	169.2493	−0.7
7	175.2407	−9.3
8	181.2450	−5.0
9	187.2421	−7.9
10	193.2452	−4.8
11	199.2500	0.0
12	205.2396	−10.4
13	211.2500	0.0
J	217.2497	−0.3
K	223.2476	−2.4
L	229.2454	−4.6
M	235.2428	−7.2
N	241.2464	−3.6
O	247.2517	+1.7
P	253.2520	+2.0
Q	259.2550	+5.0
R	265.2521	+2.1

Appendix C — Determination of system triple beat from Fig. 1.

Assuming that an operator plans to operate his system with 24 channels, *A-Q*, on one cable, the determination of the system triple beat (carrier-to-triple-beat ratio) is as follows:

In Table I, the first column is labeled "number of channels." Find 24 channels in this column and read across to the last column labeled "Total Equivalent Triple Beat" to find the total number of equivalent triple beats that the center channel has.

From Table I, 24 channels = 189.75 spurious signals on the center channel 9.

In Fig. 1, the Horizontal Axis is labeled "Total Number of Equivalent Triple Beats per Channel" and the Vertical Axis "Carrier to Third Order Intermodulation Ratio." Find the Horizontal Axis 189.75 spurious signals, and the intersection with curve 1. From the intersection of curve 1, read across to the Vertical Axis for the triple beat required in a system.

From Fig. 1, the system triple-beat specification for 24 channels would be approximately 76 dB.

As mentioned previously, this specification is for the threshold of perceptibility for unmodulated carriers. If the carriers were modulated there would be an improvement in the threshold, but the amount of improvement would depend upon the characteristics of the modulation. The unmodulated carrier is a worst-case condition for viewing third-order intermodulation distortion, but it provides for more consistent results just as synchronous modulation does for cross-modulation testing. Many systems have been able to operate with a less than worst case triple-beat specification mainly because of the modulation on the carriers, but there is no safety left for variations in system levels due to temperature changes or aging.

Appendix D — Determination of system triple beat from amplifier specifications.

Assume a system cascade of 20 trunk amplifiers, one bridger and one distribution (line extender). The number of channels will be 30, Channel 2 through channel R.

EiE Series 150 triple-beat specificaitons are:

	Triple beat	*Output at* 300 MHz
Trunk	116 dB	30 dBmV
Bridger	82 dB	47 dBmV
Distribution	88 dB	44 dBmW

The degradation for a 20-amplifier cascade would be 26 dB.

20 trunk = 116 dB − 26 dB = 90 dB
20 trunk + bridger = 90 dB + 82 dB = 79 dB
20 trunk + bridger + 1 distribution = 79 dB + 88 dB = 76.5 dB

The system specification of 76.5, for practical purposes, meets the required triple-beat specification in Fig. 1 of 77 dB.

If two distribution amplifiers are cascaded, the levels can be reduced 2 dB to maintain the specification.

20 trunk + bridger + 2 distribution = 90 dB + 86 dB + 92 dB + 92 dB = 76.6 dB

Color frame storage using a silicon storage tube

D. W. Henrichsen

This paper describes the electronic storage of "off-the-air" color tv signals by use of a silicon storage tube; a complete frame of video is stored and read out from the tube. The challenges and solutions experienced during the color-frame-storage project are described. The means of handling the tv signals external to the storage tubes and the concept modifications needed to accommodate color are described (the internal operation of storage tubes for monochrome storage is covered in the April-May, 1972, RCA Engineer).[4]

AFTER the introduction of silicon storage tubes, several means of storing color on them were proposed and developed. This paper describes a logical extension of the state-of-the-art development of color storage...in that a complete frame of NTSC video is electronically stored successfully and read out from the storage tube. The project described herein is aimed at storing off-the-air signals from a tv set and storing real-time or multiplexed-time video signals distributed by a cable. The work involves the means of handling tv signals external to the silicon storage tubes and development of modification concepts that will accommodate the color tv frames.

Early studies

An early study of color storage and retrieval from a silicon storage tube was carried out by D. P. Dorsey[1] of DSRC in

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Since writing this article, Mr. Henrichsen has left RCA.

1969-70. J. G. Amery and G. F. Awbrey of CE (Indianapolis) started similar work in early 1971 using the "Interleaved Carrier System;" this system was developed at RCA Laboratories under Dr. J. J. Brandinger's leadership. During this work, it was observed that color frame storage improved as the stripes of the storage tube were rotated toward the vertical.

J.G. Amery then converted to the use of alternate line phase reversal of NTSC chroma (de-interlaced); this approach resulted in the storage of vertical stripes of chroma dots and was first demonstrated in December, 1971. The Amery decoder reversed the chroma-phase again every other line, so that an NTSC color receiver displayed the normal crisscross pattern of NTSC chroma dots.

My responsibilities for the project began in June, 1972. At this time, a number of electron optic defects were removed that had caused severe signal shading and focus nonuniformity. The NTSC video encoding/decoding techniques were improved and a completely self-contained NTSC color storage system was delivered to DSRC in April, 1973.

Color storage

The storage of color on the silicon storage tube (C22047) was possible due to previous knowledge gained during development of the single-tube color camera at Consumer Electronics (CE).

The problems related to electron-optics defects that required answers were spot size of the electron beam, non-uniform focus, and uneven amplitude of the recovered signal (shading). A deflection non-linearity, or instability, caused the recovered chroma phase to shift or to lose chroma lock.

The video processing problems were recovered signal quality (S/N) and gray-scale tracking non-linearities in the luminance signal. The chroma level was also non-linear as the luminance changed from black to white.

Electron-optics solutions

The solutions for problems in electron optics derived for the color camera under J. H. Wharton's leadership (CE) provided the necessary groundwork for color storage.[2] The "inside-out" yoke/focus-coil arrangement provided 5 MHz response over 90% of the raster areas of the 8507-A vidicon. The focus coil was segmented, while the yoke was a slightly altered version from an early monochrome tv receiver. Because of the recovery of the burst at the start of each scan line, the spot resolution requirement of the storage tube was at least 4 MHz at the raster corners.

The need for a small spot size required a determination of the electron lens field inside the tube. The lens field was found by combining theory with trial and error. Using a conventional yoke/focus coil arrangement, the 8507-A vidicon elements G_2, G_3, and G_4 were typically set to +300V dc, 450V dc, and 600V dc respectively. The improved beam control with the Wharton yoke/focus coil allowed the G_2, G_3, and G_4 voltages to be set at 300V dc, 650V dc, and 1300V dc, respectively. This resulted in less beam ellipticity at the raster edge and greater resolving power; also, the light portholing of earlier yoke/focus assemblies was reduced greatly.

The video processing techniques used for a 5 MHz low-noise video preamplifier in the single-tube color camera were applied to the storage tube. The methods used in the color camera to compensate for the luminance nonlinearity were applicable to the storage tube; these techniques are described later in more detail.

Solutions unique to silicon storage tubes

The requirement on electron optics in the silicon storage tube necessitated the alteration of the J. H. Wharton electron lens for the 8507-A vidicon. The overall G_2-to-target distances of the vidicon and storage tube were similar, but the storage tube had a completely different G_3/G_4 aspect ratio. This difference caused the beam in the storage tube to be elliptical in the corners and produced poor corner focus. The beam landing velocity also changed over the raster producing uneven signal amplitudes. A large coil over the target end of the storage tube produced the desired spot resolution and field flatness. For satisfactory storage and recovery of the color information, the voltages on the silicon storage-tube elements must remain stable to within 0.1% during *erase*, *write*, and *read* modes of the storage tube.

The proper synchronization of the chroma signal recovered from the storage tube created additional stability requirements. No timing marks or codes are used during the *write* time, so that proper color-phase recovery depends on very noise-free and stable deflections. The *read* electron beam must exactly retrace the path of the *write* beam (Fig. 1). In our system, phase reversal of the recorded color subcarrier signal during alternate lines reduced resolution requirements of the beam and gave more tolerance to deflection stability. Notice that only the amplitude of the 3.58 MHz subcarrier changes when the deflection is non-stable or the beam non-symmetrical (Fig. 2).

Every line of stored color information starts with the burst signal. When the burst phase is wrong due to scan non-linearities, chroma quality suffers

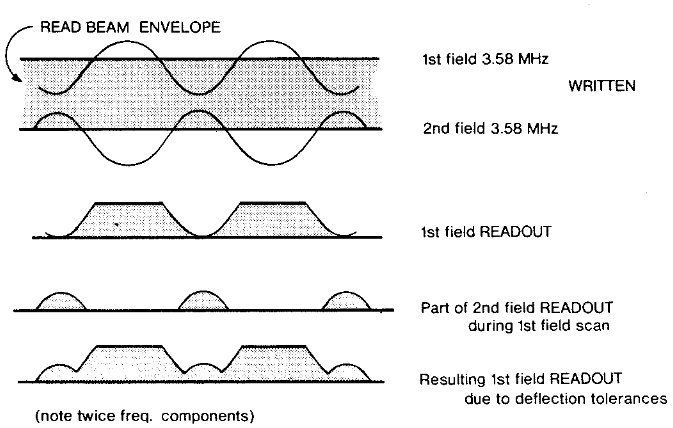

Fig. 1 — Normal 3.58-MHz readout with write/read deflection shifts.

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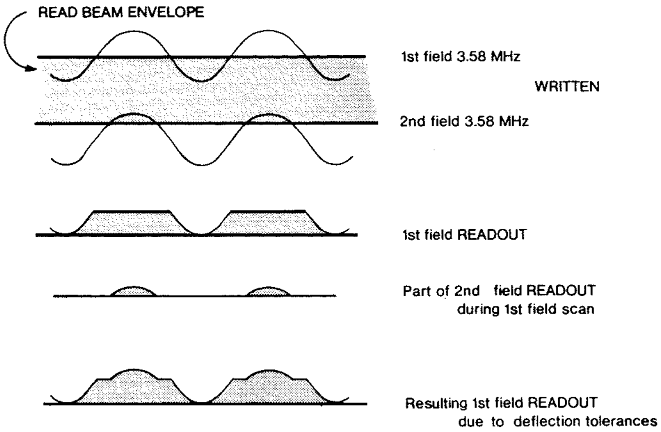

Fig. 2 — De-interlaced 3.58-MHz chroma readout with write/read deflection shifts.

through the complete scanning line. If the burst is not recovered due to poor electron optics, chroma will be lost for the complete line of scan. To recover the burst, an advanced horizontal deflection (2μs) was used to insure that the horizontal flyback spikes and scan nonlinearities wouldn't block burst recovery.

To maintain accurate phasing of the subcarrier, it was also necessary to change the voltage on the G_4 electrode in the storage tube by a factor of 2% when switching from *write* to *read*.

Timing and power supply

A general block diagram (Fig. 3) illustrates the steps required to achieve the proper waveform timing and voltage control. A delayed sync arrangement was necessary so that the timing circuits would produce a correctly placed burst gate for the stored video input. The *write* and *read* portions of the video processing each cause a 0.27 μs delay. The remaining 0.56 μs of delay is produced mainly in the tube.

The logic sequencer is clocked by vertical sync and it is activated by the front panel controls for *cycle*, *write*, or *erase*. The *cycle* button produces an erase pulse of 20 frames and a write pulse of 2 frames duration. The unit then reverts back to the *read* state. It must be noted that only one frame of video is seen by the storage tube. The sequencer blanks the cathode of the tube during the first *write* frame of video. This is done so that all switching transients in the video processing and high voltages have terminated before the recording beam is turned on. About 5% of the cathode blanking is devoted to parabolas used to remove portholing caused by beam-landing errors. These parabolas are removed during *erase* so that a more uniform tube background is present on which to *write*.

It was found that a typical untrained observer could see that the output of the storage tube displayed on the monitor changed when either the focus current, the deflection currents, or the G_4 voltage changed by 0.5%. The focus currents, deflection currents, high voltages, and power supplies are therefore regulated to 0.1% or better.

Video processing circuits

The video processing portion of the circuitry used with the silicon storage tube (C-22047) is shown in the block diagram of Fig. 4. *Read* and *write* relationships are illustrated and described below. The states of the switched tube elements are shown in Table I. The *loop thru* feature allows the operator to preview the video to be stored on the silicon storage tube.

"Write" video

The *write* mode prepares the video for storage on the tube. The switches on the block diagram are actually transistors driven from the logic sequencer during the *write* time. The luminance and chrominance signals are first separated, then processed separately, and finally recombined for writing on the storage tube.

The luminance signal may be altered to compensate for the black video compression that occurs in vidicons and storage tubes. This "black stretch" may be done either during *write* or *read* or divided up between *write* and *read*. This is a subjective decision based on signal/noise consideration and individual storage-tube characteristics. The noise is also accentuated as the video blacks are stretched.

The "white stretch" is an artistic touch used to improve the apparent picture contrast. Very small amounts of peak white stretching go a long way and this is also an optional feature. It is mandatory that the video into the luminance gamma amplifier be dc restored.

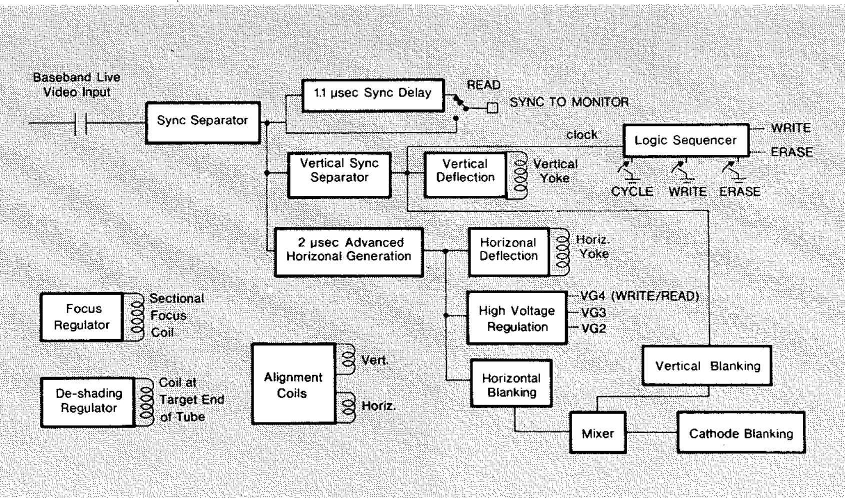

Fig. 3 — Waveform timing and dc control block diagram.

After separation, the phase of the chroma is reversed every other line as a means of reducing the required beam resolution of the tube. The chroma goes on the tube as vertical stripes of dots rather than the crisscross pattern of dots observed on the monitor when NTSC chroma is present. This phase reversal of the chroma also takes place during the *read* mode, so that the stored video appears as normal NTSC to the color receiver or monitor.

To improve the recovery of the stored chroma, a *write* chroma preemphasis of 6 to 10 dB is employed. The *write* burst is also put on a small pedestal and preemphasized an additional 6 to 10 dB so that the burst amplitude is about equal to the peak video amplitude.

Table I — States of switched tube elements.

	Erase	*Write*	*Read*
G_1	ground	−55Vdc + 15Vpp video	−40Vdc
Target	+ 18 ±3 Vdc	+ 175Vdc	+10 ±3Vdc
Cathode	+25Vdc gnd	as in *erase* but with 5% parabolas	as in *write*
VG_4	+1323Vdc	+1300Vdc	+1323Vdc
VG_3	+ 585Vdc		
VG_2	+ 300Vdc		

"Read" video

The preamplifier was designed to compensate for a simple rc-type rolloff with a −3 dB point of 300 kHz. Percival coil peaking, capacitive peaking, and other methods of high-frequency peaking were used to improve the video signal/noise ratio. The signal from the preamplifier is separated into luminance and chrominance components. The phase of the chrominance signal is then reversed during alternate lines to convert it back into a standard NTSC chrominance signal. The two components are then recombined and the resulting signal is displayed on the monitor. When properly set up, the S/N is acceptably good to most viewers. As the stored picture is *read* for longer periods of time, the S/N decreases as expected.

Possible uses of system

An example for SST system application might be its use as a color display in the Videovoice system.[3]

The storage tube may also be used in the cable tv systems of the future for providing single frame video services to subscribers. B.J. Lechner and C.M. Wine of RCA Laboratories are involved in developing this concept.[4]

The tube may also be used as a tool for the video disc research and design; sections of video can be conveniently stored and analyzed during research studies.

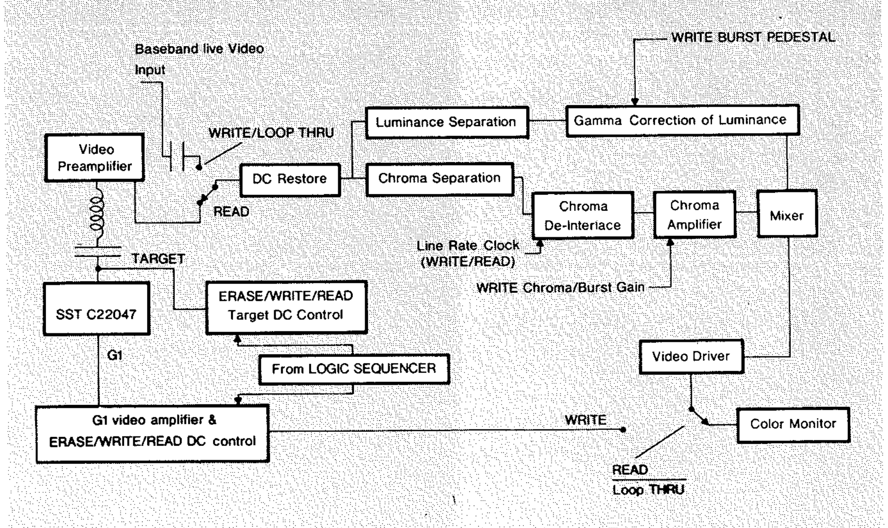

Fig. 4 — Block diagram showing erase/write/read dc control of silicon storage tube, C22047.

Conclusions

A means of storing a complete NTSC "off-the-air" type signal has been developed and the C22047 silicon storage tube is being used to store one frame of video information including the NTSC chroma burst. When properly set up, the retention time is between 4 and 10 minutes depending upon the tube. The signal/noise decreases gradually until the background is almost completely white after this 4 to 10 minute interval. The *read* picture quality is good with typical peak/peak, video/RMS noise figures of 32dB and optimum S/N figures of 40 dB.

A better method of video processing may be desirable in the future. In the present system, the chroma bandpass is too wide. This results in undesirable phase changes in the luminance components during alternate lines. It may be desirable to comb the chroma before de-interlacing.

More work on the electron optics and system dynamics is also needed, since the present system is based on experience with a relatively small number of C22047 storage tubes. A final commercial design may require joint study of the electron optics and the proper deflection/focus field-forming circuitry; such a study would involve both the circuit and the tube designers.

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Elimination of adjacent-channel interference on multi-channel CATV systems

Dr. H. G. Schwarz

Television receiver and transmitter characteristics required for the elimination of visible adjacent-channel interference effects are discussed. The interference mechanisms involved are explained. Effects of an upper-adjacent-channel signal are different in nature from those of a lower-adjacent-channel signal. To suppress upper-channel interference, little extra effort, if any, is needed in the receiver, but special care is necessary in designing the VSB filter response at the transmitter (or headend modulator). To eliminate interference from the lower adjacent channel, however, the receiver selectivity has to meet requirements which are more stringent than those for the upper adjacent channel. The quantitative analysis of the selectivity requirements is supported by experimental data. The measurement methods employed are discussed.

A PROBLEM was posed, some two years ago, to RCA Laboratories: Given a cable television system distributing programs on all 12 standard vhf channels; what is the receiver selectivity required to avoid visible interference from signals transmitted on adjacent channels?

The problem of adjacent-channel interference is as old as tv broadcasting, but for broadcast signals it had been greatly alleviated by proper channel allocations. A wealth of experimental data on adjacent-channel interference can be found in the literature.[1] Such data, however, turned out to be of little use for the task on hand. First, the problem to be solved by previous tests was posed in a different way: the receiver selectivity was considered a given fact, and the question was how to avoid adjacent-channel interference by means of geographic channel allocation, radiated power ratios, *etc*. For the problem now posed by the full 12-channel program complement offered on CATV systems, the relative signal levels in adjacent channels are the given fact (they are nominally equal on cable); the receiver's frequency response required to avoid interference represents the unknown factor. Receiver selectivity requirements for the case of CATV, furthermore, are different because the sound carrier level as transmitted on a cable system is lower than that broadcast over the air.

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A second reason for trying a new and different approach was the fact that several major parameters affecting interference visibility had been lumped together in the past, and none of them was controlled during the experiments. As a result, the recorded data on subjective interference ratings show a very wide spread.

Methods of testing

As a new and better approach, we decided to establish interference visibility threshold levels for *worst-case* conditions. This is meaningful and, in practice, also a useful definition. It, furthermore, has the advantage that the threshold levels can be determined, as we found, with an accuracy of ±1 dB by trained, critical observers. Measurements when repeated after several days or weeks still gave the same results.

Worst-case conditions were defined as follows:

- The video signal in the desired channel is chosen for easiest recognition of interference,
- The video signal in the interfering adjacent channel is chosen to cause a maximum of interference in the desired channel,
- The interference, when visible as beat pattern or video picture, must be stationary or only slowly moving,
- Finally, as already mentioned, the threshold levels are determined by trained, critical observers with good vision.

To realize such worst-case conditions:

- The "desired picture" consisted of a plain gray raster for observing interference in the luminance channel, and of a solid color field or the color-bar test pattern for observing interference in the chrominance channel;
- The interfering adjacent channels were video-modulated with multi-burst or other appropriate test patterns, or with a red-saturated color field. For some tests, the modulation was simulated by a sinewave, whose amplitude represented the maximum possible value for video modulation (*e.g.*, swing between black-level and white-level).
- Crystal-controlled sync rates were chosen so as to produce slow-moving or stationary interference patterns.

With visibility threshold data determined for "worst conditions", it becomes possible to establish the receiver selectivity requirements which guarantee complete freedom from adjacent-channel interference. These requirements can be determined experimentally with very good accuracy, as mentioned, and can provide the receiver designer with upper-limit data.

All variable parameters (such as picture content) which affect interference visibility under actual viewing conditions, can and must be investigated separately using

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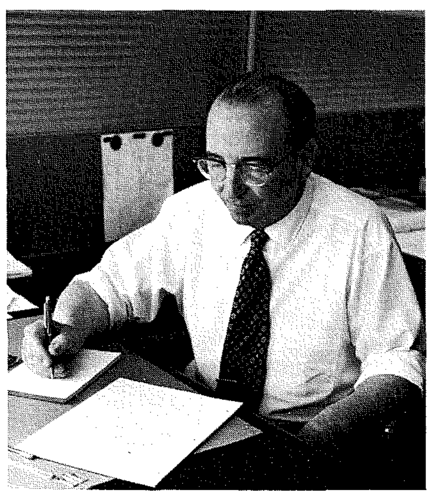


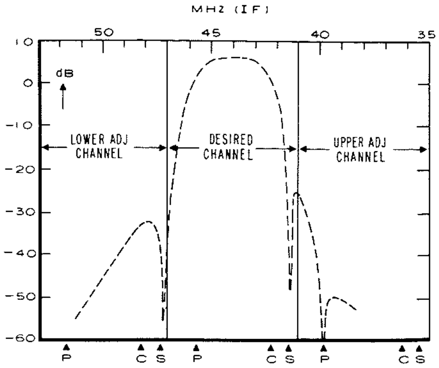


Fig. 1 — Receiver frequency response. P = picture carrier position; C = color carrier positions; S = sound carrier positions.

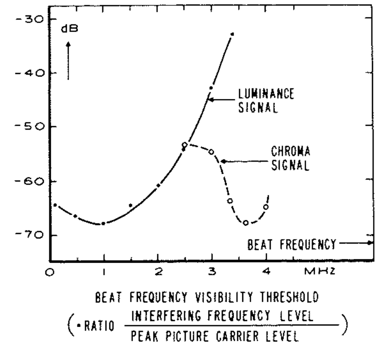


Fig. 2 — Beat frequency visibility threshold as a function of beat frequency. (Beat frequency visibility threshold is defined as the ratio of the interfering frequency level (rms) to the peak picture-carrier level (rms) at the input of the second detector.)

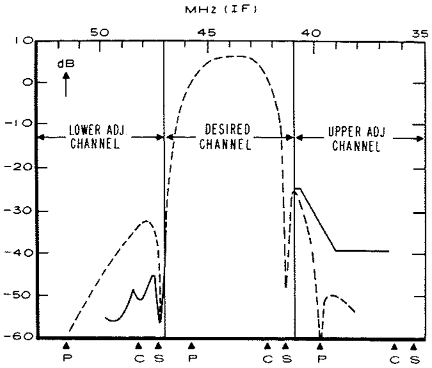


Fig. 3 — Receiver frequency response, rf + i.f. (Dotted line represents actual response; straight line represents required attenuation in adjacent channels to guarantee interference-free operation under "worst-case" consitions.)

subjective viewing tests with a larger group of observers. Properly interpreted, the statistical findings of such subjective tests can then be referenced to the accurately established "worst-case" basic visibility threshold data. This makes it possible to estimate the probability of visible interference with reasonably good accuracy for any given condition.

To obtain a threshold value, the measurement started with a high interfering signal level to acquaint the observer with the type of interference to be rated. The level of the interfering signal was then decreased in steps (minimum step = 1 dB); the interference was removed temporarily between steps. The highest attenuator setting at which the observer could still identify the presence of interference was used to arrive at the visibility threshold value. It was found that this method gave very accurate, repeatable results. The spread between the three participating observers was not greater than 1 dB.

For interference visibility tests, observers were seated at a distance of 4 ft. from the face of the CRT (equal to four times the picture height). The luminance highlight brightness was in the order of 30 to 35 ft lambert, while the room background brightness during tests was held at approximately 0.5 ft lambert.

The peak picture-carrier level in the desired channel at the receiver input was −35 dBm (5 mV across 75 ohm). The sound carrier levels were set at −15 dB relative to the associated peak picture carrier. All carrier levels could be adjusted with a relative accuracy of $\leqslant 0.5$ dB. The measurement accuracy was ±1 dB for attenuation, and ±5 kHz for frequency.

The receiver used for the interference tests was a typical contemporary color set designed for NTSC transmission standards. Its frequency response (rf + i.f.) as measured at the input to the second detector is shown in Fig. 1.

Lower-adjacent-channel interference

Let us now consider first the interference which can be caused by a signal in the lower adjacent channel. With insufficient attenuation of the lower-adjacent-channel signal, interference appears in the desired channel as beats between the desired channel picture carrier and frequency components of the undesired channel. In the past, the lower-adjacent-channel sound carrier had been found to be the main offender compared to which any video sideband components could be ignored. This, however, no longer holds for cable television transmissions which presently operate with the sound carrier at −15 ± 2 dB relative to the associated picture, as compared to values of −2 to −3 dB for tv broadcast transmitters at the time most of the previous tests were made. For cable television transmissions, therefore, interference from lower-adjacent-channel video components can become as critical as that caused by the sound carrier. A quick check of the test receiver with "worst-case" test conditions showed that the depth of the adjacent-channel sound trap was fully adequate to suppress interference. The attenuation for color and video components, however, was found to be too low. The picture carriers in both channels had equal amplitudes for this first test.

The visibility of beat interference depends not only on the amplitude ratio of interfering signal to desired signal, but also on the frequency of the beat interference. Information relating beat interference visibility to beat frequency was needed to establish receiver selectivity requirements.

To obtain the beat visibility threshold as a function of beat frequency, a variable frequency was combined with the "desired" picture. The latter consisted of a plain gray raster for observing the interference effect in the luminance channel, and of the color-bar test pattern for measurements on the chroma channel. The results are shown in Fig. 2. The ordinate represents the beat visibility threshold level, given as the rms amplitude ratio of the interfering signal to the peak picture carrier, measured at the input to the second detector.

It has to be noted that the threshold response as shown is still affected by the receiver's video and chroma amplifier responses; strictly speaking, the response applies only to the receiver type used for the test. For instance, slight peaking of the video response at 1 MHz causes the luminance threshold value to decrease with frequency, reaching a minimum at that frequency. Beat frequencies near the color carrier at 3.58 MHz, of course, appear in the picture as low frequency beats in colored areas.

Having determined the beat visibility threshold as a function of frequency, we now compute from the television transmission standards, the maximum amplitude of color signals and video sideband components that can exist in the rf television channel. With this information, it becomes possible to plot the receiver selectivity required to keep any beat-frequency interference caused by signal in the lower adjacent channel, just at the visibility threshold level. Finally, we have to add a safety factor of 3 dB for video components and 5 dB for the adjacent channel sound carrier, as an allowance for the established tolerances for picture and sound carrier levels in adjacent channels. The frequency response so computed for the lower adjacent-channel range is shown in Fig. 3 as a solid line. Its validity was confirmed by measurements with selected test signals.

For worst-case conditions, the response of the test receiver leads to visible interference for two types of signals in the lower adjacent channel:

1) The color subcarrier, because of its high amplitude for saturated colors, giving a 2.42-MHz beat frequency, and
2) Strong video components in the 2 to 3-MHz range, as they can cause easily visible low frequency beats in the chrominance information.

For normal operating conditions of a CATV system, of course, the presence of interference and its degree of visibility will depend on the picture content in both the interfering and the desired channels. This point is discussed in more detail later.

Upper-adjacent-channel interference

Fig. 3 also shows that the selectivity requirements for the upper adjacent channel are quite different from those for the lower one. The selectivity of our receiver is adequate to suppress any

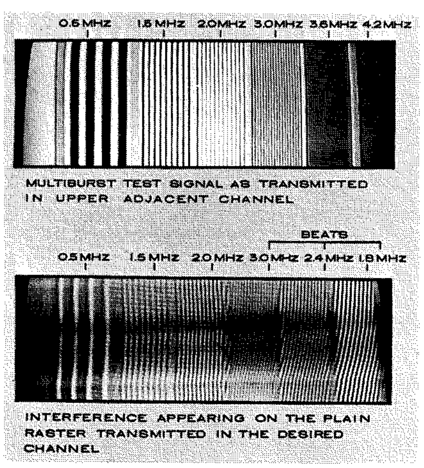


Fig. 4 — a. top) Multi-burst test signal, as transmitted and received in the upper adjacent channel. b. bottom) Interference appearing on the plain raster transmitted in the desired channel.

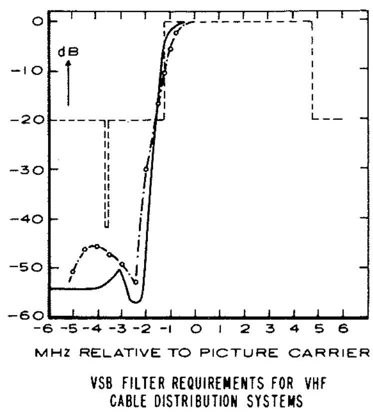


Fig. 5 — Vestigial side band (VSB) filter requirements for vhf cable distribution systems. (Solid line represents suggested minimum requirements, based on interference test results; dash/dot line represents VSB response of upper adjacent channel modulator used for interference tests; dotted line represents FCC specifications for broadcast transmitters).

interference from signals in the upper adjacent channel. There is not even a need for an adjacent-channel picture carrier trap; we actually removed this trap for tests with no ill effects at all.

To explain the less stringent selectivity requirements for the upper adjacent channel, we must consider how signal components transmitted in the upper channel can cause visible interference in the desired one. The mechanism responsible for interference from the lower adjacent channel can be ignored; any beat with the desired channel's picture carrier is higher in frequency than the desired video components and cannot pass the video amplifier.

Two other effects, however, can produce visible interference. The one commonly encountered with insufficient attenuation of the upper adjacent-channel signal is very similar to that of video crosstalk. The unwanted signal appears amplitude-demodulated in the second detector, as a faint but clearly recognizable picture superimposed on the desired one. The interfering picture generally will move slowly across the desired picture due to slight differences between the synchronizing frequencies.

There exists, however, still another form of interference which can be caused by the upper adjacent channel signal. While testing with equal signal levels in both desired and adjacent upper channel, we noticed clearly visible beats with a frequency of 1.8 MHz on the plain raster of the desired channel, when the upper channel was modulated with the multi-burst test pattern. Fig. 4a shows a slice of this pattern as seen in the upper adjacent channel. The 4.2-MHz burst can be hardly seen on the picture tube screen, as it is highly attenuated in the video amplifier of the tv receiver. In the desired channel, on the other hand, it was the 4.2-MHz burst signal that caused the 1.8-MHz beat frequency. We were puzzled at first, but the explanation turned out to be very simple. After deliberately increasing the interference by changing the carrier ratio in the two channels from 1:1 to 20:1, we observed the following interference picture on the plain raster of the desired channel (Fig. 4b). The lower burst frequencies appeared, as expected, in the desired channel similar to crosstalk; the higher burst frequencies, however, appeared as beats of the video sideband components with a frequency displaced 6 MHz from the upper adjacent picture carrier. The beats were the result of insufficient attenuation of the vestigial sideband filter in the upper channel transmitter (in our case a CATV modulator); thus, the lower sideband components of the upper channel signal produced co-channel interference in the desired channel.

We determined that a ratio of approximately 30 dB between desired and upper adjacent picture carriers is sufficient to suppress visible interference. The visibility threshold for lumped video cross-talk, on the other hand, is known to be close to 60 dB.[2] The apparent discrepancy can be explained by a masking effect taking place during demodulation of the two signals.[3,4] Ch. B. Aiken has analyzed the detection of two modulated waves by a linear rectifier. He could show for the case of two carriers with greatly different amplitudes that the weaker signal becomes masked by the stronger one. In a first approximation, a strong carrier reduces the weaker video signal in the demodulator by a factor equal to one half the ratio of weak to strong carrier before demodulation. This masking effect is responsible for the upper adjacent channel's less severe selectivity requirements.

A second type of interference can occur in the presence of saturated colors in the desired channel; in this case, the color subcarrier can produce a 2.42-MHz beat with the picture carrier in the upper adjacent channel. Here again, the picture carrier in the desired channel provides some masking, and the selectivity as shown in Fig. 3 is sufficient to suppress this beat-type interference.

It is quite obvious that the vestigial sideband attenuation requirements for cable-system modulators must be considerably more stringent than those established by the FCC for broadcast transmitters. In Fig. 5 we compare the tv broadcast transmitter specifications (dotted line) with the vestigial sideband response of the modulator we used for the upper channel signal (dash-dotted line). It can be seen that this response peaks for a 4.2-MHz lower sideband frequency. The solid curve finally shows the vestigial sideband response needed to eliminate visible interference for worst-case conditions.

Subjective observer tests

Returning to Fig. 3, the solid lines showing the receiver response needed to meet "worst-case" conditions can be considered the selectivity desired for monitor-type receivers. The designer of home-entertainment receivers, naturally, would like to relax these specifications and now looks for answers to more practical questions, such as:

- What is the probability that we actually shall experience "worst-case" conditions on a CATV system? Or,
- By how much can the selectivity requirements be relaxed before a most critical observer considers the interference objectionable?

To answer questions of this kind, potential interference sources should be separated into two categories:

1) Interference whose level depends on the picture content, and
2) Interference whose level is independent of the picture content.

The latter type, certainly, should be suppressed below the visibility threshold level, as the interference, once visible, would stay visible all the time. The test receiver already meets "worst-case" specifications for this type of interference, *i.e.*, interference caused by the lower adjacent sound carrier and by the upper adjacent picture carrier with sidebands of the synchronizing signal. [Assuming proper alignment of the adjacent-channel sound trap, and adequately tight frequency tolerances for the picture carriers. Obviously, these conditions must be met to avoid visible interference caused by the adjacent-channel sound.] As a matter of fact, so do most contemporary color tv receiver models. We, therefore, need concern ourselves primarily with beat interference caused by color subcarrier and sideband components in the lower adjacent channel.

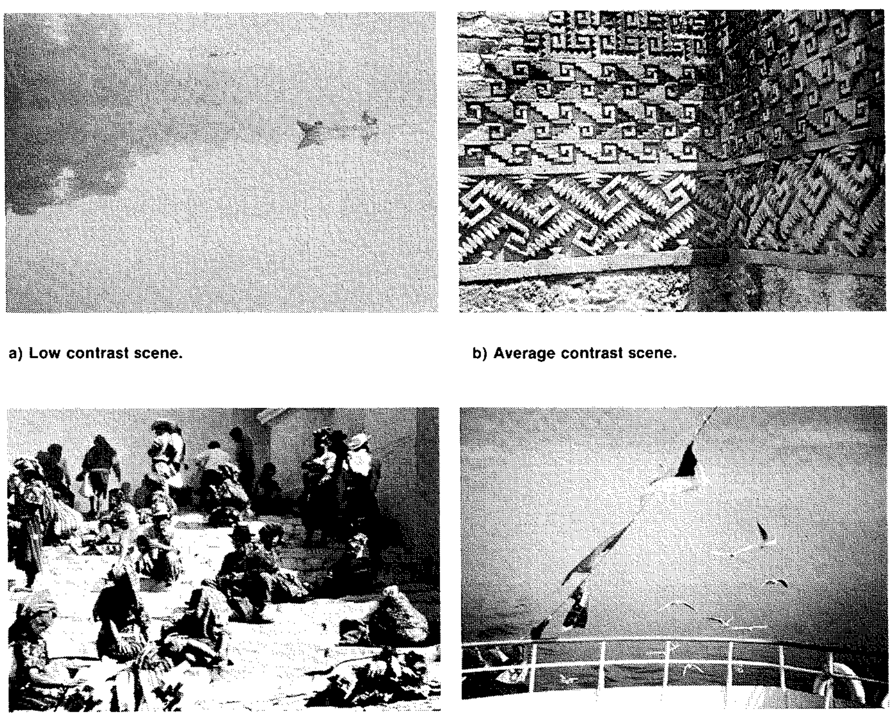
There are four main factors which affect the visibility of beat interference:

1) The *beat frequency* itself (as we have already determined);
2) The *speed* at which the beat pattern moves through the "desired"; picture. Stationary or slow-moving beat patterns are most easily recognized. At fast speeds, the beat visibility threshold increases by typically 7 to 10 dB. Use of this effect is made to suppress co-channel interference by employing precisely off-set carrier frequencies. For the case of adjacent-channel interference experienced on a cable tv distribution system, of course, the speed at which a beat pattern moves is beyond the control the the operator;
3) The *frequency spectrum* of the interfering video signal. The stronger a specific frequency component of the interfering video signal, the more likely it is to cause interference. In practice, the most critical case is that of the color carrier for saturated red or cyan, resulting in a 2.42-MHz beat. The observer, as we have seen, is also very sensitive to beat frequencies falling near the color carrier of the desired tv channel.
4) The effects of the *picture content* on the observer reaction. Such effects can be of a physiological as well as of a psychological nature.

Viewing tests on a small scale were conducted to get a feel for effects of the viewed picture content on observer ratings, since no quantitative information was found in the literature. These viewing tests served, futhermore, the purpose of estimating with limited, yet useful accuracy by what amount the "worst-case" selectivity requirements can be relaxed before the viewer considers the interference objectionable.

Data was obtained from a group of nontechnical observers. The observers were shown slides with various controlled amounts of interference present. The source of interference was the lower adjacent channel color carrier (saturated red color field). This still represented a "worst condition" for the interference source, not normally encountered with tv program material. The observers were asked for each interference level shown to rate the interference as "not visible, "perceptible but not objectionable", or "objectionable".

The pictures to be viewed were chosen as follows. One slide pair was selected for differences in contrast and the size of areas with uniform brightness and color. A second pair was chosen for the fact that the background (a hazy blue sky, in which the interference first became visible, was practically identical in brightness and hue for both pictures, while the foreground showed completely different scenes, Other slides were typical color tv test



a) Low contrast scene.

b) Average contrast scene.

c) Interference in area of low interest.

d) Interference in area of interest.

Fig. 6 — Black-and-white reproductions of two pairs of color slides used in viewing tests.

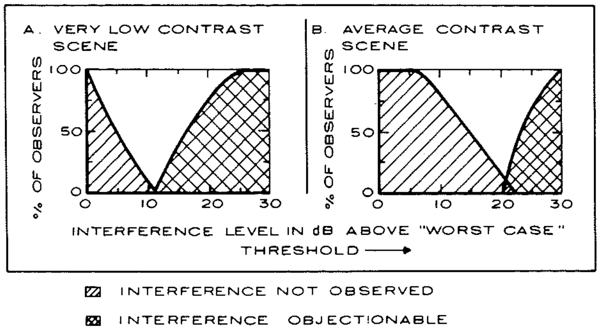


Fig. 7 — Observer ratings for different picture content (Figs. 6a and 6b).

slides, with normal contrast and plenty of color, in which beat interference generally is not so easily detected.

A comparison of some of the test results clearly shows the effects of picture content on the observers' reaction to interference. The first pair of test slides is reproduced in black-and-white as Fig. 6a and 6b. They were selected for their difference in contrast. The first picture (Fig. 6a) has large areas of extremely low contrast and therefore, is quite sensitive to beat interference visibility. The second picture (Fib. 6b) was chosen for average contrast but many details. The observers' reactions are shown as graphs in Fig. 7. It can be seen that the interference ratings for the two pictures differed by approximately 10 dB. The observer ratings for typical color tv test slides were practically the same as those shown for the picture in Fig. 6b.

The difference in ratings between the next pair of test slides (Fig. 6c and 6d) is caused by a psychological effect. The observer is particularly critical of interference appearing in that part of the picture on which he focuses his attention. As mentioned before, brightness and hue of the background in which the interference first becomes visible, is very similar for both test slides. However, the observer's attention is drawn to the foreground of the first of the slides, away from the critical area in the upper left corner, while the action in the second slide (sea gulls) is centered in the sky. Thus, at an interference level of 21-dB above worst-case threshold, none of the observers rated the interference objectionable for the first slide, but 66% of the observers did so for the second one (Fig. 8). It can be expected that, for similar reasons, an observer judges interference less critically for moving pictures than for stills.

Results and conclusions

The combined effect of the factors affecting the observer's reaction to beat interference due to video signals can cause a total spread of experimental subjective ratings in the order of 40 dB, from "worst-case" conditions to a "best-case" combination. One, therefore, has to exercise extreme caution in attempting to derive general conclusions regarding receiver selectivity requirements exclusively from subjective viewing tests.

The results of subjective viewing tests, however, are very helpful when properly used in combination with the data obtained for "worst-case" interference visibility thresholds. With the test material on hand it becomes possible to determine or estimate, respectively, for each of the factors mentioned before, by what amount it statistically increases the threshold level of interference. These factors can be considered to be uncorrelated. Therefore, the probability that certain amounts of interference become visible on the screen of a receiver with marginal selectivity can be computed and presented in percentage of time.

This information, in turn, enables the design engineer to make a knowledgeable decision when weighing cost and complexity of i.f. filters against the possibility of an occasional picture impairment due to adjacent-channel interference.

We have seen (Fig. 3) that the selectivity curve of the test receiver was substantially poorer for the lower adjacent channel than the selectivity needed to guarantee complete freedom from interference for "worst-case" conditions. From our subjective viewing tests, nevertheless, we obtained the estimate that no beat interference should be visible on this receiver for typical television program material transmitted on adjacent channels. The probability of experiencing a case of objectionable interference came out to be 10^{-4} or less; in other words, at an average of a fraction of a second per hour. The actual performance of the receiver, connected to a cable system with 12 active channels, agreed well with his prediction. This result, I have to admit, came as a pleasant surprise to those of us who had seen in the laboratory the objectionable interference visible under "worst-case" conditions. It also proved the feasibility of our method of testing.

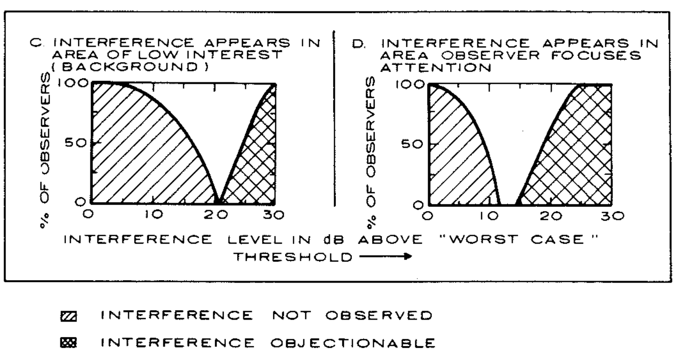


Fig. 8 — Observer ratings for different picture content. (Figs. 6c and 6d).

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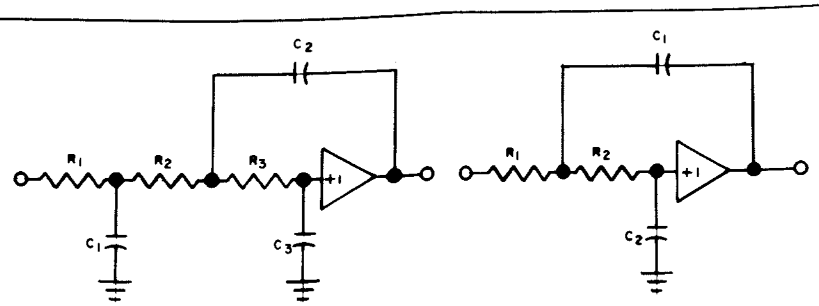
Computer-aided design of active filters

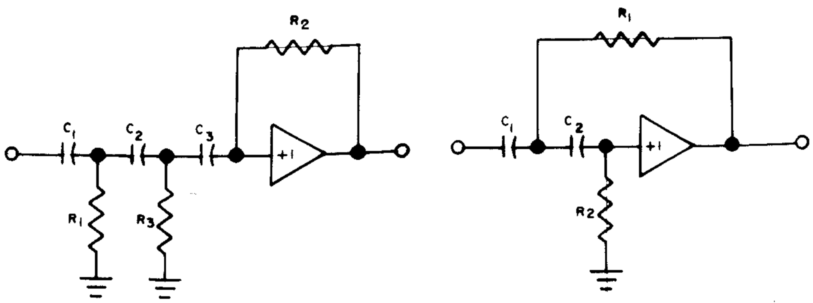
A. Jugs

This paper explains a new iterative approach in a computer program by which any active odd or even pole, lowpass, highpass, or bandpass filter including the standard categories (Butterworth, Chebychev, Bessel, Butterworth-Thompson) may now be designed with standard capacitor values. This eliminates the need for stocking a large variety of nonstandard precision capacitors and greatly simplifies the designing of active filters. The computer program is based on pole locations; therefore, the user has the flexibility to specify his own pole locations for design of other categories of filters. A bandpass filter can be realized by cascading lowpass and highpass sections. Asymmetrical rolloff requirements may be met where the lowpass section has one rolloff and the highpass another.



(a) THREE-POLE (b) TWO-POLE

Fig. 1 — Lowpass filter sections.



(a) THREE-POLE (b) TWO-POLE

Fig. 2 — Highpass filter sections.

FOUR basic configurations[1] used as building blocks to realize the active filters covered in this paper are shown in Fig. 1 and 2.

Even-pole sections are implemented by cascading two-pole sections; odd-pole sections are implemented by cascading a three-pole section with two-pole sections.

Analysis of two-pole section — lowpass

The following discussion shows how any two-pole section may be designed using standard capacitor values.

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The gain transfer function for the two-pole lowpass section shown in Fig. 1 is:

$$A(s) = 1/[R_1C_1R_2C_2s^2 + (R_1+R_2)\,C_2s+1] \qquad (1)$$

The generalized transfer function for a two-pole lowpass filter may be expressed in the form:

$$\begin{aligned} A(s) &= a_0/(s^2 + a_1s+a_0) \\ &= 1/[(s^2/a_0) + (a_1/a_0)\,s + 1] \qquad (2) \end{aligned}$$

where a_0 and a_1 will be determined by the filter characteristics: Butterworth, Chebychev, *etc.* (See the later section on the derivation of active filter coefficients.)

The above normalized equation may be scaled to some cutoff frequency ω_o, then $s=s/\omega_o$

$$A(s) = 1/[(s^2/a_0\omega_0) + (a_1s/a_0\omega_0) + 1] \qquad (3)$$

By equating coefficients in Eqs. 1 and 3:

$$\begin{aligned} 1/a_0\omega^2 &= R_1C_1R_2C_2 \\ a_1/a_0\omega_o &= (R_1+R_2)\,C_2 \qquad (4) \end{aligned}$$

These equations contain four unknowns: C_1, C_2, R_1, R_2. Some restrictions are required for the solution of capacitors and resistors to be positive real.

If $R_1 = R_2 = R$, the capacitors will always be positive real.

The simultaneous solution of Eq. 4 with all R's set equal, will therefore yield:

$$\begin{aligned} C_1 &= 2/(a_1\omega_0R^2) \\ C_2 &= a_1/(2a_0\omega_0R) \qquad (5) \end{aligned}$$

However, the two solutions for the capacitors probably will yield nonstandard values.

To remedy the situation, choose the nearest standard value for C_1. The simultaneous solution of Eq. 4 yields:

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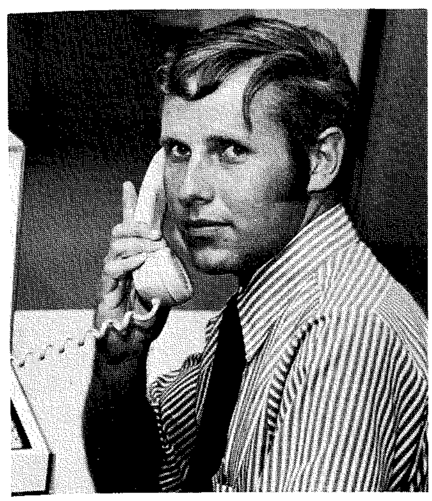

$$R_1, R_2 = (a_1/2\omega_0 C_2 a_0)$$
$$\{1 \pm [1 - (4C_2a_0(C_1a_1^2)]^{1/2}\} \quad (6)$$

A restriction for positive real resistors is that

$$4C_2a_0/C_1a_1^2 < 1$$

or that

$$C_2 < C_1a_1^2/4a_0 \quad (7)$$

with C_1 chosen as a standard value, C_2 may then be chosen as a standard value subject to the above restrictions.

Once C_1 and C_2 have been chosen, R_1 and R_2 may then be computed from Eq. 6 yielding the final solution.

Analysis of three-pole section — lowpass

The first step is to derive capacitor values for a three-pole section. The transfer function for the three-pole lowpass section (Fig. 1) is:

$$A(s) = 1/\ \{R_1R_2R_3C_1C_2C_3s^3 + [C_2C_3(R_1R_3 + R_2R_3) + C_1C_3(R_1R_3 + R_1R_2)]s^2 + [C_3(R_1+R_2+R_3) + R_1C_1s+1] \quad (8)$$

A general normalized transfer function for a three-pole filter may be expressed in the form:

$$A(s) = a_0/(s^3 + a_2s^2 + a_1s + a_0) = 1/[s^3/a_0) + (a_2s^2/a_0) + (a_1s/a_0) + 1 \quad (9)$$

Where the coefficients a_0, a_1, a_2 are determined by the type of filter; *e.g.*, Butterworth, Chebychev, Bessel, (see the derivation active filter characteristics). If Eq. 9 is scaled to some cutoff frequency ω_0, the transformation $s = \delta/\omega_0$ may be made. Then:

$$A(s) = 1/\ [(s^3/a_0\omega_0^3) + (a_2s^2/a_0\omega_0^2) + (a_1s/a_0\omega_0) + 1] \quad (10)$$

If the coefficients of Eqs. 8 and 10 are equated, then:

$$1/a_0\omega_0^3 = R_1R_2R_3C_1C_2C_3 \quad (11)$$

$$a_2/a_0\omega_0^2 = C_2C_3\ (R_1R_3 + R_2R_3)\ (R_1R_3 + R_1R_2) \quad (12)$$

$$a_1/a_0\omega_0 = C_3\ (R_1 + R_2R_3) + R_1C_1 \quad (13)$$

To solve for C_1, C_2, C_3 some restrictions are required for R_1, R_2, R_3 since many solutions for C_1, C_2, C_3 are negative or imaginary. If all the R's are set equal, C_1, C_2, C_3 will always be positive.[2] Therefore, setting $R_1 = R_2 = R_3 = R$ in Eqs. 11, 12 and 13, the solutions for C_1, C_2, and C_3 are of the following form.

$$C_1^3(\omega_0^3R^3) - C_1^2(a_1\omega_0^2R^2/a_0) + C_1(1.5\omega_0Ra_2/a_0) - (3/a_0) = 0 \quad (14)$$

$$C_3 = (a_1/3a_0\omega_0R) - (C_1/3) \quad (15)$$

$$C_2 = (a_2/2a_0C_3\omega_0^2R^2) - C_1 \quad (16)$$

After solving the cubic equation for C_1 and finding the real root, C_2 and C_3 may be found. The Newton-Raphson method (last section) for one variable lends itself nicely to solving the cubic equation. However, C_1, C_2, and C_3 will probably be of nonstandard values.

To remedy the situation, C_1, C_2 and C_3 are chosen in the next step as the nearest available standard values from a table.

Eqs. 11, 12, and 13 are then solved for R_1, R_2, and R_3, knowing C_1, C_2, and C_3. However, the simultaneous solution of the three nonlinear equations is difficult if not impossible using explicit methods.[3] By using an implicit method (Newton-Raphson[4], described in the last section) to solve simultaneous nonlinear equations, R_1, R_2, and R_3 may be found.

Examining Eqs. 11, 12, and 13 and using the same notation as in the last section, let:

$$A = R_1R_2R_3C_1C_2C_3 - 1/a_0\omega_0^3 \quad (17)$$

$$B = C_2C_3(R_1R_3 + R_2R_3) + C_1C_3(R_1R_3 + R_1R_2) - a_2/a_0\omega_0^2 \quad (18)$$

$$C = C_3(R_1 + R_2 + R_3) - a_1/a_0\omega_0 + R_1C_1 \quad (19)$$

ΔR_1, ΔR_2, ΔR_3 may be computed using the formulas found in the last section.

Lowpass-to-highpass transformation

Highpass filter sections may be implemented as in Fig. 2 using a low-pass-to-high-pass transformation where each C is replaced by $1/R$ and each R by $1/C$ and s by $1/s$ (ω_0 by $1/\omega_0$).

The program

The design of active filters may be mechanized in a computer program (see Fig. 3) using Basic, Fortran, or any other standard programming language.

For a two-pole case, see Fig. 4; for a three-pole case, Fig. 5.

All capacitors can be chosen to be equal to one standard value for a realizable solution. (This is the inverse of choosing all resistors equal in the lowpass case.)

Fig. 6 shows a sample run of an interactive filter design program, developed by RCA, Palm Beach, Florida.

User-supplied information is circled. The total terminal time used was 4 minutes; CPU time used was 2.16 minutes on an RCA Spectra 70/45.

Deriving the active filter coefficients

The filter characteristics (Butterworth, Chebychev, Bessel) are determined by the location of the poles in the complex frequency plane. These pole locations in turn lead to the filter coefficients a_0, a_1, and a_2 which are then used in the design of the filter.

The most useful pole locations are those normalized to a radian cutoff frequency of $\omega_0 = 1$ at which the gain is -3dB ($|A| = 2^{-1/2}$). These normalized pole locations[5] may be derived analytically or may be stored in a table for recall in the computer program.

Two-pole sections

Each pair of complex conjugate poles corresponds to a two-pole section. Since no impedance loading exists between sections, each section can be individually magnitude scaled.

A pair of normalized complex conjugate poles may be expressed as: $s_1 = sa$ and $s_2 = sa^*$, where sa^* if the conjugate of sa.

$$(s - sa)(s - sa^*) = s^2 + a_1s + a_o$$

Then the normalized filter transfer function may be expressed as:

$$A(s) = a_0/(s^2 + a_1 + a_0)$$

Three-pole sections

A filter with an odd number of poles requires one three-pole section. This section is then cascaded with two-pole sections to yield the desired number of odd poles.

A three-pole section requires a real root in addition to the pair of complex conjugate roots. Since only one real root may exist per filter, only one three-pole section per filter may be realized.

A pair of normalized complex conjugate poles, plus a real pole may be expressed as:

$$s_1 = sa$$
$$s_2 = sa^*$$
$$s_3 = sb$$

and expanded

$$(s - sa)(s - sa^*)(s - sb) = s^3 + a_2 s^2 + a_1 s + a_0$$

A three-pole normalized transfer function may be expressed as:

$$A(s) = a_0 / (s^3 + a_2 s^2 + a_1 s + a_0)$$

where at dc, $A = 1$ and $at/s = j1, |A| = 2^{-1/2}$

Order of sections

The order in which filter sections are placed is important. The first section should have the most gradual rolloff with each following section having a steeper rolloff.

The overall response of the filter may be flat in the passband with unity gain, but a particular section with a steep rolloff, examined by itself, will have a gain. (for example, in a seven-pole Butterworth filter, the steepest two-pole section has a gain of 8.) The result could, therefore, be an overdriven amplifier.

The location of each conjugate pole pair with respect to the imaginary axis determines the steepness of rolloff of that section. Therefore, pole pairs furthest from the imaginary axis should be realized and placed first with the corresponding sections following.

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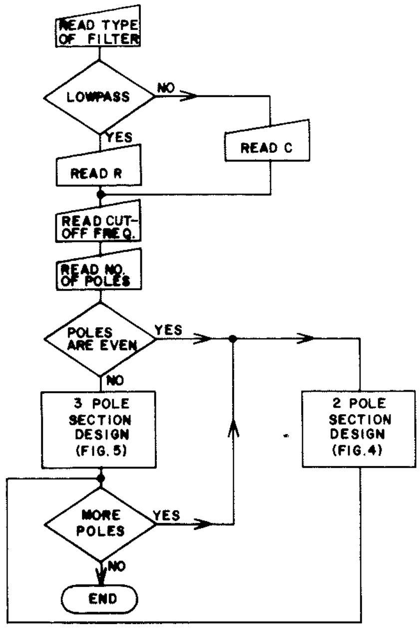


Fig. 3 — Flowchart for composite design of two, three or more poles.

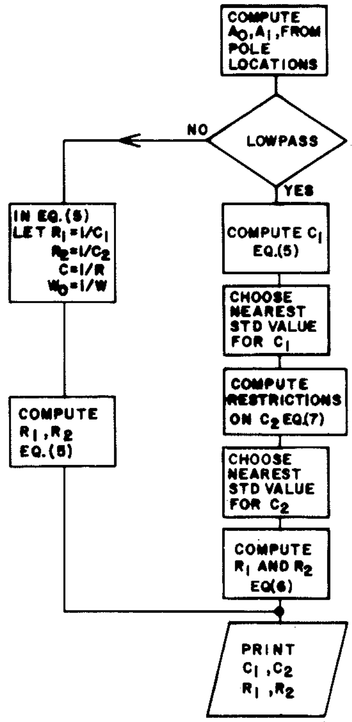


Fig. 4 — Two-pole section design.

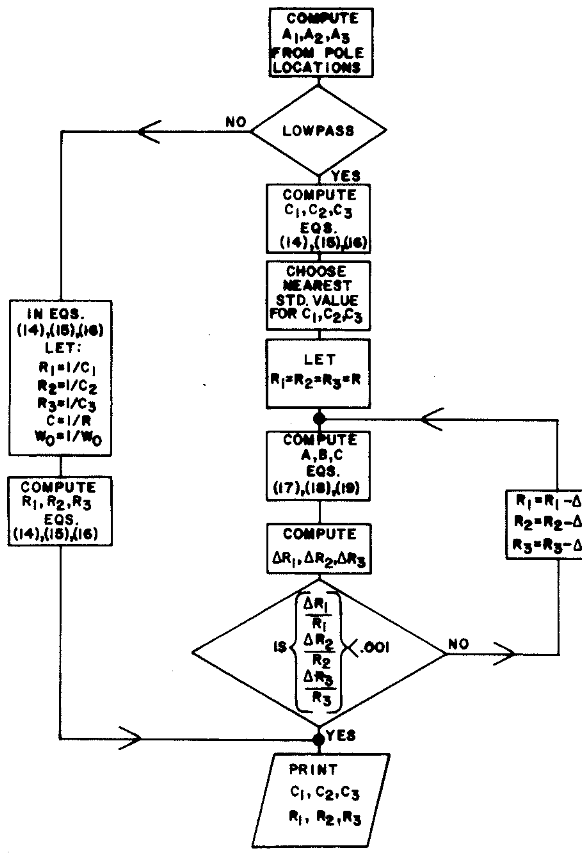


Fig. 5 — Three-pole section design.

Viterbi decoding

D. Hampel | K. Prost

CMOS Viterbi decoders can be designed with state-of-the-art technology, representing more efficient implementations than previous approaches. First a description of the convolutional encoder is given, followed by an overview of Viterbi decoder operation. This is followed by a more detailed description of the design of the major functions in a decoder, and a comparison of alternate implementations based on CMOS LSI. Finally, a description and test results of a feasibility model of a Viterbi decoder built with standard CMOS parts is given.

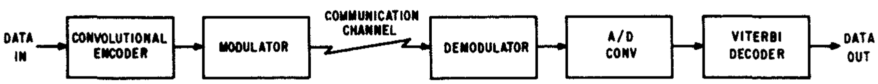
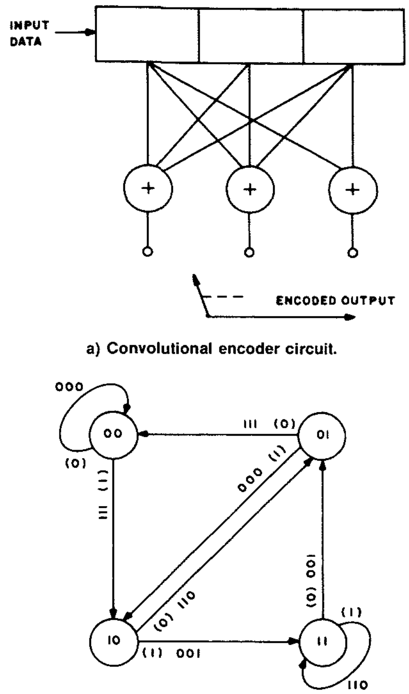


Fig. 1 — Convolutional encoder and Viterbi decoder in a conventional communication link.

CONVOLUTIONAL encoding and decoding schemes have been developed in an effort to provide improved performance in communication links. Lower bit-error rates (BER) are achievable at a given signal-to-noise (denoted as energy per bit over the single-sided noise spectral density, E_b/N_o), for a given modulation scheme, while using more bandwidth. Such codes have been shown to be superior to block codes for many applications.[1]

Several practical sequential decoding algorithms for decoding convolutional codes have been developed.[2,3] The sequential decoder examines the error-corrupted quantized received sequence and chooses the data sequence corresponding to a transmitted code sequence which has a high (but not the greatest) probability of being transmitted. Thus, it is a sub-optimum decoder. The Viterbi decoder offers an alternate approach.

The convolutional encoder and Viterbi decoder relate to an overall communication link as shown in Fig. 1. Before the data are prepared for transmission (generally with phase-shift-keying modulation, for best performance), it is encoded with, typically, two or three bits for each data bit. At the receiver end, the demodulator prepares the incoming, noise-corrupted signals for the decoder. For optimum performance, the demodulator incorporates a matched filter and an A/D converter. The latter provides soft decision, or quantized data, which the decoder can process to provide improved gain compared to hard decision data.

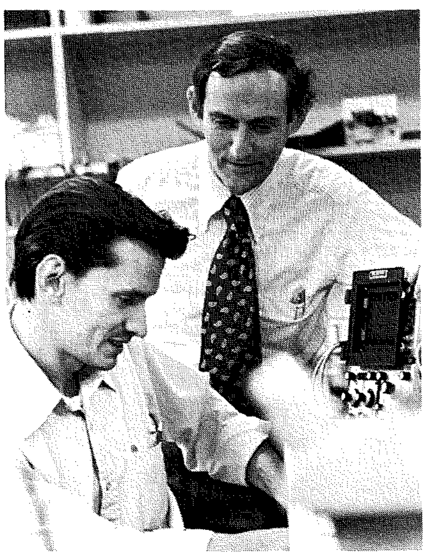


a) Convolutional encoder circuit.

b) State diagram.

Fig. 2 — Convolution encoder operation.

The Viterbi decoder examines the error-corrupted quantized received sequence and chooses the data sequence corresponding to the transmitted code sequence which has the greatest probability of being transmitted, *i.e.*, the most likely sequence. In so doing, it is limited to relatively small constraint lengths (≈ 8 or less) because (as will be seen) of the exponential increase in hardware with constraint length. However, the Viterbi decoder operation is independent of the information rate (as long as the logic circuits are fast enough for the highest



Authors Prost (left) and Hampel.

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K. Prost, Advanced Communications Laboratory, Communications Systems Division, Somerville, NJ, studied at RCA Institutes (1947 and 1958); Hofstra University and CCNY (1958 to 1962)' and Newark College of Engineering (1968 to 1972). Mr. Prost joined RCA in 1966 and was assigned to the Advanced Digital Techniques Group. Since joining the group he has been involved in work on integrated threshold gates, logic design, and radiation hardening. In these areas Mr. Prost has assumed responsibility for design and testing. He is now performing logic designs for modems. From 1963 to 1965 he was with IBM where he wrote test procedures for automatic checkout of digital circuits, and initiated changes for existing test equipment. From 1958 to 1963, Mr. Prost worked for the Arma Corporation in Long Island. During this time he worked on System Test, Quality Control, and Field Engineering on the Inertial Guidance System for the Atlas Missile. Mr. Prost has co-authored three papers in the digital techniques area.

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rate). Sequential decoders, although they can accomodate long constraint lengths ($\approx$ 40 or more) are also limited in information rate due to the finite probability that the number of operations required to decode a particular bit can be quite large, causing input buffers to overflow. Comparisons[4] have been made to show the relative advantages of sequential and Viterbi decoding for given situations.

The Viterbi decoding algorithm, then, can be viewed as a method for achieving many of the advantages to be gained for decoding convolutional encoded data with manageable hardware requirements at high data rates.

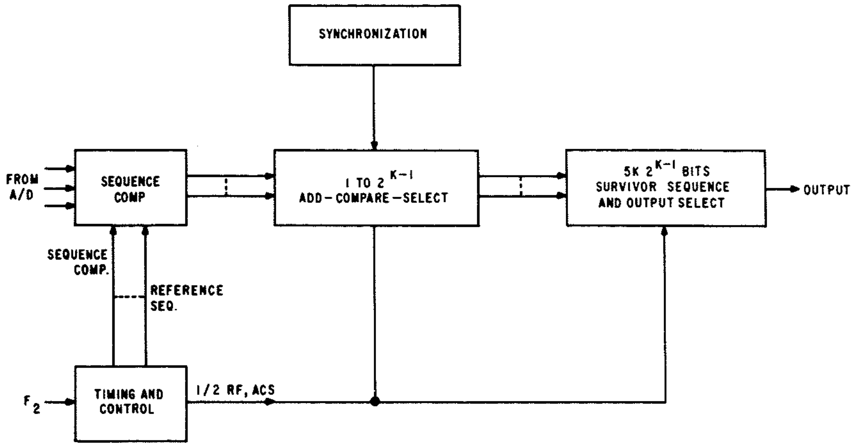


Fig. 4 — Viterbi decoder block diagram.

Simulations[5] have shown that Viterbi decoding systems can provide up to 4 to 5 dB of signal-to-noise improvement at bit error rates in the order of 10^{-4}, for relatively simple hardware. The power dissipation of the decoding circuitry and its speed ultimately determine system performance improvement.

Convolutional encoding

A convolutional encoder simply consists of a K-stage shift register, with taps, feeding V modulo-2 adders as shown in Fig. 2a. For each input bit to the shift register, there are V encoded output bits. The codes rate is R and is defined as

$$R = 1/V$$

and indicates the number of information bits/transmitted bit (the reciprocal of the bandwidth expansion). The length of the shift register, K, is the constraint length, and indicates for how many bit periods an input bit influences the output sequence. Each of the V modulo-2 adders can have inputs from 1 up to K shift register taps. The particular taps used on the shift register to the modulo-2 adders define the code.

It has been shown by analysis and simulations that particular codes are superior[4] for given constraint lengths and code rates and that others could result in catastrophic error propagation.

The operation of a convolutional encoder, which is useful in later describing Viterbi decoding, can be viewed with the aid of a state diagram. The K=3, R=1/3 encoder as shown in Fig. 2a, would have a diagram as shown in Fig. 2b. There are 2^{K-1} possible states for the $K-1$ stages of the encoder shift register shown by the circles. From any state, an input bit will cause the register to assume any one of two other states, shown by the bits in parentheses. Hence, any state can be entered from any one of two input states. This pattern is the same, no matter what the code rate or the taps are. Now, the message transmitted during each transition between two states is code dependent and is labeled along each transition path. For example, if the last two bits in the shift register were 10, after a 0 was shifted in, the outputs of the three modulo-2 adders feeding the output (of Fig. 2a) would be

first adder $0 \oplus 1 \oplus 0 = 1$
second adder $0 \oplus 1 \oplus 0 = 1$
third adder $0 \oplus 0 = 0$

The output message sequence for any input data sequence can be determined from a modified state diagram by arranging the possible states in columns and labeling all the transition sequences; the upper line emanating from a state corresponding to a 0 input and the lower, a 1 input. This is shown in Fig. 3.

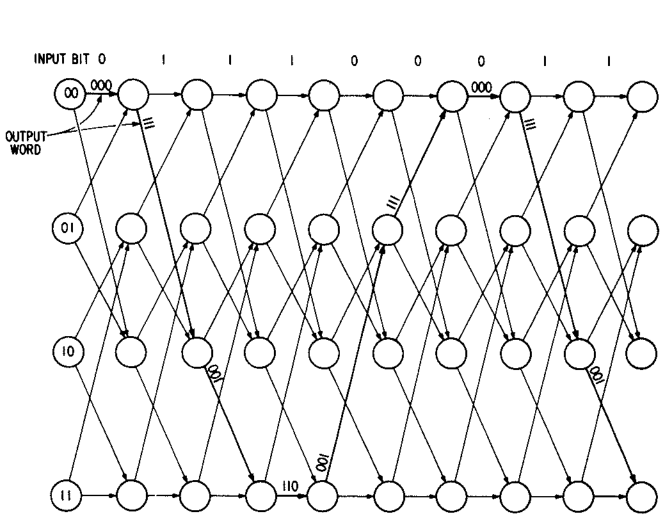


Fig. 3 — Trellis structure for determining output message of convolutional encoder.

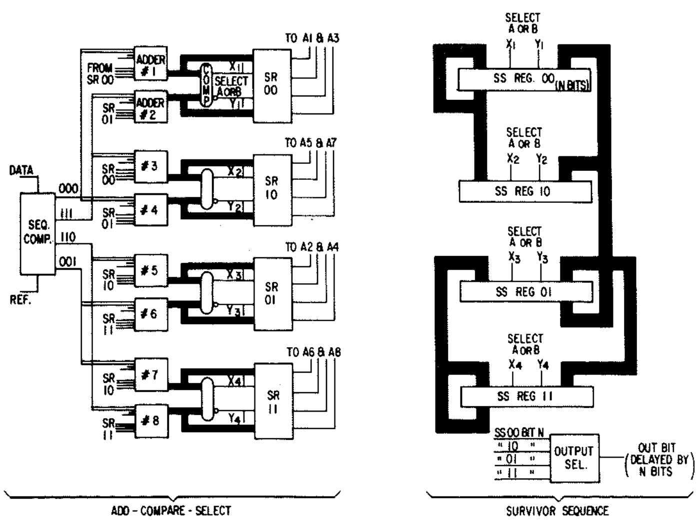


Fig. 5 — Functional diagram of Viterbi decoder.

Another way of interpreting the operation of the convolutional encoder is to view the first column of the diagram corresponding to time slot 0 and the second column, time slot 1, *etc.* As time progresses, V information bits are transmitted for each time slot transition (and state sequence) as denoted by the allowable bit patterns, for the particular code taps.

Viterbi decoding description

Viterbi decoding represents a methodical approach for choosing the most likely input data stream from a noise-corrupted transmitted (and received) stream. It does this by measuring the discrepancies between successive input "words" and all possible input words emanating from, and merging at, each state of the expanded state diagram and maintaining a record of accumulated "errors" as well as the responsible data sequences. The following discussion gives the detailed operation of a Viterbi decoder, with respect to its block diagram, Fig. 4, and its functional diagram for a K=3 system, Fig. 5, for the particular code exemplified in Figs. 2 and 3.

Sequence comparator logic

First, each input bit data stream is compared with each possible input data stream. For V=3, there would be a maximum of 8; certain codes will result in fewer possibilities. This comparison consists of determining the number of bit positions which differ between the input and each possible stream. This value, ΔS or state metric, can vary from 0 (if they agree) to V (if they all disagree). The circuit that does this is referred to as the sequence comparator. The implementation consists of an *exclusive-or* gate for each allowable sequence; one input is the transmitted data stream and the other input is one of the locally generated, possible sequences. Disagreements are gated into V-state counters to register each ΔS for subsequent processing.

Add/compare/select logic

The outputs of the sequence comparator feed the add/compare/select unit which keeps a running total of the state metrics and makes decisions after each summation to store the lowest total state metric of each pair terminating in each state as shown.

For example, for each permissible state there are two entry sequences. There are two corresponding adders which accept the branch metrics generated by these sequences (in the sequence comparator) and provide total metrics for that state. These two sums are compared, and the lowest one is selected and gated into a score register. Each of these registers (one for each state) is connected back to one of two adders of each state for subsequent addition with the new ΔS's. An expanded state diagram is shown in Fig. 6 for a given input sequnce, with the total metrics shown along each state after each V-bit input sequence. After each add/compare/select operation, there are 2^{K-1} possible paths leading to the next operation. In those cases when the metrics entering a state are equal, either one can be selected.

The running totals, or metrics, enable the determination of the transmitted bit sequence. The adder and score register capacity can be limited since the difference between the minimum and maximum metrics is bounded. This difference is dependent upon the code rate as well as on whether the Viterbi decoder is fed by hard or soft decisions data (to be described later). In implementation, the adders are monitored and when a specific minimum value is reached, by all of them, a constant is subtracted from all adders during the next operation.

Survivor sequence logic

The next and final function of the Viterbi decoder is that of determining the most likely information bit from the add/compare/select data. To do this, a set of survivor-sequence registers is provided (one for each state), each of which is loaded with a 1 or 0 bit depending upon which path led to the lowest metric at each state. These registers are loaded and information is transferred among them. For example, after the first sequence is received at t=1 (Table I), the survivor sequence for the 00 state (SS_{00}) is 0, SS_{01} is 0, SS_{10} is 1 and SS_{11} is 1. After t=2, since state 01 led to state 00, by virtue of a 1 being transmitted, the contents of SS_{00} will consist of the contents of SS_{01} (0), followed by a 1. Similarly the contents of SS_{01} will consist of the contents of $SS_{10}{}^{1}$, followed by a 0, *etc.* Table I shows the contents of each sequence comparator and survivor-sequence register for a par-

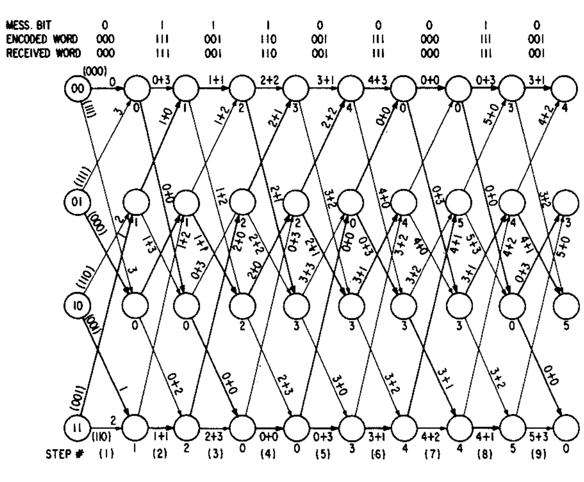


Fig. 6 — Decoder's path through trellis structure (no noise).

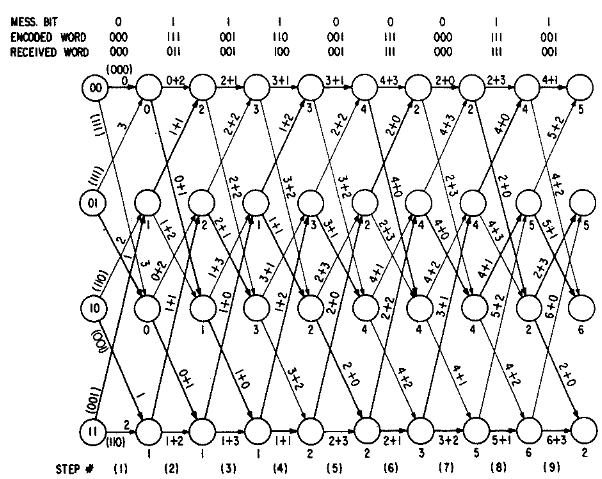


Fig. 7 — Decoder's path through trellis structure (noise).

Table I — Survivor sequence registers (no noise).

Step No.	*From reg.*	SS_{00}	*From reg.*	SS_{01}	*From reg.*	SS_{10}	*From reg.*	SS_{11}
1		0		0		1		1
2	01	00	10	10	00	01	11	11
3	00	000	11	110	01	101	10	011
4	01	1100	10	1010	00	0001	11	0111
5	01	10100	11	01110	01	10101	11	01111
6	01	011100	10	101010	01	011101	11	011111
7	00	0111000	11	0111110	00	0111001	10	0111011
8	00	01110000	10	01110010	00	01110001	11	01110111
9	00	011100000	10	011100010	01	011100101	11	011100011
10								

ticular input sequence.

Another example is shown in Fig. 7 and Table II, with error bits added to the input sequence. Whereas in the error-free case all survivor-sequence registers agreed after a couple of periods (after initialization), it is seen that more time is required for the corrupted input sequence to agree. Simulations have shown that with survivor-sequence register lengths of 5k, all registers will, with high probability, agree on the initial few bits of the survivor sequences, with a moderate number of received errors. After an initial delay of 5k, outputs bits are selected from the survivor-sequence registers, one per V-bit input sequence. The output from the survivor-sequence register whose associated sequence-comparator register has the lowest score at the time can be chosen for an optimum decision in a high noise environment. Alternately, an output which agrees with the majority of the survivor-sequence registers can be chosen. The former method requires minimum hardware.

Soft decision

Simulations[5] have shown that 3-bit quantized (8 level) soft decisions provide about 2 dB of additional coding gain (compared to hard decision). Futhermore, for a given constraint length and code rate, higher-order quantization yields relatively little additional gain. Since the hardware requirements for both the A/D, and the added amount for the Viterbi decoder to handle the quantized bit stream, are not very involved, soft decision is recommended. The block diagram of the decoder of Fig. 4 showed soft decision interface. The major impact is on the add/compare/select capacity — 3 bit inputs instead of 1 result in larger state metrics and hence larger sums and scores to process.

Synchronization

Along with the data, a $2f$ clock is supplied to the decoder. This $2f$ line is then divided by twice the basic clock frequency and all necessary control lines are generated from this. The only synchronization requirement is that the decoder is able to detect which bit or groups of bits (if soft decision is used) represents the first bit in each word.

As was mentioned previously, if there is noise on the incoming data stream, the metrics for all states will increase. If the decoder is out of synchronization, it has the same effect as noise would have on the system. By monitoring the rate of increase of all metrics over a fixed period of time one can determine if the system is synchronized. Once a decision has been made that the decoder is out of synchronization, the timing pulse can be made to skip, effectively slipping the data stream by one bit. This will continue until the decoder is locked on the proper bit. A tradeoff exists regarding the period of time over which the metric increase is monitored. If this period is made too short, then noise will cause the decoder to lose synchronization frequently, while on the other hand, if it is made too long it will take the decoder long to lock on, resulting in an increase of output errors.

Table II — Survivor sequence registers (noise).

Step No.	*From reg.*	SS_{00}	*From reg.*	SS_{01}	*From reg.*	SS_{10}	*From reg*	SS_{11}
1		0		0		1		1
2	01	00	11	10	00	01	10	11
3	00	000	11	110	01	101	10	011
4	01	1100	11	0110	01	1101	11	0111
5	00	11000	11	01110	01	01101	10	11011
6	01	011100	11	110110	00	110001	11	110111
7	00	0111000	11	1101110	01	1101101	11	1101111
8	01	11011100	10	11011010	00	01110001	11	11011111
9	00	110111000	10	011100010	01	110110101	10	011100011
10								

Implementation

The three major functions to be performed by the decoder are: 1) determination of the branch metric; 2) determination of the new state metric; 3) generation of the survivor sequence. The hardware implementation will depend upon the constraint length (K), the code rate (R), and also if hard decision (Q=2) or soft decision (Q=8) is used on the incoming data stream. Other factors that ultimately affect hardware complexity are the data rate and the power and size constraints imposed by the system. If the data rate is high (with respect to the type logic being used) then an all-parallel approach has to be taken, resulting in a maximum amount of hardware. On the other hand, for low-data-rate systems, a serial or serial-parallel implementation can be used, reducing gate count. The complexity of the first of the three major functions, the input sequence comparator, or the branch metric section, is mainly affected by the code rate, and whether hard or soft decision is used. For each unit increase in the code rate there is an extra bit to be processed by the sequence comparator. For soft decision (Q=8), where each incoming bit is effectively represented by three bits, the sequence-comparator capacity must be increased accordingly. A change in K has no effect on the sequence comparator, since this unit only operates on the incoming data and is not required to maintain any history of operation.

The add/compare/select and the surviorsequence logic complexity are directly related to K, and for every increase in K the hardware increases exponentially in a parallel organization. The code rate and decision process does not affect the survivor-sequence logic and only moderately affects the add/compare/select logic. As more bits are processed due either to a change in the data rate or soft decision, the number of errors that each state can have increases. Therefore, the bit capacity in each add/compare/select unit will have to increase.

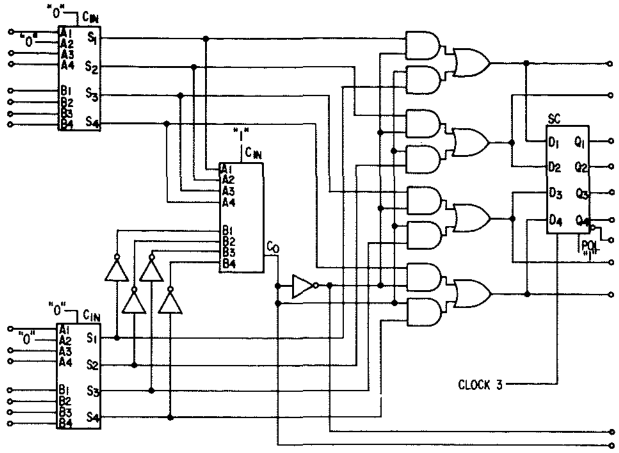


Fig. 8 — Add and compare logic.

Parallel organization

Add/compare/select logic

An add/compare/select unit implemented with CMOS logic is shown in Fig. 8. In an all-parallel organization, 2^{K-1} identical units would be required. Basically, each unit consists of two adders and a comparator which determines the adder with the lowest number, and a multiplexer, which (in conjunction with the comparator outputs) selects the adder with the lowest number and gates it into the register. The register holds the value for use when calculating the next metric. This output value represents the normalized errors accumulated for that state since the start of transmission.

Of the two sets of inputs to each adder, one group represents the state metrics from other states while the second group of inputs is from the sequence comparator containing the new branch metric. These inputs depend upon the code taps in the encoder and are changeable. Table III lists the state transfers for a K=7 decoder. For example, to enter state 25, the previous state must have been either 12 or 44. It should also be noted that odd-number states indicate that the latest data bit into the encoder shift register must be a 1, while even states indicate a 0 bit input. Therefore, state 12 can either go to state 24 or state 25, depending upon the data bit.

During operation of the system, each adder calculates the new state metric based upon the previous state metric and the new branch metric. The object is to determine which one has the lowest metric, since the state with the lowest metric is the one that has the highest probability of having been transmitted. Once the sum outputs are avialable , a 2's complement addition is performed between the two sums in a third adder. This adder functions as a binary comparator, the carry output line conveying the comparator's results. If C_{out} is high, then $A \leqslant B$; if C_{out} is low, $B<A$. The C_{out} line is the select input to the multiplexer and gates the lowest metric into the register.

In a parallel system, the delay in the add/compare/select block represents the longest single delay path in the decoder and hence dictates maximum data rates. In terms of gate delays, it can be assumed that each adder is 6 bits long and (for K=7) that alternate stages of carry lookahead are used. If, further, it is assumed that there is one gate delay for the carry and two delays for the sum, then the 6-bit sum would be available in a maximum of four gate delays in a custom circuit. As the sum outputs become available, the comparator block is performing its operation; therefore there is one additional gate delay contributed by the comparator plus the delay of an inverter. This brings the total to six delays, with a maximum of two delays for the multiplexer block. Based on the type of logic used, then, one can estimate the maximum data rate. For example, if a CMOS gate delay is approximated at 40 ns, then a maximum data rate of about 4 Mb/s can be accommodated for a custom design.

Table III — State transfer for K=7 decoder.

New state	Original state		New state	Original state	
0	0	32	32	16	48
1	0	32	33	16	48
2	1	33	34	17	49
3	1	33	35	17	49
4	2	34	36	18	50
5	2	34	37	18	50
6	3	35	38	19	51
7	3	35	39	19	51
8	4	36	40	20	52
9	4	36	41	20	52
10	5	37	42	21	53
11	5	37	43	21	53
12	6	38	44	22	54
13	6	38	45	22	54
14	7	39	46	23	55
15	7	39	47	23	55
16	8	40	48	24	56
17	8	40	49	24	56
18	9	41	50	25	57
19	9	41	51	25	57
20	10	42	52	26	58
21	10	42	53	26	58
22	11	43	54	27	59
23	11	43	55	27	59
24	12	44	56	28	60
25	12	44	57	28	60
26	13	45	58	29	61
27	13	45	59	29	61
28	14	46	60	30	62
29	14	46	61	30	62
30	15	47	62	31	63
31	15	47	63	31	63

Add/compare/select capacity

As was mentioned previously, the add/compare/select units keep a continous record of all errors associated with each state, and if there was no noise, one state would always have zero errors. The errors for the other states would range up to some finite value, depending upon the code rate and also on hard or soft decision. This range from zero to some maximum is fixed for a given code rate and the type of processing on the incoming data. Once noise is introduced into the system, the spread between the minimum and maximum errors remains the same, but the absolute value of both increases as the amount of noise increases. Therefore, adders of finite length could overflow resulting in errors by the decoder. One method of preventing the overflow is to monitor the outputs of all add/compare/select units, and when all units have reached some minimum value, subtract this value from *all* add/compare/select units. The length of the adders would have to be such that when all units are at or above the minimum, the maximum error state does not overflow. By subtracting this minimum value from all units, the absolute value of errors is changed but the relative difference between each state remains the same.

The second major function performed by the decoder is to generate survivor sequences for each state of the system. Computer simulations have shown that survivor sequences of about $5K$ in length are sufficient. Therefore, as K increases, the number of survivor-sequence states also increases, as does the number of bits per state.

Survivor-sequence logic

In a parallel system, the survivor-sequence logic is built using parallel-in-parallel-out (PIPO) shift registers as shown in Fig. 9. On each input line, there is a two-input multiplexer with the select line for the multiplexer being the comparator output of the appropriate add/compare/select unit. Since the even-numbered states can only be entered if the data bit is a "zero", the LSB of all even states is "hard wired" to a low while the LSB of all odd states is "hard wired" to a high.

The function of the survivor-sequence logic is to reproduce the data stream that is entering the encoder shift register. Initially, all bits of the survivor-sequence memory are set to a zero state, and after the first data bit has been processed, the LSB's of all odd-numbered states are set high, while the even-numbered states are set low. After the next bit, add/compare/select units would have selected the most likely path for each state and the comparator outputs will then select the appropriate transfer for the survivor-sequence register. For example, the register holding the survivor sequence for state 25 has at its inputs terminals the survivor sequences for states 12 and 44 (shifted by one bit position). Now, if the add/compare/select unit for state 25 selects state 12 as the best metric, the comparator output line will select survivor-sequence-12 data for the inputs to survivor sequence 12. Therefore, all 2^{K-1} survivor sequences are shifted one bit position and transferred at the same time. Once the comparator has made a decision, the thruput delay for the survivor-sequence logic is the same as for the add/compare/select unit; a multiplexer delay and delay of the flip-flop.

One of the major drawbacks to the parallel system is the number of I/O lines that are required. For K=7, there are 64, 32-bit shift registers required and each register has to have all of its output bits available — as well as two input lines for each bit. This is a total of 96 I/O pins plus the necessary control lines, severly reducing the number of such bits that can be placed on a single custom chip for a parallel organized survivor sequence.

Serial organization

If the data rate was low enough, a serial system could be used. A completely serial implementation would require only one add/compare/select unit along with a multiplexer for all of the inputs, and a memory (64 words $\times$ n bits for K=7, where n is the adder size) large enough to hold all of the state metrics. This is the minimum amount of storage required to hold all metric values but insufficient to operate the system. As new state metrics become available, they cannot be entered into memory because the original metric for the state may be required at a later point in the calculations. One way of eliminating this is to double the memory size and use half of the memory to hold previous state metrics while the second half of the memory is being filled with the new metric values. At the end of each cycle, the function of each memory half is reversed. One advantage of this method is that the metric for any given state can always be found in the same memory location. During the time that the add/compare/select unit is calculating the metric, memory transfers have to be set up to obtain the necessary state

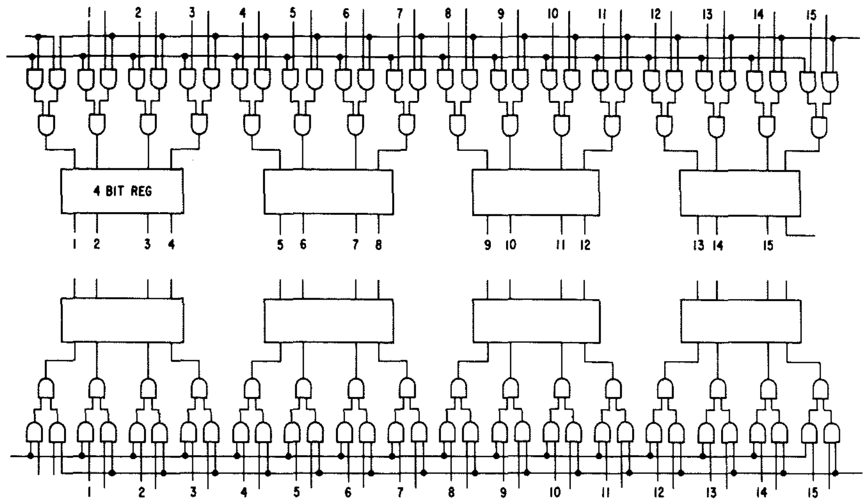
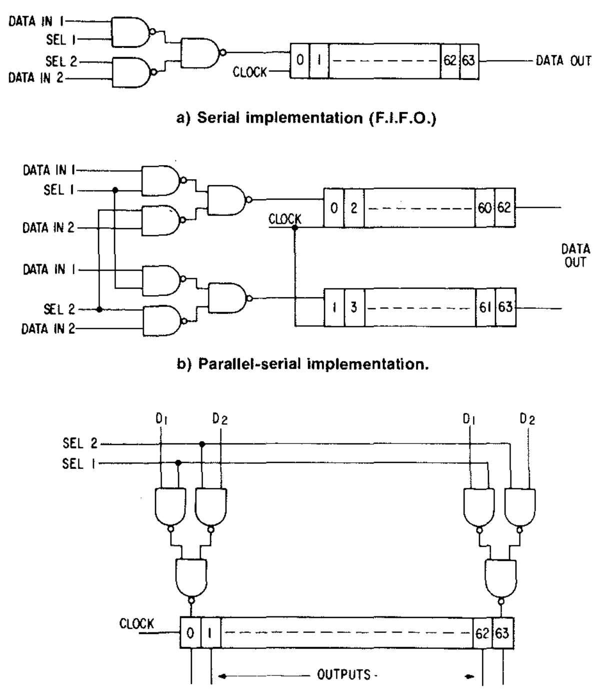


Fig. 9 — Logic for survivor sequence registers.



c) Parallel implementation (P.I.P.O.)

Fig. 10 — Survivor sequence implementations.

metrics as inputs during the next cycle.

For the survivor-sequence logic, a serial implementation was possible; instead of a PIPO register approach, one could go to the other extreme and use a first-in-first-out (FIFO) approach. Fig. 10a shows a typical register along with the gating required for the serial input data. The speed of such a serial system would be limited by the clock rate of the shift register (32 clock cycles required to completely update all of the survivor sequences). A compromise can be made between the FIFO and the PIPO register approaches (outlined in Fig. 10). The tradeoffs include operating speed, package count, and I/O pins/register.

Fig. 10b shows how the FIFO register can be split in half, effectively doubling the system speed. Or, it can be cut into quarters, etc. The limit results in a PIPO register organization.

Data outputting

After a fixed period of time ($\sim 5 \times K$), the decoder must make a decision for output data, providing a new data bit for each subsequent clock pulse. In a noise-free system, the MSB of all registers will be the same and the output could be taken from any arbitrary state. In the noise environment, the decoder must determine what state to select for each output bit.

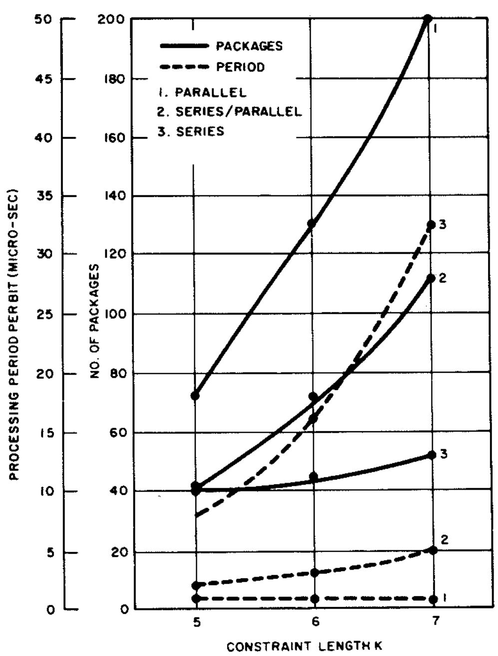


Fig. 11 — CMOS Viterbi decoder performance.

The optimum method of outputting data would be to determine which add/compare/select unit has the lowest metric and then output the MSB from the correspondent survivor-sequence register. The major drawback with this method is the amount of hardware required. Each state must be compared to determine which one has the lowest count. For K=7, this implies that 64 n-bit comparator outputs have to be checked. An alternate method of selecting the output bit would be to take a majority vote. This would require a logic network to determine the binary vector of the 64 MSB's. This method does not result in as optimum a decision as the previous approach, but, on the other hand, it requires much less hardware for its implementation.

Custom LSI implementation and performance

Based on the previous analyses, and the use of two customized LSI chip types (one for add/compare/select and one for survivor-sequence logic and memory), estimates were made for speed and total package count for different decoder organizations. These are shown in Fig. 11. The all-parallel approach results in a constant processing period of about 0.25 μs, independent of K. Package count increases rapidly. For a serial organization, the package count is relatively constant, but the processing period increases rapidly with K. The series/parallel approach falls in between, as expected. In all cases, the power dissipation is less than 1 W. Each logic technology applied to Viterbi decoding would yield different profile curves. However, the power dissipation can become an overriding limitation for all technologies other than CMOS, especially for spaceborne applications. Also the extent of LSI realizable is not as great as for MOS. Although a TTL decoder can use a higher degree of serial organization for a given bit rate, the power would be considerably higher and the package count could be higher as well, depending on the circuits used.

The custom LSI circuits hypothesized for the package count are based on chips with 40 or less bonding pads and having the equivalent of 300 or fewer gates — both resulting in high yield, low cost production. By extending these same designs to silicon-on-sapphire (SOS) technology, the data rates handled can eventually be increased from 2 to 4 Mb/s to over 20 Mb/s. Power dissipation and chip count would remain virtually the same.

Feasibility model

To demonstrate the operation of a CMOS Viterbi decoder, a model having K=3 and V=3 was built using commercially available parts in a completely parallel organization. A convolutional encoder was also built along with a test and evaluation unit. The test and evaluation unit contained the following:

1) A PN generator to provide data to the encoder,
2) Two PN generators to insert digital noise in the channel,
3) A comparator to compare the encoder output before and after the noise,
4) A set of counters to display the number of input errors,
5) A comparator to compare the original input data to the encoder with the decoder output,
6) A set of counters to display the number of errors not corrected by the decoder,
7) A message length control which will send out 1×10^6 or 8×10^6 bits and then turn off the data and the noise source,
8) A multivibrator to serve as the master clock and associated circuitry for resetting the system.

The transmission channel was a cable and all timing and control signals were generated in the encoder and transmitted to the decoder along with the data.

The input sequence comparator logic consists of an *exclusive-or* gate, counter, and register for each word. The reference words for the *exclusive-or* gates are generated in the encoder, and, for every mismatch between the reference and the input data, the counters are advanced by one. After the third data bit, the content of the counters is stored in the latch for use by the add/compare/select unit. The counters are then reset and are ready to process the next word. No node synchronization is required, since all timing pulses are supplied. If the timing was internal, then the decoder would have to determine which was the first bit of each word.

Each add/compare/select unit requires six IC's, four bit adders (one of which is used as a comparator), one package of inverters, one *and-or* select package and four bit register packages. Since hard decision was simulated, the difference in errors between the best state and the worst state can never exceed 6. Overflow-

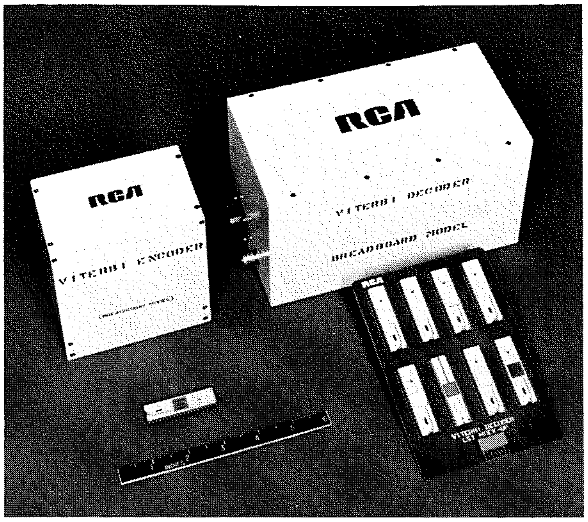


Fig. 12 — Feasibility model of convolutional encoder and K=3 Viterbi decoder.

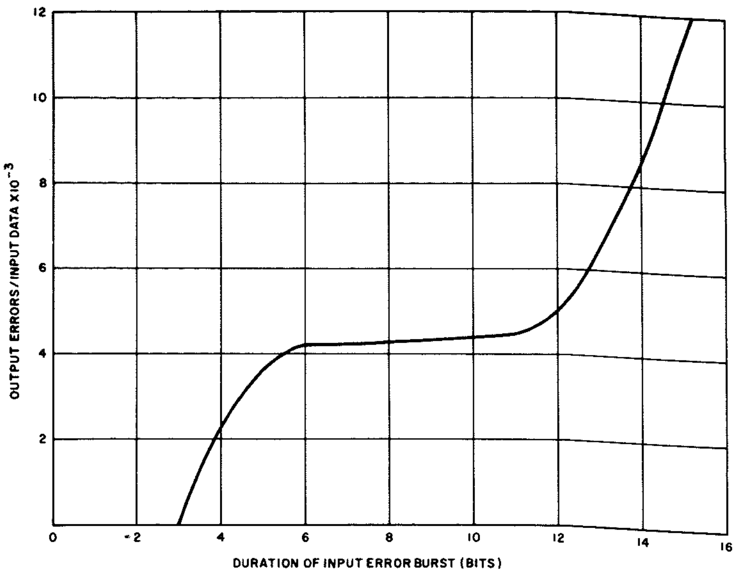


Fig. 13 — Burst-error characteristics of K=3 feasibility model Viterbi decoder.

ing is prevented by monitoring the most significant bit of each state output and subtracting eight from each metric when the minimum metric is $\geqslant 8$.

The survivor sequence registers contain 14 bits for each state and uses four CD 4035 for storage and four CD 4019 for data selection for each state. Output selection is determined by comparing all state metrics and selecting the best metric (lowest score) and then gating the most significant bit of that particular register to the data output line.

The K=3 feasibility model is shown in Fig. 12. It uses 7 packages for the encoder and 75 for the decoder. The test and evaluation circuitry (not shown) uses 55 packages. The decoder power consumption was approximately 30 mW at a 5V supply when operating at 100 kb/s. Mock-up circuits are shown next to the feasibility model for containing LSI packages. In that case, the same size decoder box could contain all the circuitry for a parallel organization, K=7 system for operation at a 1 to 2 Mb/s information rate.

Test results

Two types of tests were performed on this feasibility model. The first consisted of introducing isolated error bursts of variable duration to check for the error-correcting capability of the decoder. The results are shown in Fig. 13. The Viterbi decoder *corrected all errors of* 1-, 2-, and 3-*bit long sequences,* occuring at a separation much longer than 5 constraint lengths. (In the test case, the separation of the error bursts was 256 bits.) Then, the output error rate increases to about 4×10^{-3} to burst durations up to 6. Error rate remained relatively constant to burst durations of 11 after which it again increases very rapidly with increasing burst durations.

The other type of test consisted of introducing random errors into a bit stream and checking the output errors for message lengths of 10^6 bits. Both the message and errors were derived from PN generators, providing a qualitative performance figure. The following table gives some of the results of these tests.

No. of input noise bits in 10^6 message bits	*output errors*
2,300	0 or 1
10,000	20 to 35
25,000	25 to 60
65,000	150 to 500

Summary and conclusions

Viterbi decoders have been proven by many through computer simulations, and actual equipment is being used effectively in communications systems. The use of CMOS LSI is suggested for further enhancing the performance of a system employing Viterbi decoding and for extending application areas. In those cases where the power dissipated or the complexity required by the decoder (*e.g.*, in satellites) is too high, it might well be prevented from being used. (Increased transmitter power or receiver gain would be two alternatives). The CMOS technology can now accommodate information rates of 2 to 3 Mb/s. The same logic designs and partitioning for LSI can be extended to the emerging CMOS/SOS technology for handling bit rates to 30 Mb/s with virtually the same low power dissipation and packing density, thus making Viterbi decoders useful for the bulk of present and near future communication channels.

The Viterbi decoder for burst-error correction is also of value besides its proven performance in the Gaussian noise environment. Even a small constraint length unit (*e.g.* K=3) corrected all errors produced by noise bursts up to three bits in duration.

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Electronic PABX using large-scale integrated-circuit devices

E. D. Taylor | L. Kolodin | N. Hovagimyan

A solid-state PABX has been designed with a wide variety of service features and with a system capacity up to 600 lines. The switching matrix, line circuits, trunk circuits and common control are all implemented with large-scale integrated-circuit devices. To complement the use of this technology, the PABX is manufactured with extensive use of automated assembly, wiring and testing, to insure the product is economically attractive. The PABX is small in size, completely modular and suitable for many applications. The paper demonstrates how the advantages of modern solid-state technology are successfully applied to fulfilling the requirements, functions, and interfaces of conventional PABX's.

THROUGH A SERIES OF R&D programs, a solid-state switching matrix has been designed using large-scale integrated-circuit devices of the PMOS type. This approach—coupled with the use of LSI devices in station line circuits, trunk circuits, and common-control circuits—has led to an advanced design of a fully electronic PABX (private automatic branch exchange) equipment. To complement this design concept, the PABX is manufactured by extensive use of automatic (software controlled) assembly wiring and testing.

Design objectives

In the concept of the electronic PABX certain basic objectives were pursued. The first objective was to design a product with a range of service features suitable for both business and hotel/motel applications. The second objective was to insure that the product was sufficiently flexible to cover a wide range of size classes and applications using a minimal family of building blocks. The third objective was to design a product using hard wired logic control for basic systems but adaptable to the use of stored program control for more sophisticated systems.

Reliability, maintainability, ease of installation and general operating efficiency were all carefully considered in the total product concept. The extensive use of LSI devices and production automation was made to reduce manufacturing costs and human errors and to improve MTBF. The equipment requires no adjustments and the maintenance philosophy is based on plug-in module replacement, with subsequent repair completed at a centralized depot.

Authors Hovagimyan, Taylor, and Kolodin (left to right).

Table I — Standard features.

Station-to-station dailing
Direct outward dialing
C.O.-trunk/VCA interface
Multiple-trunk groups
Class of service (10)
Howler-tone (permanent call alarm)
Toll restriction on 0 and 1
Automatic trunk dropout
Flexible numbering plan
Station call transfer
Three-way conference
Consultation hold
Hot-line lobby phone
Attendant console
Key per trunk operation
Call splitting
Camp-on busy with indication
Attendant busy override
Attendant call transfer
Attendant trunks
Attendant conference
Attendant recall on no-answer
Night alarm
Equipment alarm to attendant
Room-to-room dialing restriction
Extend dial tone to restricted station
Processor interface for future options

Performance features

The standard features included in the basic PABX are summarized in Table I. Most of these service features are self-explanatory but some qualifying comments are worth noting. A flexible numbering plan permits intermixed 1-, 2-, 3-, and 4-digit dialing so that correct room-number-to-telephone-number correlation can be achieved in hotel or motel applications. The standard toll restriction on 0 and 1 will also permit the informa-

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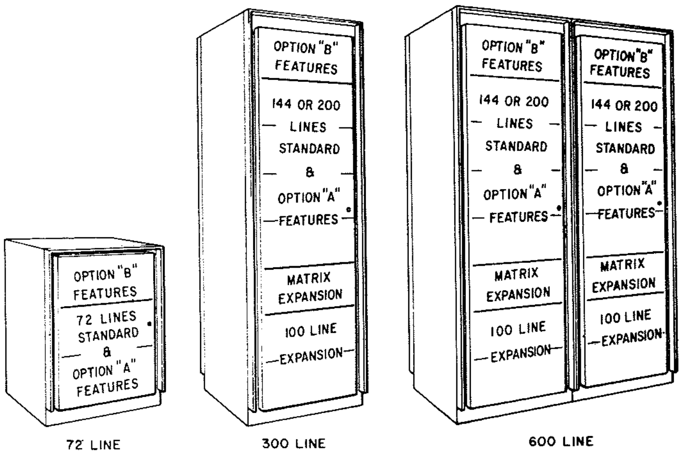

Fig. 1 — Equipment layout.

tion code (411) and the emergency code (911) to be dialed if applicable to the Central Office. Up to three multiple-trunk groups can be provided with separate toll restriction arrangements for each group. Any incoming call processed by the attendant, encountering no answer from the called party will automatically return to the attendant as a recall signal after ringing for a designated period of time. Classes of service may be assigned on a per-station basis. At the attendant's discretion, Central Office dial tone may be extended to stations normally restricted from this class of service. Howler tone is applied to any line giving a permanent off-hook signal exceeding 30 seconds duration.

The optional service features designed into the system are summarized in Table II. As with the standard features, most are self-explanatory but, again, some qualifying comments are in order. The *station busy* lamp field uses light-emitting diodes for long life and high packaging density. A message-waiting system can be provided over the normal telephone voice pair. In hotel/motel systems, local calls are recorded by message units in an electronic memory. This memory may be accessed via a small display console. When a hotel guest checks out, the console provides a means for changing the class of service of the room telephone from normal direct outward dialing (D.O.D.) service to a hot line to the attendant so that unauthorized D.O.D. calls cannot be made. The night patch service provides assigned night answering facilities for up to five PABX stations. The calling-station number display gives the attendant an immediate readout of the telephone number of any incoming call on an attendant (0-level) trunk. Alternatively this facility may be provided at any specific station, for example the room service station in a hotel. Progressive conference permits up to five parties to be conferenced by a dial-up sequence. A dictation trunk permits either dial-pulse or DTMF (dual-tone multifrequency) signals to be used to control the dictation equipment. Tie-line service includes ringdown, automatic, and two-way dial-repeat operation.

Louis Kolodin, Ldr., Government Communications Systems, Communications Systems Division, Camden, NJ, graduated from Rutgers University in 1954 with a BS, Physics, and obtained an MA, Physics, from Temple University in 1959. He joined RCA in 1954 and was assigned to Advanced Development, working on digital circuit development. He was project engineer for the sun-angle computer in the TIROS ground station, and was responsible for design of digital circuits on the AN/GKA-5. He subsequently designed transfer channels for the RCA 601 computer and the Autodin CDP. In 1961 he became Engineering Leader, responsible for the design of subscriber subsets on UNICOM, modifications to the Autodin CSU, and the VOCOM switchboard. Subsequently, Mr. Kolodin worked on a TDM switch matrix and was responsible for the preliminary equipment design on the Mallard Study Project. He was responsible for design of the circuit switch hardware for the Integrated Circuit Message Switch and for the development of a family of tactical switching equipment using PMOS-LSI. He is currently Leader, responsible for the electrical design of the RCA 600 PABX. Mr. Kolodin is a member of the IEEE and has been an instructor at Rutgers University, Camden, and for the RCA Engineering Program.

Norman Hovagimyan, Government Communications Systems, Communications Systems Division, Camden, NJ, graduated from Tufts University in 1952 with the BSEE and received the MSEE from Brooklyn Polytechnic Institute in 1964. He joined Western Union in 1952 and worked on various design projects, including multiplex equipment and interface equipment for use between computers and switching systems. Mr. Hovagimyan joined RCA in 1960 and has been engaged in the design and development of communication systems and equipment. He was assigned responsibility for system definition of the Mallard Study program, including signaling and supervision trade-offs, and the Mallard uniform directory. Subsequently he designed a Time Division Multiplex switching matrix, using micro-electronics and MSI techniques. He was then assigned to the VOCOM program as Project Engineer and performed many design tasks from system design to field test implementation. He is currently Project Engineer on the commercial electronic PABX. Mr. Hovagimyan was a contributing author to the *Communications Systems Engineering Handbook*, published by McGraw-Hill, and has patents on PMOS solid-state crosspoints, time-division-multiplex universal I/O module, and a one-wire crosspoint.

Equipment layout

A single standard cabinet 80 inches high, 28 inches wide and 24 inches deep is used to house a 300-line PABX, complete with all standard and optional features. A second cabinet of identical size can be mounted alongside the first cabinet, with the side panels removed, to provide a system up to 600-line capacity. For lower size class applications, up to 72 lines, a smaller cabinet is used.

The standard 80-inch PABX cabinet mounts up to 8 rows of plug-in modules. The equipment layout arrangements are shown in Fig. 1. The basic equipment building block is a four-row backplane that can mount sufficient plug-in modules for a complete 200-line PABX including both standard and option-*A* features (see Tables I and II). A three-row alternate version of the basic backplane mounts 144 lines, with standard and option-*A* features. One hundred additional lines may be added using a two-row backplane at the bottom of the cabinet, and for additional traffic capacity in a one-row matrix expansion backplane is added at the top of the cabinet, and for additional traffic capacity, a one-row matrix expansion backplane is used. For option *B* features, a one-row backplane is added at the top

Table II — Optional features.

A options
3-digit toll restriction
Station hunting
Paging trunk
Off-premise extension
Station-busy lamp field
Camp-on busy with indication & recall
Unassigned night answering
Calling station number display
Message waiting (2-wire)
Electronic message registration
Cut off of dialing at check-out
CEK interface
Rapid wake-up calling
Multiple-console operation
Battery back-up
Night patch
B options
DTMF operation (or mixed DTMF/DP)
Progressive conference
Dictation trunk
Tie-line service
Trunk-to-tie-line connections
More extensive 3-digit toll restriction
C options
Expanded matrix - 48 links
Expanded matrix - 64 links

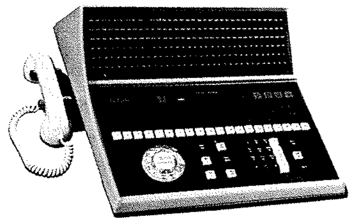

Fig. 2 — Attendant console.

of the cabinet. PABX systems over 300 lines and up to 600 lines use two standard 80-inch cabinets and appropriate combinations of backplanes. The small cabinet mounts a 72-line PABX with standard and option-*A* features in a two-row backplane. A one-row additional backplane is used to add option-*B* features.

Mechanical design

The basic mechanical component of the PABX is the backplane assembly. The backplane mounts as many as 21 plug-in modules per row, with up to 8 rows in a standard cabinet. The plug-in modules are spaced on 0.950 inch centers and each module uses a 120-pin edge-card connector in the backplane. In the rear of the backplane is a wire-wrap pin field and the assembly tolerances are carefully maintained so that wiring can be performed by automatic wire-wrap machines. The backplane is made of glass epoxy with printed-circuit wiring on both sides for power distribution and other bus-line connections. The metal side members insure that the total assembly has proper rigidity. A typical plug-in module is 9 inches long by 7.75 inches wide. The board is glass epoxy with printed circuits on both sides and plated through-holes. The 120-pin edge connector is keyed to preclude improper insertion, and the tabs are gold plated. Forty edge-connector test points are brought out to the front of each board for testing and maintenance purposes.

Attendant console

The attendant console has been designed for ease of operation and aesthetic appeal. Its contoured lines and neutral colors will blend with any office decor. All incoming and two-way trunks are terminated on individual answering keys which have integral long life incandescent lamps. Similar keylamps and keys are used for attendant trunks and operator functions such as *hold*, *release trunk*, *call splitting*, *camp-on*, *etc.* A keyset is used for rapid processing of incoming calls to local stations while a rotary dial is used to originate outgoing Central Office and Tie-Line calls. If the PABX and the local Central Office are equipped for pushbutton dialing, the rotary dial may be eliminated and the keyset used for originating all calls. The attendant console, shown in Fig. 2, is equipped with 20 trunk appearances, but an optional console provides two rows of trunk keys to accommodate up to 40 trunks. Input jacks for the attendant's handset are available on both sides of the console and the handset hangers may be mounted for use by either right- or left-handed attendants. The optional *station-busy* lamp field at the top of the console uses light-emitting diodes for extremely long life and can be equipped with up to 200 lamps. For systems using more than 200 lines, the busy-lamp field is housed in a separate cabinet.

Message register display console

As an advance over the method of local call registration by electro-mechanical counters, a system of electronic message registration has been designed using electronic memory storage. The memory is accessed by means of the message-register display console (Fig. 3) which can be mounted remotely from both the switch and the attendant console. The display console is about 6 inches wide and 9 inches long and requires only one 25-pair cable for any size system. The memory can be accessed by keying in a specific room number. The *room* and *calls* displays are immediately given on light-emitting diode arrays. The number of calls can be reset to zero by pressing the *reset* key. Similarly the room telephone can be changed to a hot line to the attendant by pressing the *cut-off* key. Following a *cut-off* condition, a room telephone can be restored to normal service by pressing the *connect* key. As an added precaution, the system memory is protected against power failure.

System concept

The PABX has a solid-state space division matrix and a distributed control which is multiplexed on common buses. Fig. 4 is a simplified block diagram of the PABX. The rectangular matrix consists of a number of PMOS-LSI chips mounted on matrix plug-in modules. In the interest of total system economy, telephone switching is on one wire with ground return. A careful analysis and computer simulation model showed that system crosstalk and noise objectives could be met with this method of switching, as long as certain equipment layout and wiring rules were followed. The use of automatic wiring under control of a computer-generated card deck insures that the system crosstalk objectives are met in production.

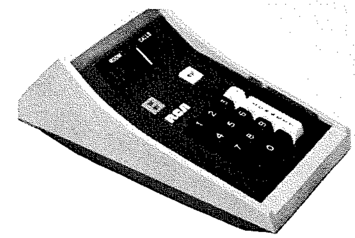

Fig. 3 — Message register display console.

Each matrix chip contains 64 crosspoints, and the associated crosspoint control logic to serve four lines on 16 links. The chips are assembled and partitioned so that the standard system is equipped with a 32-link rectangular matrix. Optional expansion matrix modules may be used to expand the system to 48 or 64 links as required.

The solid-state matrix carries only analog signals; all normal supervision and signaling is performed by time-shared multiplexed bus lines. The use of LSI chips in line circuits and trunk circuits permits a system of distributed control where line and trunk status is continuously monitored by a line scanner. Similarly, link and register scanners are used to monitor and allocate links and registers respectively.

Extension telephones are connected to the system through line-circuit modules, each of which terminates eight telephones. In addition to providing control logic, the line circuit also provides talk battery and ringing to the line and isolates the line from the matrix.

Trunks and tie lines are interfaced to the matrix by trunk-circuit modules, each of which terminates four trunks.

It is worth noting that many of the functions distributed on a per-line or per-trunk basis in LSI logic circuits have traditionally been located in common control in either call-stores, registers, or junctor circuits. Since LSI permits many functions in a single chip, the common control is simplified.

A typical call from station-to-station will be initiated by the calling station going *off-hook*. The line scan detects the line status change and interfaces both register scan and link scan to obtain an idle register and an idle link. A connection is then made between the register and the calling line over the link via two crosspoints. The register returns dial tone to the calling line via the analog path through the link, and the calling party then transmits dial-pulse information to the register over the information bus during an assigned time slot. The register accumulates and stores the dialed number and also stores the calling-equipment location. After the called number has been accumulated, the register determines if the line is idle or busy by placing the number on the address bus. If the called number is idle, the register will initiate a final connection between the two parties over a link and will set ringing in the called line circuit and then release. Otherwise, busy tone will be returned to the calling party.

The address-and-command bus has been designed so that a micro-processor can be accessed to control future optional features. However, all the service features described in this paper are controlled by hard-wired logic and the register-processor interrupt is permanently strapped to a continue mode.

Power sub-system

The standard power sub-system is driven from a 120-Vac 60-Hz single-phase supply. The logic power supply is mounted inside the rear of the cabinet and provides outputs of +12Vdc, −12Vdc and +5Vdc. A talk supply having an output of −48Vdc is mounted external to the PABX cabinet. For systems requiring protection against power failures, an optional battery back-up system is used. Power dissipation in the equipment is moderate and forced-air cooling is not required, even for fully populated systems at a maximum ambient temperature of +50°C.

Technical approach

A major difference between this PABX and most other commercial electronic switches is the extensive use of LSI devices throughout the design. A typical PMOS chip is mounted in a 40-pin dual in-line ceramic package measuring 2 inches by 0.6 inches. Eleven special chip types have been developed for this product, and they are used on an average of approximately two chips per line. Some of the specific advantages of the PMOS-LSI devices are reproducibility, low cost, high reliability, low power and small size. As an indication of the packaging density achieved, the line-circuit chip, when simulated in conventional DTL integrated circuit logic, consists of 69 chips in regular 14-pin dual in-line packages.

Production and test automation

A prime consideration in the PABX design effort was the automation of ordering, manufacturing, and testing to minimize the labor costs and to maximize production speed and accuracy. Custom orders are keypunched and processed by computer to produce a number of outputs. These include bills of materials for stock pull to the production assembly area, standard and custom card decks for automatic wiring control, wiring continuity test tapes, plug-in module location charts, and business management summary data sheets. In production, many of the components are automatically inserted into boards, and essentially all wiring is performed either by fully automatic or semi-automatic machines. After assembly, all wiring is fully tested on an automatic tester under control of standard and custom tapes. Additionally, all plug-in modules are computer tested by simulation programs, and the average test time for a plug-in module, including several hundred tests, is about 12 seconds.

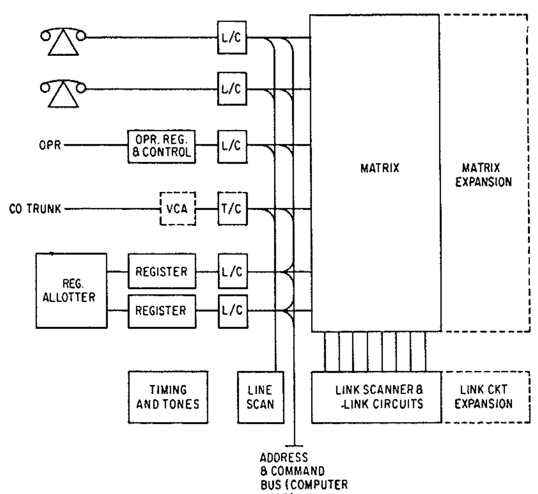

Fig. 4 — PABX system.

Installation and maintenance

Careful design consideration has been given to simplifying the field installation task and also to the tasks required to expand or modify an existing PABX without disruption of service. In particular, all external connections to the PABX cabinet are completed through standard 50-pin telephone type connectors. This permits stub cables to be connected to the MDF (main distribution frame) before the PABX is delivered to the installation site. When the switch arrives on site, the connectors may be quickly plugged into the appropriate sockets at the back of the cabinet. Additionally, the comparatively light weight and small size reduces shipping costs and simplifies equipment handling and floor space problems.

Any failure of a fuse or of a major equipment component is automatically brought to the operator's attention by a flashing red lamp on the attendant's console. Simple diagnostic procedures are used to isolate faults, first to subsystems, and then to a specific plug-in module. Fault diagnosis is performed using a special plug-in visual display tester which interfaces to special backplane system test sockets and to the front access test points on the plug-in modules. Facilities are also provided for determining idle/busy status and temporarily busying out lines, trunks, registers and links.

Conclusions

The PABX has been designed to meet the requirements of commercial telephone switching using LSI technology. The product has the special advantages of small size, light weight, low power consumption, and high reliability. In addition, it can be manufactured quickly and accurately, and installed and maintained by personnel without a high degree of specialized training, so that both installed first costs and life-cycle costs are minimized.

RCA's broadcast antenna engineering center

R. L. Rocamora

This paper describes RCA's Broadcast Antenna Engineering Center where 70 engineers and associated technicians combine their talents to design, build, and produce about 20 basic types of fm and tv transmitting antennas in addition to custom-engineered antenna systems. Such products vary from the timeproven RCA Superturnstiles to complex multistation antenna arrays that satisfy unique coverage requirements for a variety of viewer area contours. By employing modern computer-aided design techniques, plus full-scale horizontal- and vertical-radiation-pattern field testing ... upwards of 2000 fm and tv antenna structures and systems have been designed and produced in cooperation with engineers and consultants of the television broadcast industry. Several of RCA's antenna designs are reviewed in this article, and performance criteria and measurement techniques are described.

THE RCA ANTENNA Engineering Center at Gibbsboro, N. J., started in 1954 with a single building and one antenna-test-positioner (turntable) on a few acres of ground. At that time, only two or three types of broadcast antennas were being produced.

Today, the Gibbsboro complex encompasses 135 acres. It includes three large antenna test sites, two scale-model test sites, and more than 20,000 square feet in modern engineering labs, machine shops, offices and assembly buildings. The Center is now the design, development and production site for up to 20 basic fm and tv antenna types, associated transmission lines, filters, and complete custom-designed equipment systems for linking transmitter and antenna.

An organization chart of the Antenna Engineering Center and a picture of the engineering staff are shown in Figs. 1 and 2. The Center employs approximately 70 people of which about 25 are engineers. An aerial view of the center is shown in Fig. 3.

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Richard L. Rocamora, Manager, Antenna Engineering Center, Broadcast Systems, Gibbsboro, New Jersey, received his degree in electronics from New York University in 1946. He also attended graduate courses at the Polytechnic Institute of Brooklyn and the Moore School of Engineering at the Universtiy of Pennsylvania. Mr. Rocamora joined RCA in 1952 with an engineering background which included design experience in relay logic, power control systems and test instrumentation. His work at RCA included the development and design of transistorized digital communications equipment. In 1959, Mr. Rocamora was made a leader of this activity. He was subsequently promoted to Manager, Digital Equipment Engineering of the Communications Systems Division, a position he held until 1964, when he assumed his present duties.

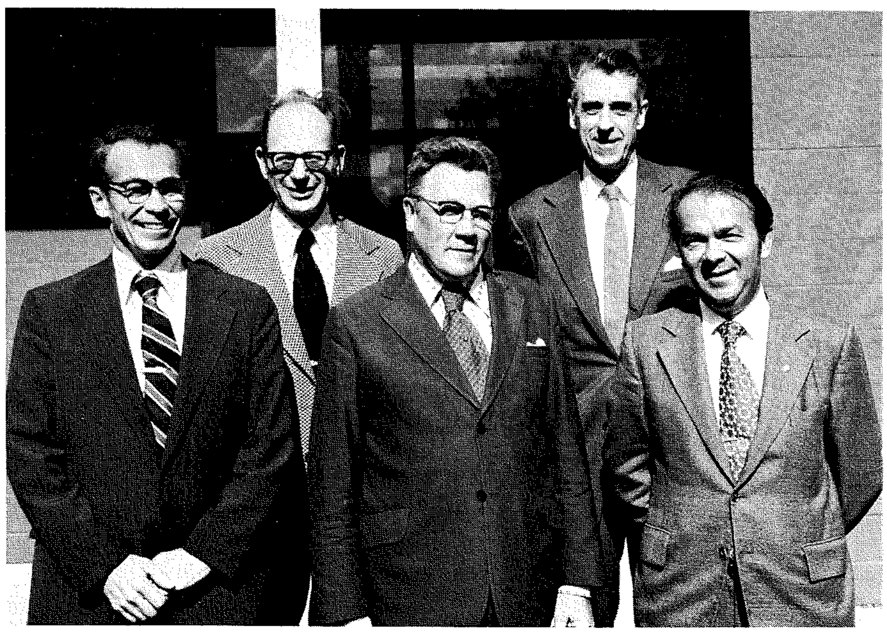


Fig. 1 — Gibbsboro staff - from left to right: D.G. Hymas, R.L. Rocamora, Dr. M.S. Siukola, H. H. Westcott, and N. Nikolayuk.

Antenna design evolution

Early tv antenna designs were omnidirectional vhf types with gains of less than 10; a few stacked dipoles were produced, but by far the bulk of deliveries were RCA Superturnstiles.

Superturnstiles

The Superturnstile antenna is constructed with batwing shaped radiators stacked one above the other. Because of its simplicity and superb performance, this pioneer vhf antenna remains today the standard of comparison world-wide. Some 700 RCA Superturnstiles have been placed in service.

As tv allocations began to escalate in frequency, the need for new designs in high-gain antennas became apparent. On vhf channels 7 to 13, for example, the complex mechanical feed system of the Superturnstile tended to limit reliability of the antenna. Thus, a simple design with lower wind load was indicated.

TW antennas for vhf

The low wind load need led to the development of the Traveling Wave (TW) antenna and the new feed system for which it was named.[1] The rugged,

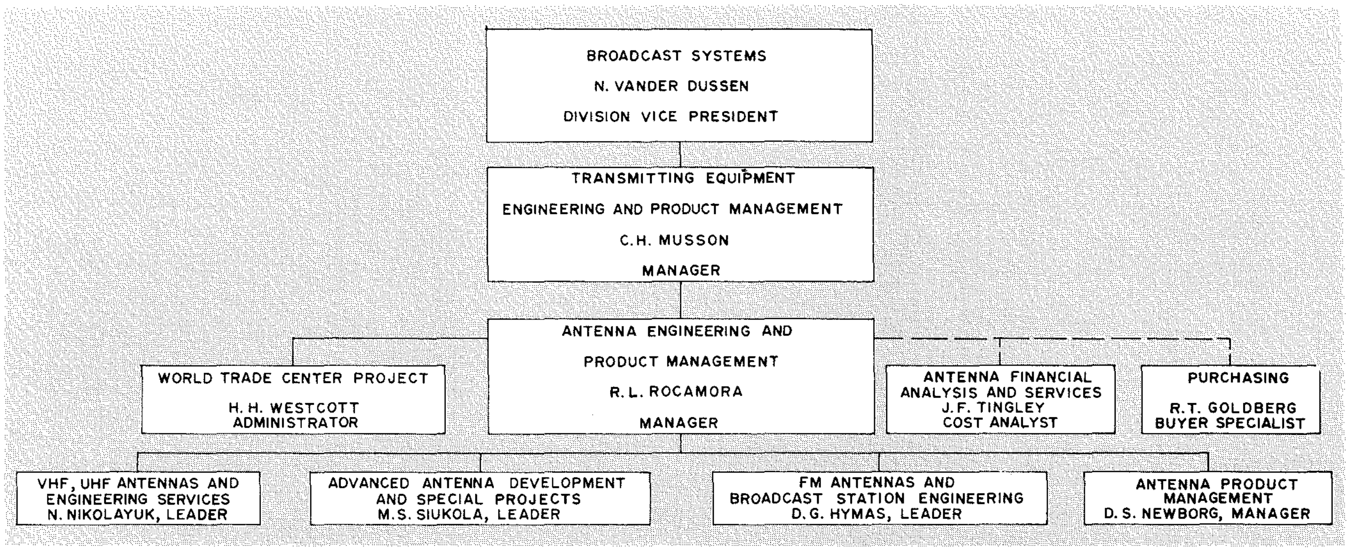


Fig. 2 — Gibbsboro antenna engineering and product management organization.

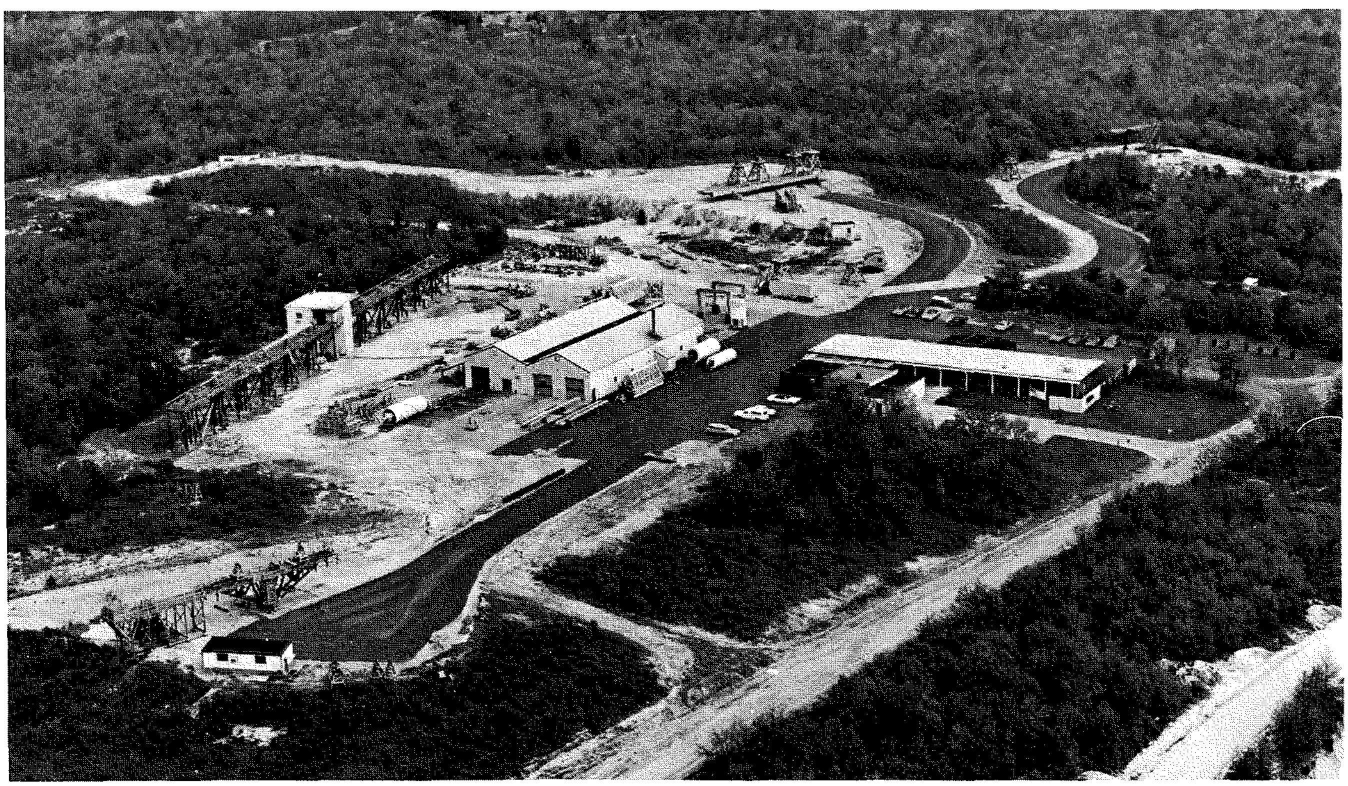


Fig. 3 — Gibbsboro Antenna Engineering Center.

enclosed feed system, connected at the base of the antenna, consists of a single inner conductor which sets up a traveling wave between it and the supporting outer conductor. Pairs of slots on opposite sides of the cylinder are fed by capacitive pick-up probes. The Traveling Wave antenna not only provides gains of 9 to 18, but gives smoother vertical patterns and better circularities than the Superturnstile. With RCA's 25-kW transmitter, the system achieves the FCC maximum 316-kW effective radiated power (ERP) for vhf channels 7 to 13 with power to spare. The TW is still one of the most popular of the vhf high-band antennas. Over 110 have been built and installed.

UHF Pylons

Gibbsboro continued its high-gain antenna development program and, in 1954, scored a major breakthrough in uhf design. This antenna, the first ultra-gain Pylon for uhf, is a slotted-cylinder type, 100 feet high, with nominal gains of up to 60.[2] Again, the radiator is an integral part of the structure, the slots being energized by a standing wave on the inner conductor. The feed system is a simple design with built-in provisions for beam-tilt adjustment. The Pylon antenna made possible the first uhf station with one million watts ERP. As a measure of broadcast industry acceptance, RCA Pylons are in use by the majority of uhf stations in this country; approximately 400 have been delivered.

Trends of the "fifties"

Beginning late in the '50's, Gibbsboro engineers sensed a changing market and responded with new products to meet its needs. The new market needs included:

- Increases in ERP and tower heights.
- Increases in use of directional antennas.
- More sharing of towers and antennas by broadcasters or sharing of common tower structures with individual antennas.

The station owner's bid for signal dominance (as reflected by the demand for increases in ERP) resulted in much original design work by RCA on antennas, transmission-line components and supporting structures. Requirements were usually for radiating the maximum power from towers ranging in height up to 2,000 feet. Several installations were made.

Zee, VeeZee, and Butterfly types

Although many directional applications were filled by the slotted cylinder antennas, the radiation patterns of these types were not optimum in some cases. This problem was met with the development of a series of uhf and vhf panel antennas known as the "Zee", "Vee-Zee" and "Butterfly." Zee and Vee-Zee panels utilize zig-zag elements.[3] The Butterfly panel[4] is based on the hardware of the Superturnstile. Panels can be face-mounted from one to five around the tower to provide omnidirectional as well as directional patterns of any desired shape. Panel antennas, with their tower-like mainframes, are advantageous in multiple antenna systems since they can be used alone or as supports for other antennas.

To meet the requirements of multiple station systems, it was necessary to develop new antennas as well as modify existing designs so they could be stacked side by side, one above the other, or both.[5] Auxiliary equipment was also introduced to permit stations to share antennas and antenna apertures.

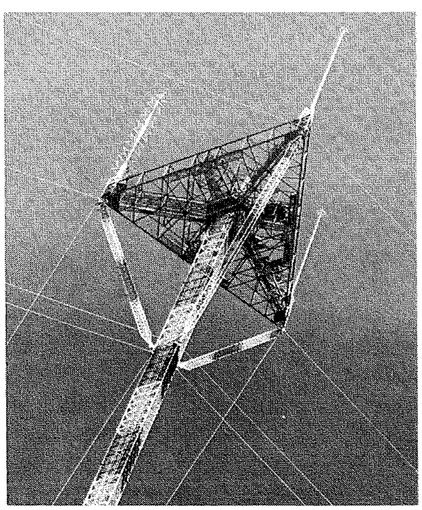
Broadband fm antennas

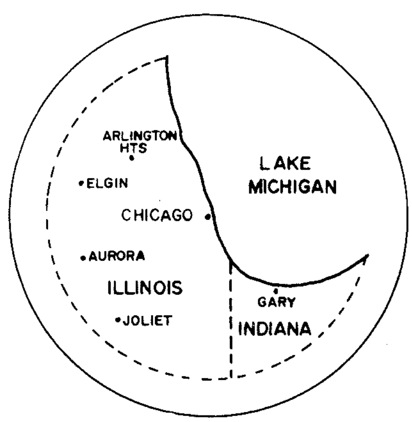
With the move to multiple-station systems and their ancillary services, the need arose for a broadband fm antenna design. Development of a broadband, circularly polarized antenna[6,7] made it possible to radiate both horizontally and vertically polarized fm signals from the same antenna. This design became the basis for future multi-station fm systems.

The mechanical and electrical constraints of stacking antennas on a platform at the top of a tall tower are numerous. And where broadcast groups form to share facilities, individual preferences must be met, sometimes under the most adverse conditions.

Multi-station arrays and candelabra designs

All broadcasters in a multi-station system vie for the same audience. Thus each is vitally interested in, among other things, his coverage and the position of his antenna on the tower. Since matters of this kind are often decided by other than engineering considerations, the antenna, which for obvious technical reasons should be at the top of the tower, may actually wind up at the bottom. But the designer must still satisfy the requirement





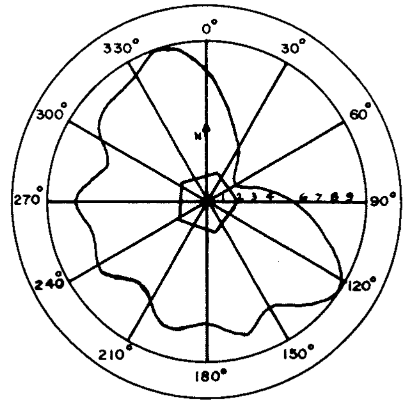


Fig. 4 (left) — Side-by-side Candelabra antenna system on tower at Sacramento, California. Fig. 6 (above) — Directional patterns of John Hancock Polygons.

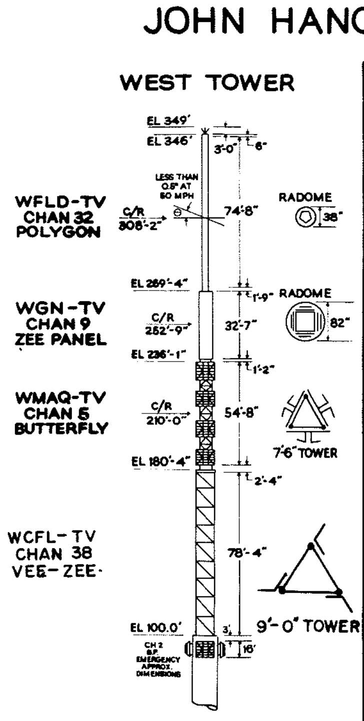


Fig. 5 — John Hancock multi-station antenna system.

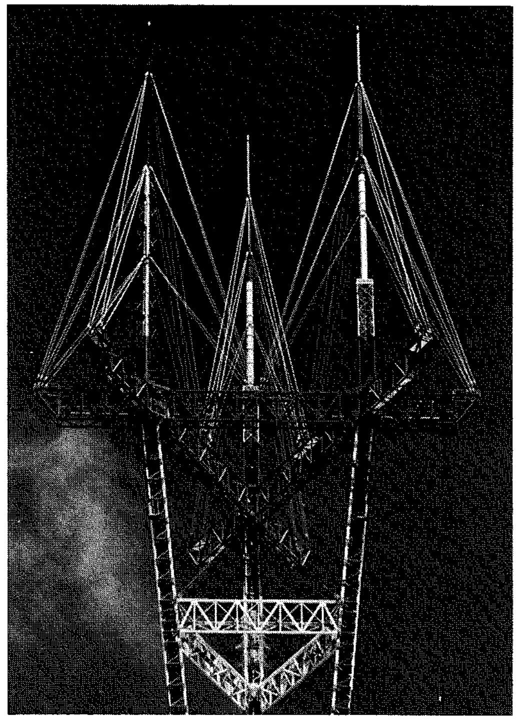


Fig. 8 — Mt. Sutro antenna system.

for a solid structure and provide antenna designs that assure good performance characteristics. This partly explains the need for the wide variety of antenna types and system combinations.

Obviously the design of multiple-station tv antenna systems calls for considerable experience in analyzing the effects of scattering on the patterns of antennas operating in close proximity to each other. This was the subject of intensive study[8] as early as 1957, and the findings were used for several candelabra designs that were to follow.

In early installations, specifications of pattern circularity were verified and demonstrated by means of scale models. Subsequently, however, equally accurate theoretical methods of determining gain were formulated.[9] For several years now, these computer-based techniques have been used in design and cost estimating calculations, and most recently to provide customized vertical patterns for antenna arrays.[10]

Typical multi-station systems

The first of the multiple-station systems, following the historic Empire State Building array, were the candelabras consisting of two or more tv antennas horizontally stacked on the same platform at the top of a tower (Fig 4). RCA has equipped about 10 of these "side-by-side" systems in this country, using Superturnstiles, Pylons and Traveling Wave antennas.

Typical of more sophisticated designs are the RCA multi-station systems on John Hancock Center Building in Chicago, and on Mt. Sutro near San Francisco (Figs. 5 and 8).

Both the 6-station John Hancock system and the 11-station (7 tv, 4 fm) Mt. Sutro system combine horizontally and vertically stacked antennas. The John Hancock system employs five types, one of which is the new 5-mW uhf Polygon. There are two Polygons on John Hancock (channels 32 and 44) with patterns tailored to direct radiation around Lake Michigan (Fig. 6). This is the first use of the Polygon, a 5-sided self supporting structure consisting of zig-zag panels. The Polygon has a gain capability of 60, and can be designed for power inputs up to 200 kW.

Antenna performance criteria

All antenna types (Fig. 7) are adjusted for proper input impedance before delivery to the customer. As to other tests required, RCA broadcast antennas are categorized as follows:

1) *Product-line types* requiring no pattern or gain tests. These are the omnidirectional, medium-gain Superturnstile, Traveling-Wave and Butterfly types which have a history of strict adherence to calculated radiation patterns and gain figures. Proof-of-performance tests are conducted only at customer request and at extra charge.
2) *Product-line types* for which verification of vertical pattern and gain is recommended. These are the omnidirectional high-gain antennas such as the uhf Pylon and Polygon. Tolerances in the manufacture of these antennas usually contribute to some deviations from the calculated patterns, making measurement with pattern trimming advisable.
3) *Directional vhf and uhf types*. Pattern and gain tests are required for all directional antennas which, in addition to their directivity, may have a variable beam-tilt design in the azimuthal plane.
4) *Multiple-station systems*. Scale models are used to determine the effects of adjacent antennas.
5) *Circularly polarized antennas*. Special test-procedures include measurements of axial ratio and certain other parameters.[11]

Superturnstile	Zee panel
Butterfly	Vee-Zee panel
Traveling Wave	Polygon
Pylon	Mark 3 panel
FM	

Fig. 7 — Antennas in production at Gibbsboro.

Measurement techniques

Vertical patterns are recorded with the antenna lying horizontally on a rotating positioner. Azimuthal patterns are recorded on the vertical pattern positioner by rotating the antenna on its longitudinal axis, spit fashion. Azimuthal patterns can also be recorded with the antenna mounted vertically on a positioner. Where size or complexity make it impractical to mount an antenna system on standard test facilities, scale models are fabricated and tested in free space. Gain is determined by mechanical integration of the recorded patterns.

Throughout pattern measurements, the antenna under test is used as a receiving antenna in accordance with the reciprocity principle. Source antennas are selected and located so as to provide illumination of the antenna test range. Upon delivery and erection of an antenna at the customer's site, field-service engineers conduct antenna system performance tests. Since all antennas have been adjusted at the factory for proper input impedance, field tests should be minimal.

Systems must be checked after installation of transmission lines to assure that there are no excessive discontinuities, a minimum of reflections and low mutual coupling in the case of multiple installations. An rf pulse technique[12] using a vestigial sideband signal to simulate the television signal is used for reflection tests. Impedance transformers are employed when required to achieve optimum reflection levels.

Antenna center facilities

The 135-acre facility is laid out to provide the best possible conditions for efficient antenna handling and testing (Fig. 9). Included is a model range for design and development of antennas and filter equipment.

The Center is built around two antenna ranges, one 18,000 and the other 10,000 feet in length. Supplementing the ranges are four antenna positioners, each tailored for special requirements. Each of the following facilities (Figs 10, 11, and 12) has its own control building equipped with the most modern measurement equipment available:

1) *North Hill* is the site of the largest positioner. This facility resembles a 130-foot long barge located diametrically on a 90-foot circular rail track. Positioner capability

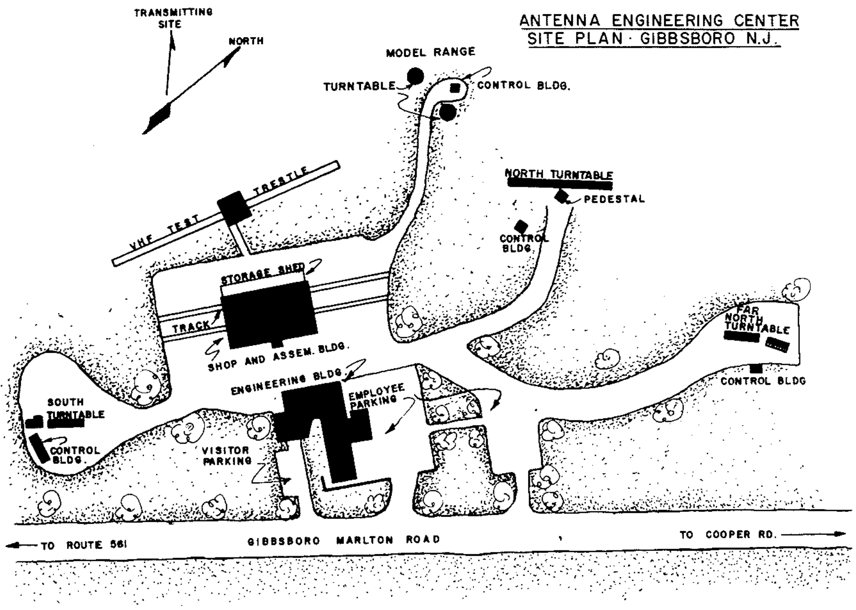
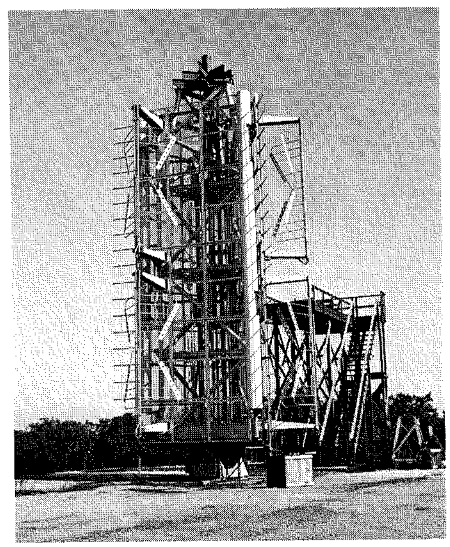


Fig. 9 — Gibbsboro facility layout.



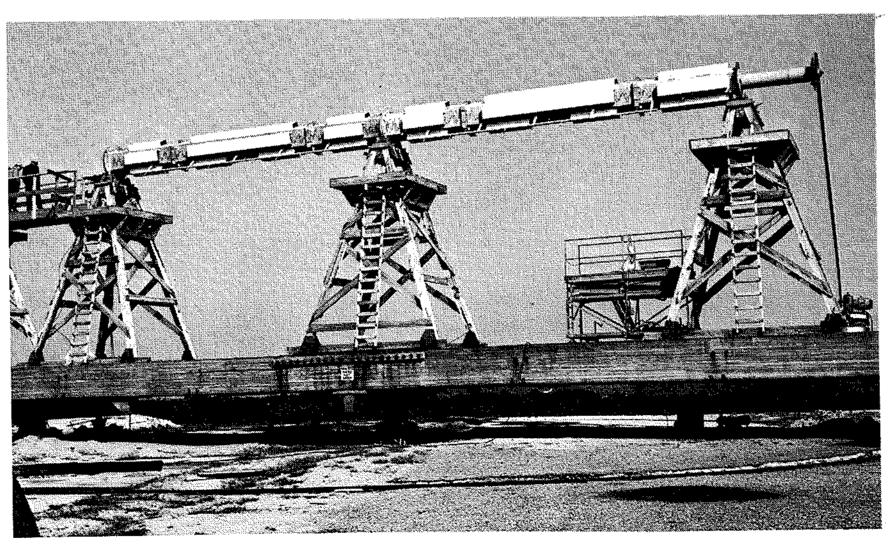


Fig. 10 (left, above) — Scaled antenna on model-range positioner. Fig. 11 (right, above) — UHF Polygon on North Hill positioner. Fig. 12 (above) — Butterfly antenna on Far North positioner.

was recently extended to handle antennas as long as 150 feet and weighing up to 45 tons. Antennas for all tv frequencies can be supported horizontally up to 25 feet above the barge, which has an automatic spit turning system. Also at North Hill is a pedestal type positioner which is used for azimuthal pattern measurements and impedance testing of low band vhf antennas.

2) *South Hill* is the oldest facility at the Center. The positioner at this site is used primarily for uhf Pylons and high band vhf (Traveling Wave) antennas up to 120 feet long and weighing 15 tons. Panel antennas are also tested on the South Hill.

3) The *Far North* facility was added in 1967 to accomodate increased uhf business, starting with the Kentucky Educational System. The basic equipment is a pedestal-type positioner. A support trestle is normally used to measure horizontally positioned Pylon and panel antennas up to 80 feet long and weighing .10 tons. The main pedestal positioner also serves for impedance and azimuthal pattern testing of vertically mounted massive antennas such as those recently tested for the World Trade Center project.

4) The *Model Range* incorporates two multi-axis positioners for scale model antenna and system measurements.

Filters and transmission line

An equally important assignment of skills at Gibbsboro is the design of various high power filters and coaxial transmission line for use with fm and tv transmitters operating over the broadcast range of 54 to 890 MHz.

Harmonic filters are necessary to reduce harmonic output of fm and tv transmitters to the stringent levels prescribed by the FCC. Another type of filter, a combiner, is commonly used today to combine the signals from two or more transmitters for higher output power. A filter of this type was recentiy developed to combine the output of three 15 kW vhf transmitters into a single antenna.

The vestigial sideband filter is a single shaping filter used at the output of the television transmitter to attenuate a portion of the visual lower sideband while maintaining the passband within a fraction of a dB. The diplexer is a form of filter that must be used to combine the visual and aural transmitter outputs for vhf antennas (such as Traveling Wave antennas) onto a single transmission line feeder.

The functions of vestigial sideband filtering and diplexing have been combined in a vhf filterplexer recently introduced for the RCA TT-50FH 50 kW transmitter.[13] Filterplexers have been designed to handle power as high as 220 kW at uhf.

To meet the requirements of voltage, power and efficiency at these frequencies, filtering elements must be made of coaxial transmission line sections or waveguide cavities, rather than of less bulky lumped constants. Even so, the heat developed often requires some form of built-in temperature compensation to eliminate filter detuning. One example is the new high power Filterplexer,[13] where cavity probes were compounded of Invar and steel to maintain proper length with temperature change.

New filter designs are assisted by computer analysis of mathematical models. They are now air cooled rather than water cooled. The capability of coaxial tv line has been increased by the use of Freon to accelerate heat transfer from the inner to the outer conductor. Proper exploitation of Freon, therefore, may be expected to increase present power ratings of coaxial transmission lines and components.

Custom designed systems

Frequently the fm and tv broadcaster desires a transmitting system different from the standard and tailored to his specific needs. To meet this requirement experienced Gibbsboro engineers work closely with the customer and Broadcast Sales to first define the optimum system and to set system specifications. A proposal is then drafted and presented to the customer.

Upon approval by the customer, the custom system is then finalized in design, fabricated and tested at Gibbsboro. On site installation guidance is often offered with the more complex systems.

The custom system engineering ranges in complexity from transmitting room installation plans showing equipment placement and transmission line routing, to RF output switching and logic circuitry for multi-transmitter parallel operation. Typical systems can be found in the Empire State Building, John Hancock Building in Chicago, Mt. Sutro in San Francisco, New Jersey Educational Broadcast Facilities, and in most major cities of the U.S. as well as some overseas.

Several custom systems originally designed by Gibbsboro are now being offered as standard items with current transmitters. For instance, approximately the first 20 paralleled tv transmitters were designed on a custom basis; a standard paralleled transmitter is currently offered as a catalog item. A circuit to sense transmitter exciter output and automatically switch to a spare in event of failure is now standard with the "F-Line" tv transmitters, while custom packaging of rf output switches in unitized cabinets led to the standard "OPTO Switch" design for vhf-tv transmitting systems.

Conclusion

The need for new types of broadcast antennas and associated equipment has grown substantially in the last two decades. The quest for higher power, the desire of broadcasters to share antenna structures, increased use of directionals, extensive activity in UHF, all have placed new emphasis on design integrity. RCA engineers at Gibbsboro have kept abreast of these requirements by carrying out the following programs:

1) Initiating state-of-the-art solutions to system problems,
2) Maintaining a sophisticated test range to verify concepts, and most importantly,
3) Continuing to provide antenna equipment with a performance and acceptance record that is unmatched anywhere in the industry.

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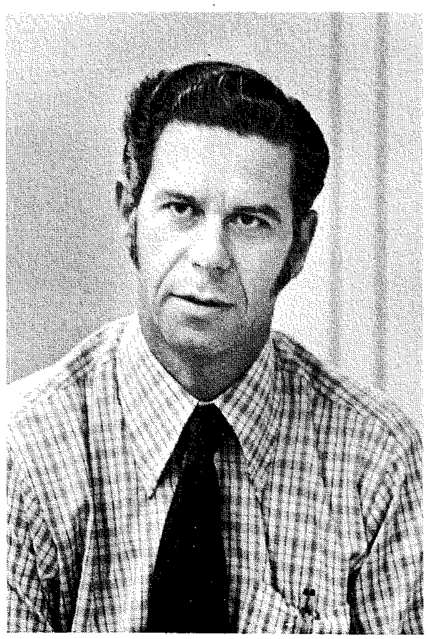
Trade-offs in rf transistor design

D. S. Jacobson

The geometric design of a microwave transistor is a primary factor in determining reliability and rf performance. This paper compares the major types of structures and the advantages of each.

David S. Jacobson, Mgr., RF and Microwave Devices Design, Solid State Division, Somerville, New Jersey, received the BChE from the Polytechnic Institute of Brooklyn in 1954. He then joined the General Electric Company where he participated in the Chemical and Metallurgical Training Program. After a tour in the U.S. Navy, Mr. Jacobson joined the Semiconductor Division of Texas Instruments in 1958 and worked on processing techniques for grown-junction silicon transistors. Mr. Jacobson joined the RCA Semiconductor and Materials Division in 1961 as a device development engineer; he engaged in the development of germanium double-diffused transistors and germanium and gallium-arsenide tunnel diodes. He was promoted to Project Leader of the Varactor Device Design and Development group where his major responsibilities included the development of high-reliability varactors for the LEM program. Mr. Jacobson transferred to the RF Transistor Design group in 1965 where he worked on the 2N3866 overlay transistor. Subsequently, he was responsible for the design and development of microwave power transistors, including the 5-watt, 2-GHz 2N5921 and 2N6266 and the 2-watt, 2-GHz 2N5920 and 2N6265. Mr. Jacobson was promoted to Engineering Leader of the Microwave Transistor Design group in 1968. In that position he was responsible for the design and development of microwave power transistors, low-noise transistors, and transistors for CATV applications. He is presently responsible for the design of all RF and Microwave Transistors and Microwave Integrated Circuits.



A NUMBER of transistor design geometries have been used in the industry to achieve high power at rf and microwave frequencies. The three basic or most common types are the overlay, the interdigitated, and the mesh geometry. Each of these structures has its own unique advantages as well as limitations. Consequently, each new design must be examined from the viewpoints of the state-of-the-art of the technology available, the rf performance, and the reliability demands of the application before the appropriate design is selected. To be specific, a transistor designer must address at least three practical factors:

- What is the optimum level of technology that the transistor manufacturer can successfully develop to a production capability?
- How much does the operating frequency requirement merit compromises with reliability?
- Which structure offers the best trade-offs of performance, cost, and reliability?

While these considerations are vital to a microwave transistor designer, some insight into the various design constraints of microwave power transistors also is helpful to transistor users.

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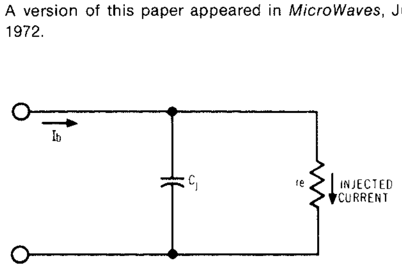


Fig. 1 — The input equivalent circuit for a microwave power transistor requires minimal junction capacitance for injecting current into the base.

$$EP = I_c \; / \; I_{m/l} \qquad (2)$$

where EP is periphery in mils, $I_{m/l}$ is the maximum collector current per mil of emitter periphery in mA/mil, and I_c is the collector current in milliamperes.

To optimize frequency in an rf transistor, it is essential to pack the required emitter periphery into the smallest practicable base area while also using the minimum emitter area. This requirement for minimum emitter area is essential to minimize the input capacitance to reduce shunting of the emitter-base diode. The input equivalent circuit (Fig. 1) shows that the current, which is injected into the base region, flows through the resistor, r_e. However, this component is shunted by the input junction capacitance, C_1. As the operating frquency is increased, junction reactance is decreased resulting in a larger portion of the base current by-passing r_e and, therefore, not injecting into the base. For this reason, high-frequency transistor designs generally emphasize a high ratio of emitter periphery to emitter area (EP/EA).

Table I — Effect of geometric design factors on transistor characteristics.

Transistor characteristic	*Geometric design factors*
Thermal resistance (θ_{jc})	Collector-base periphery Number of base cells Cell spacing Aspect ratio of base region
Current distribution (to minimize debiasing of portions of emitter)	Design of metallization pattern Design of base grid (overlay) Design of emitter grid (mesh)
Current-densities-metallization (to minimize metal migration)	Design of metallization pattern Finger width and finger spacing defines maximum metalization thickness and thus current density.
Emitter ballasting	Geometry should permit addition of sufficient ballasting to assure reliable operation with minimum compromise in EP/BA
Parasitic capacitances	High EP/EA High EP/BA Minimum emitter-collector MOS capacitance Minimum base-collector MOS capacitances (due to bond pads over collector oxide)
Parastic inductances	Sufficient number of bond pads and sufficient pad area to permit the use of optimum number of wires of optimum diameter to achieve low input Q and low common lead inductance

The output-current shunting is similar. Fig. 2 shows that the load current is reduced by a shunting reactance related to C_{ob}. C_{ob} also acts as a shunt capacitance across the real part of the output impedance (Fig. 3) and decreases the output impedance. This effect becomes severe at higher frequencies, where the transformation from a low impedance to 50 Ω becomes very "lossy." Because the other major degree of freedom in minimizing C_{ob} is to increase collector resistivity, which promotes base widening[1] and limits power output, the designer must optimize ratio of emitter periphery to base area, EP/BA.

In addition, to the optimization of these two key design ratios, EP/EA and EP/BA, within the limits of the available technology, a transistor designer must also consider those geometric factors which affect thermal resistance, current distribution, current densities in the metallization fingers, adaptability to emitter ballasting, and parasitic capacitances and inductances. These factors are summarized in Table I.

The three major design geometries which have been used to fabricate rf and microwave power transistors are the overlay or base grid structure, the mesh or emitter grid structure, and the interdigitated structure (Fig. 4). Some transistor designs incorporate combinations of these basic geometries to optimize for certain results.

Interdigitated emitter structures

Interdigitated emitter structures represent the oldest class of rf power designs, but have been used on transistors rated up to 3 GHz. This structure or its modifications is largely used by TRW as well as several other manufacturers.

Design requirements

The starting point in the design of any rf power transistor is its geometric structure, which is determined by the power level required and the frequency of operation. The power level at the desired operating voltage and an estimated collector efficiency defines the current-handling capability requirements, I_c, of the transistor by

$$I_c = \frac{P_o}{V_{cc}\,\eta_c}\,(100) \qquad (1)$$

where I_c is the collector current in amperes, P_o is the power output in watts, V_{cc} is the collector bias voltage in volts, and η_c is the collector efficiency in percent. Because most of the current at microwave frequencies is injected near the edge of the emitter sites as a result of transverse debiasing under the emitter, the required current-handling level must be translated into an emitter-periphery requirement. This requirement can only be defined if the current-handling capability per mil of emitter periphery is known. Values ranging from 1 to 1.5 mA/mil are typical. The emitter periphery, EP, is then defined as follows:

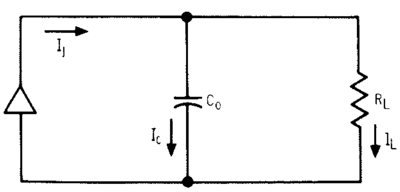


Fig. 2 — The output of a transistor also can suffer from current shunting with excessive Cob. It requires a high emitter periphery-to-base area ratio.

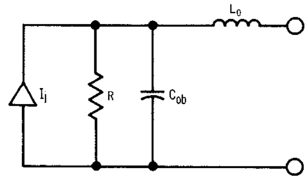


Fig. 3 — Load current is reduced by shunting reactance Cob, which also acts as a shunt capacitance across the real part of the output impedance.

Either a straight comb structure is used, (Fig. 4a) where the emitter is a continuous region, or a variation where the emitter fingers are discrete and are interconnected by the metallization, as shown in Fig. 4b. Another variation is a "tree" or "fishbone" version in which the emitters are interdigitated with the base structure in two dimensions, as shown in Fig. 4c. Of these structures, the "tree" or "fishbone" version results in geometric *EP/BA* ratios comparable to those of the metal-grid overlay and "mesh" approaches. These structures permit more of the emitter periphery to be directly contacted by emitter metallization than the "mesh" approach, and more of the base-conducting region at the surface to be contacted by the base metallization than the standard p^+ overlay approach. Therefore, this design should be capable of achieving relatively high current-handling capability per mil of emitter periphery. Processing is simpler than in the overlay approach.

To achieve high *EP/BA* ratios with the interdigitated design requires that narrow, closely spaced, interdigitated metallization be used. This results in higher current densities and greater potential for migration failures depending on the metallization system used.[2] Although finger ballasting can be used when the emitter sites are disconnected (Fig. 4b), the ballast resistors cannot be located as close to all segments of the emitter site as in the overlay structure.

Overlay design

The overlay structure[3] in Fig. 4d has been in use by RCA since 1964 on various rf power transistors operating from 30 MHz to 5.0 GHz. The standard p^+ grid overlay structure uses a number of discrete emitter sites which are connected in parallel by a broad emitter metal finger running perpendicular to the length of the emitter sites. Base current is conducted from the central portions of the emitter length to the base contacts on either end in a high-conductivity p^+ grid. The *advantages* of this structure are

- The relatively wide emitter metal fingers and the greater spacing between the base and emitter fingers permit thicker metallization, which results in large cross-sectional areas and in lower current densities compared to the other major structures.
- The discrete emitter sites and the disposition of the emitter metal fingers relative to these sites permits the introduction of a polysilicon layer (PSL) between the emitter metal and the emitter sites (Fig. 5). By controlling the doping and contacting geometry of the PSL, ballast resistors may be placed in series with each emitter site. This arrangement permits control of hot spots through negative feedback at the source rather than at the end of the metallization finger as is done with finger ballasting. Consequently, hot spotting may be minimized with lower levels of ballasting and with less compromise of power gain.
- The PSL layer also provides protection against alloy spike failures of the aluminum metallization through the shallow emitters and dielectric failures as a result of aluminum shorting through pinholes in the base oxide.
- *EP/EA* ratios as high as 50 mils/sq. mil have been achieved.

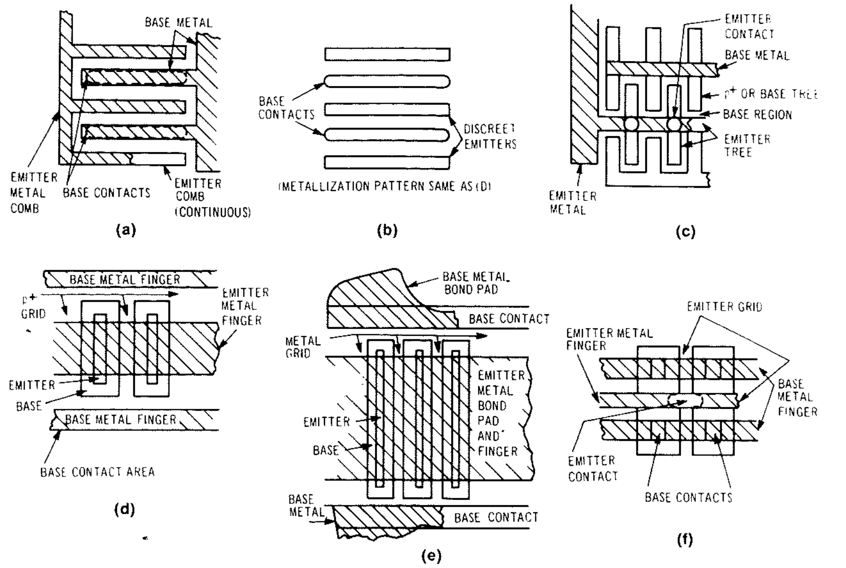


Fig. 4 — Cell structures of rf transistor geometries are shown here for (a) interdigitated emitter comb; (b) interdigitated emitter-discrete; (c) integrated emitter-"tree," (d) normal overlay p+ grid; (e) metal grid overlay; and (f) "mesh" or inverse overlay.

The *limitations* of the overlay structure relative to the achievement of higher-frequency microwave performance are as follows:

- Design ratios *EP/BA* have been limited to approximately 3.5 mils/sq. mil because of the lateral diffusion of the p^+ grid. This factor sets a minimum spacing between the emitter and the p^+ grid to avoid the intrusion of the emitter edge into the highly doped p^+ region. Such intrusion would result in poor injection efficiency and early power saturation.
- *EP/BA* ratios are also limited by the debiasing of the central section of the emitter as a result of voltage drops in the P^+ grid. This limitation defines a maximum emitter length

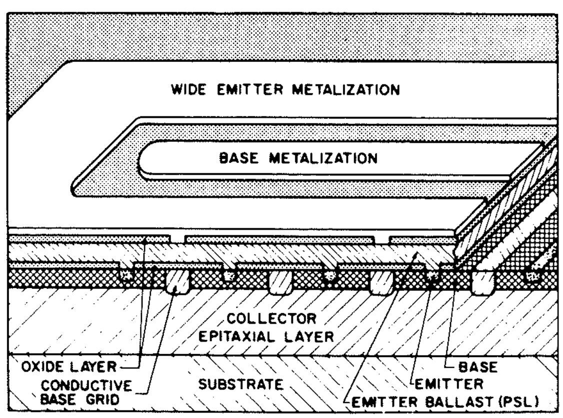
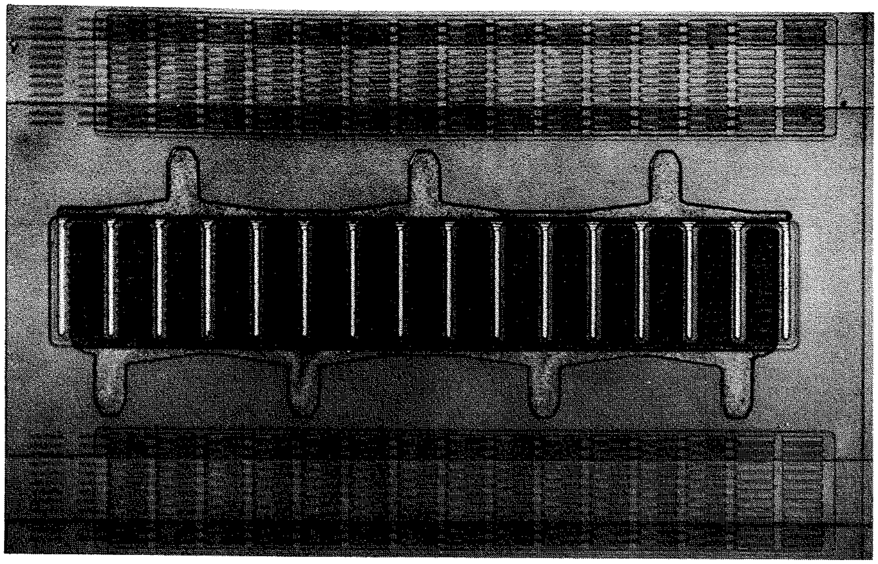


Fig. 5 — Cross section of an overlay transistor shows how emitter ballasting may be placed in series with each emitter site by controlling the doping and contacting geometry of the polysilicon layer (PSL).



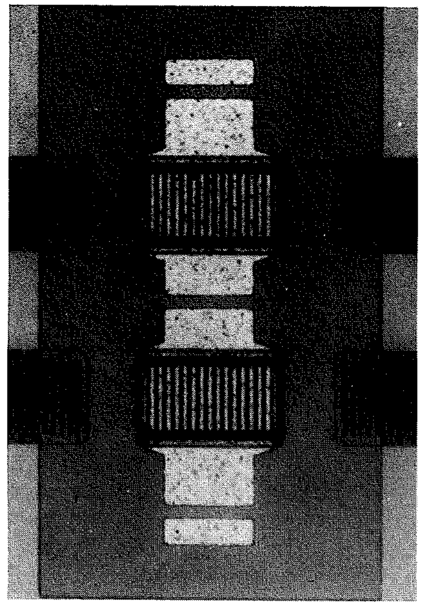


Fig. 6 (above) — The overlay geometry of this 2N6267 transistor has an effective EB/BA ratio of 6.3 relative to devices with a current-handling capability of 1 mA/mil. Fig. 7 (right) — Replacing the p+ grid with a high-temperature metal grid allows geometric EP/BA as high as 8 to be obtained.

to obtain efficient use of the emitter periphery.

New developments in overlay improve performance

Recent developments in overlay technology have resulted in higher-conductivity p^+ grids and in the use of lower-resistivity epitaxial collectors with no sacrifice in collector-to-base breakdown voltages (approximately 50 V). Combined with the introduction of PSL ballasting, these factors have permitted an increase in the current-handling capability for a given emitter periphery from 1.0 mA/mil to 1.8 mA/mil for the same geometry (*i.e.*, emitter length and grid width). These factors have increased the effectiveness of the emitter periphery. Thus, if Eq. 2 is solved for I_c assuming a given emitter periphery, the collector current may be related to an effective periphery as follows:

$$I_c = EP(I_{m/l})$$

This is defined as *EP* effective. Thus, the geometric *EP/BA* ratio of 3.5 is equivalent to an effective *EP/BA* ratio of 6.3 relative to devices with a current-handling capability of 1 mA/mil. This theory was demonstrated in the RCA-2N6267 (10 W at 2 GHz) which has double the power rating of the 2N6266 (5 W at 2 GHz) but the same geometry (Fig. 6).

A further level of development of the overlay structure can be obtained by replacing the p^+ grid with a high-temperature metal grid (Fig. 7). By minimizing the lateral diffusion problem and by improving the grid conductivity by an order of magnitude relative to the p^+ grid, closer emitter-to-grid spacing and longer emitter sites can be used. This technique permits geometric *EP/BA* design ratios as high as 8 to be obtained. The effective *EP/BA* ratio is primarily limited by the optimization of this technology. Another advantage of this variation of the overlay is that the longer emitter sites permit the use of wider emitter fingers, which permit emitter bonds to be placed directly over the emitter region (Fig. 7). This approach eliminates the emitter-to-collector MOS bond-pad capacitance which is a feed-through factor in common-base operation.

Mesh or emitter-grid structure

The mesh structure is essentially an inverse overlay geometry (Fig. 4e). It and its modifications are largely used by Microwave Semiconductor Corp. for devices up to 4 GHz, which achieves high geometric ratios of *EP/BA* by placing the emitter in the grid. Discrete base-contact areas "perforate" the emitter structure in this approach; *EP/BA* design ratios as high as 8 mils/sq. mil have been reported by MSC. This design has the additional advantage of requiring fewer processing steps than the overlay approach.

The mesh structure is not without its own limitations, however:

- The interconnected emitter grid makes it difficult to achieve effective emitter ballasting because each segment of the emitter may be fed from alternate paths.
- Although high geometric ratios of *EP/BA* have been achieved, the current-handling capability in terms of mA/mil of emitter periphery is limited by the voltage drops in the emitter grid. This results in more debiasing of segments of the emitter periphery remote from the emitter metallization than obtained with overlay or interdigitated-type structures. The voltage drops are greater than in structures where most of the emitter periphery is directly contacted by the emitter metallization. This difference can be attributed to the relatively high sheet resistance of a shallow diffused emitter and to the fact that emitter current is significantly greater than base current. Consequently, even when the resistivity of the collector is reduced, it is difficult to obtain more than 1 mA/mil and the effective *EP/BA* as defined above is typically the same as the geometric *EP/BA*.
- Finally, to minimize the debiasing in the emitter grid, the emitter and base metal fingers are spaced closer together and the widths are narrower than in the overlay structure. This places a lower limit on the thickness of the metallization which can be defined and results in higher current densities. This results in a greater susceptibility to failures from metal migration, especially when the metallization system is aluminum.

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Practical applications of the Weibull distribution to power-hybrid burn-in

L. Gallace

Weibull distribution analysis is a statistical method of examining failure mechanisms to determine optimum burn-in conditions. In this paper, a number of burn-in conditions are analyzed before the final selection is made. Finally, reliability data is shown, and the results of the inferior, or "guessed at", and superior Weibull-derived burn-in are compared.

It is not sufficient to determine burn-in conditions for solid-state devices incorporated in complex hybrids merely by arbitrarily specifying burn-in or aging conditions because very often this procedure is inefficient, performs no useful function, and uses up valuable time and facilities. In fact, one must determine in the first instance whether a burn-in will be applicable to the product in question; *i.e.*, will greater reliability be achieved with a burn-in.

The data used for analysis in this paper is taken from actual tests on the HC2000 operational amplifier, a hybrid device consisting of basically two separate sections mounted directly on an integral, common base plate. One section contains the complete driver circuit, including 23 thick-film resistors, 7 chip capacitors, 6 diode chips and 7 transistor chips on an alumina substrate. The second section contains two output-power-transistor chips and two diode chips.

Reason for screening hybrids

Manufacturing operations cannot improve the reliability of a product line beyond certain limits determined by the design of the product. In the case of power hybrids, where independent operations are performed on sections (signal circuits, power-output transistors, *etc.*) which are then interconnected before packaging (either plastic or hermetic) an approach to theoretically perfect in-process quality control would be a costly and time-consuming job.[1] Therefore, to supplement the in-process quality-control work, a thermal, power-cycling burn-in is used to identify that part of the product population that has defects. In a thermal, power-cycling burn-in, changes in power dissipation cause temperature variations which result in cyclic mechanical stresses at the interface of parts within the device because of the difference in thermal expansion of these parts.

Measuring effects of burn-in

Since the effects of a thermal-cycling burn-in are best represented by a statistical phenomenon, various types of statistical methods are used to study burn-in results. These methods include normal, chi-squared, Weibull, log-normal statistical distributions, and stress-strength distributions. However, the best of these is the Weibull method because the distribution can show a decreasing, increasing, or constant failure rate. It is important to know if the failure rate is constant, increasing, or decreasing with time because a burn-in can only achieve higher reliability when a decreasing failure rate exists. For example, a constant failure rate only reduces the amount of product available for use after burn-in, and an increasing failure rate indicates that the burn-in conditions have taken the devices to the wearout part of the life cycle. Thus, with the Weibull method, one can determine whether his burn-in conditions are correct.

Derivation of Weibull function

The mathematical derivation of the Weibull function has been handled elsewhere in very good detail.[2] The Weibull density function for the random variable X is:

For $x \geqslant r$,

$$f(x) = \frac{\beta}{\eta}\left(\frac{x-r}{\eta}\right)^{\beta-1}\{\exp[-(\frac{x-r}{\eta})^{\beta}]\}$$

For $x < r$,

$$f(x) = 0$$

and the cumulative distribution function is

$$F(x) = 1 - \exp[-(\frac{x-r}{\eta})^{\beta}] \qquad (1)$$

where η, $\beta > 0$; β is the shape parameter; η is the scale parameter; and r is the location parameter.

The location parameter (r) implies that an item failure cannot start until a product has been under test for r hours. Most often, the location parameter is assumed to be equal to zero, but this is not necessarily a good assumption, as the value depends on the type of product being tested. If the location parameter is set equal to zero, then the two-parameter Weibull is used. The cumulative distribution function for the two parameter

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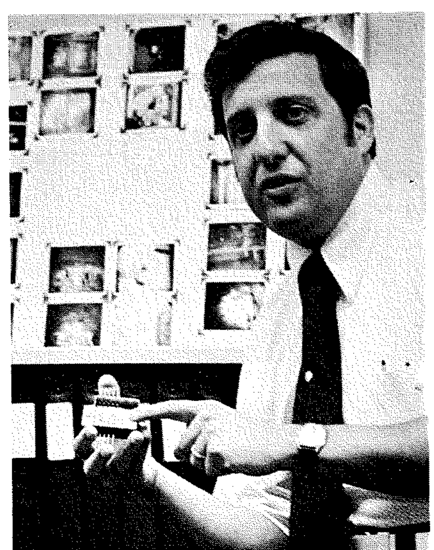

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Weibull is:

$$F(x) = 1 - \exp[-(x/\eta)^{\beta}]$$

Since the Weibull distribution is most often skewed, the location of the mean life depends upon the amount of skewness (shape parameter, β). The mean does not coincide with the median except when $\beta = 3.5$. The characteristic life, however, can always be determined from the same percentile estimate regardless of the value of the shape parameter. For example, if $x = \eta$ in $F(x)$,

$$F(\eta) = 1 - e^{-1} = 0.632$$

or the point where 63.2% of the product has failed in terms of cycles or hours. This value is known as the characteristic life (scale parameter), and can always be determined from the 63.2 percentile of any Weibull distribution.

Derivation of Weibull graph paper

From Eq. 1, which defines the cumulative distribution function (CDF) of the Weibull distribution,

$$1 - F(x) = \exp\left[-\left(\frac{x-r}{\eta}\right)^{\beta}\right] \qquad (2)$$

Now taking the reciprocal of Eq. 2, and the natural logarithm twice:

$$\ln \ln \{1/[1-F(x)]\} = \beta \ln [(x-r)/\eta] \qquad (3)$$

and

$$\ln \ln \{1/[1 - F(x)]\} = \beta \ln (x - r) - \beta \ln (\eta) \qquad (4)$$

Let

$$Y_1 = \ln \ln \{ 1/[1-F(x)]\}$$

$$a = \beta$$
$$b = - \beta \ln (\eta)$$

and

$$x_1 = \ln (x-r)$$

Then, from Eq. 4

$$Y_1 = a\, x_1 + b$$

which is linear. Therefore, a true Weibull sample or population can be plotted as a straight line.

If the plot is not straight, it can sometimes be straightened by subtracting the proper constant from the data, as can be seen from $x_1 = \ln(x-r)$. The proper constant is simply that value which straightens the line. This constant, already discussed above in connection with the derivation of the Weibull function, is known as the location parameter, r, or minimum life. A sample of the Weibull paper is shown in Fig. 1.

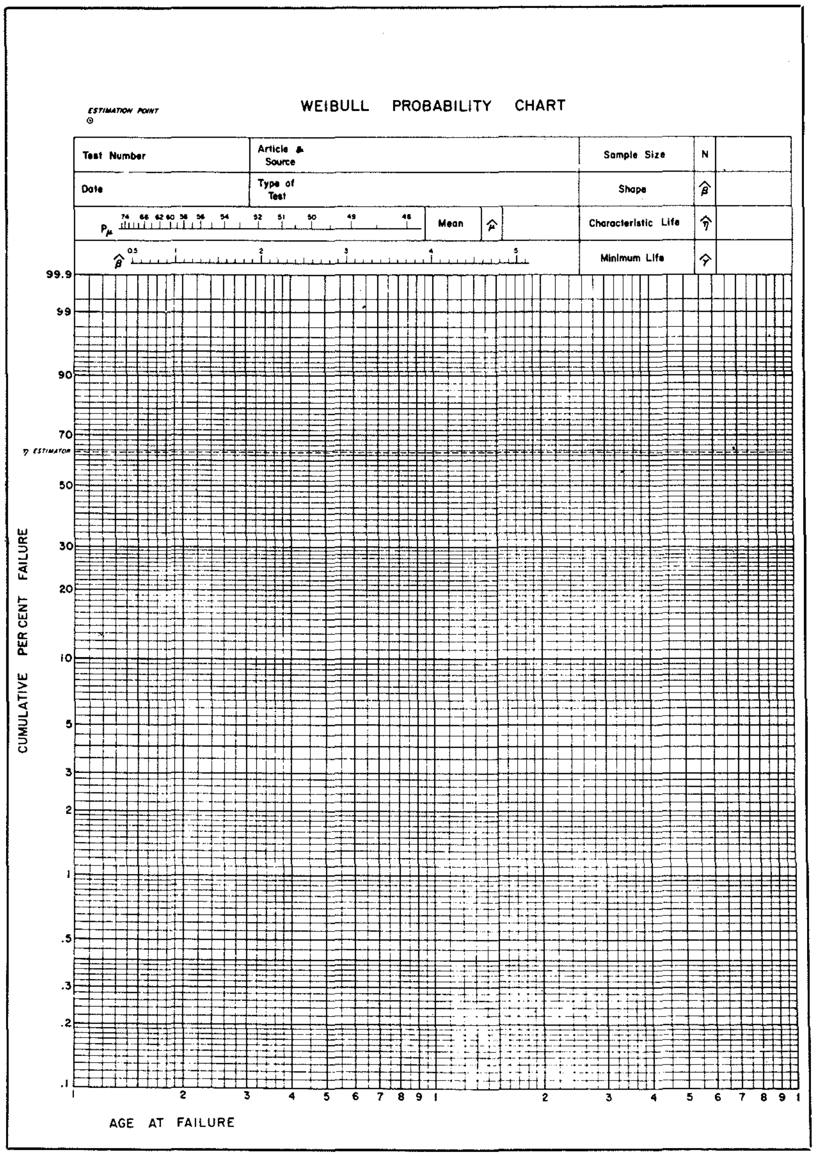

Fig. 1 — Weibull probability chart paper.

Practical application of the Weibull method

The first burn-in applied to a sample of hybrid amplifiers was a power cycling typical of what the device would see in an application (20W, T_c = 40 to 60°C). However, life tests performed to evaluate operating life after the power burn-in produced early failures which, when analyzed, were typical of what should have been culled out in the burn-in if proper conditions had been used. This is an essential point, because for a screen or burn-in to be effective, early-hour life-test failures must be eliminated. Therefore, the final decision on a burn-in must be made after burned-in devices are subjected to long-term life tests and the failure distribution analyzed. Consequently, either the initial burn-in was not efficient or the mechanisms of failure were not the type that could be screened.

Usually, the type of mechanisms which cannot be screened are design oriented; *i.e.*, a burn-in of a marginal design will screen out some units and weaken others to the point where they will fail early in service life. Thin silicon chips which will crack easily and insufficient bond area for contacts on chips are two examples of design failures which normally cannot be screened. Failure analysis of the mechanisms resulting from the initial burn-in indicated, however, that they were predominantly of the process-control and assembly type and, therefore, capable of being screened.

At this point it was decided to look at the failure distribution of the amplifiers at higher power levels. The first level (30W, T_c = 35°C to 100°C), shown on the Weibull paper in Fig. 2, curve A, plotted as a straight line.* Failure analysis showed three different types of failures occurring at different periods in the cycle life, and a further increase in the power-cycling level (40W, T_c – 35°C to 135°C) showed that the straight-line failure distribution was actually composed of three distinct failure regions, curve B, each characterized by a different Weibull shape parameter, *B*.

Fig. 3 shows a plot of failure rate versus time for three different values of Weibull beta parameters;[3] note that only a Weibull distribution with a $\beta < 1$ will

*From the theory on order statistics it can be shown that a good estimate of the cumulative percent-failure points would be:

cumulative percent $= i/(n+1)$

where i is the i^{th} order failure and η is the sample size.

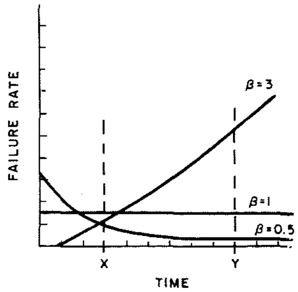

Fig. 3 — Failure rate for three beta parameters.

have a failure rate decreasing with time.

To determine the value of the shape parameter β of the data plotted in Fig. 2, a draftsman's right triangle is used to draw a line perpendicular to the plotted line and passing through the estimation point at the top, left-hand corner of the paper. The β or scale parameter is then read from the scale labelled β. The $P\mu$ scale indicates the percentile at which the mean should be read for the β parameters determined. The characteristic life, which has been discussed above, is always determined from the 63.2% line; this is the "η estimator" line on the Weibull paper. The value of the minimum life parameter, r, defined above in this paper as the location parameter, is the value of the constant used, if any, to straighten out the data.

The Region I failures at a shape parameter equal to 0.5 show a decreasing failure rate, and, through failure analysis, these failures are shown to be due to the process-control and assembly of the product. Regions II and III are the result of wearout mechanisms because they show a statistically increasing failure rate as verified by failure analysis.

Region I failures are those that are to be eliminated, and 100 cycles is dictated by technical and economic reasons. Technically, the break between the decreasing failure rate, $\beta = 0.5$, and the increasing failure rate, $\beta = 1.1$, regions is approximately 150 to 200 cycles. However, 100 cycles can be accumulated in an 8-hour period (12.5 cycles/hr), which allows scheduling problems on a factory-shift basis to be minimized.

Groups of product were selected and evaluated on both the original burn-in conditions and the new Weibull-derived conditions. Table I contains the reliability-test results compiled after the product has been burned-in under both "guessed at" and Weibull-determined conditions. The table shows, in the *total failures/cause* column, that early failures have been eliminated and that no long-term effects have been introduced as a result of the higher-level, power-cycling test. The latter is indicated by the fact that no failures occurred after 1000 hours. The new burn-in, under Weibull-determined conditions, has been demonstrated, over the long run, to be 3 to 4 times more efficient than the original "guessed at" burn-in for culling out early failures.

Failure analysis

Region I failures were shown by failure analysis to be:

1) Gold leaching of lead frame contacts on signal alumina substrate;
2) Poor wire bonds on signal chips;
3) Chip capacitor failures (cracked);
4) Epoxy mold in contact with substrate; and
5) Poor location of *E-B* clips on output transistors.

Photographs of these defects are shown in Figs. 4a through 4e. Improved inspection is eliminating more of these failures in addition to pointing out new ways to improve processing techniques.

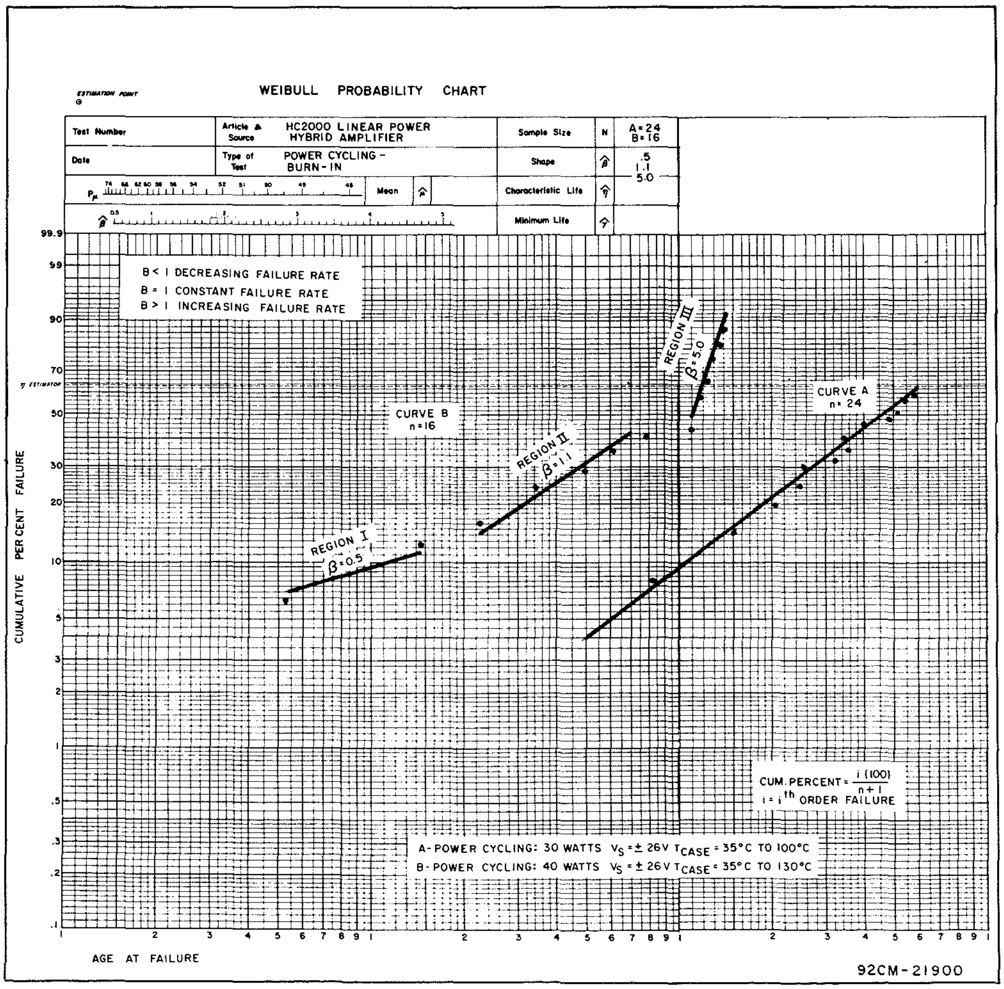

Fig. 2 — Power-cycling burn-in for HC2000 linear power hybrid amplifier.

Table I — Original burn-in versus Weibull-derived burn-in.

Test	*Screening conditions*	*Devices tested*	*Total cycles or hrs*	*Total failures/cause*	*Failing cycles/hrs.*
±26.5V, 13V rms	Original burn-in	8	5000 cycles	1/(open)	1 to 720 cycles
T_c = 40 to 80°C	Weibull burn-in	12	5000 cycles	0	
Operating life					
±27V, 13V, rms	Original burn-in	12	3166 hrs.	1/(low ac output)	250 hours
T_c = 100°C	Weibull burn-in	12	3166 hrs.	0	
Temperature cycle					
−65°C to 150°C	Original burn-in	12	200 cycles	1/(pos. clipping)	1 to 10 cycles
	Weibull burn-in	12	200 cycles	0	
150°C *shelf life*					
	Original burn-in	6	3024 hrs	0	
	Weibull burn-in	9	3024 hrs.	0	

Original burn-in: 20W, T_c = 40°C to 60°C
Weibull burn-in: 40W, T_c = 35°C to 135°C

Conclusions

No product screen on burn-in should be applied to a semiconductor product line unless the proper statistical evaluation, coupled with an analysis to determine failure mechanisms, is performed. The Weibull distribution analysis is a good graphical method to analyze the failure distribution for power-cycling testing.

Proper levels of burn-in must be chosen so that in the case of hybrids, where different failure mechanisms exist, the burn-in condition will distinguish between decreasing and increasing failure-rate regions.

Definitions

1. *Burn—in*—The operation of an item to stabilize its failure rate.
2. *Thermal cycling*—The *on-off* power operation causing variations in device temperatures.
3. *Thermal fatigue*—A wearout type of failure that may occur in power devices as a result of thermal-power cycling causing cyclic mechanical stresses because of differences in thermal expansion of the materials of the device.
4. *Probability density function*—The relative probability of the rate at which a variable will occur.
5. *Cumulative distribution function*—The probability that the random variable x takes on any value less than or equal to the stated value of x.
6. *Distribution*—The arrangement of a set of numbers.
7. *Distribution function of lifetimes*—The probability that a device starts at time (0) and fails at time (t).
8. *Mechanism of failure*—The physical process which results in device failure.
9. *Failure*—The probability of a failure per unit of time of the devices still operating.
10. *Screening tests*—Tests employing non-destructive environmental, electrical, and/or mechanical stresses to identify defective items.
11. *Statistical analysis*—The mathematical treatment and evaluation of statistical information.
12. *Random variable*—A function defined over the sample space of an experiment.
13. *Failure analysis*—A detailed study to determine the mechanism of failure.
14. *Early failure period*—An interval, immediately following final testing, during which the failure rate of devices is relatively high.
15. *Hybrid*—A combination of active semiconductor technology joined to a passive network of conductors, capacitors, and inductors with either thickfilm or thin-film techniques.
16. *Order statistic*—Analysis performed on data when they are arranged in order of increasing or decreasing magnitude.

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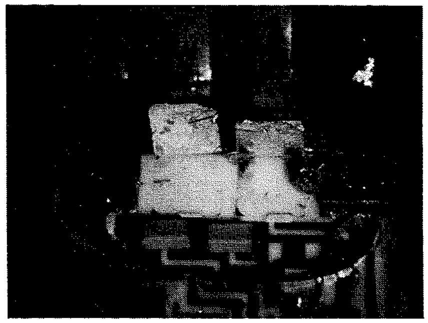
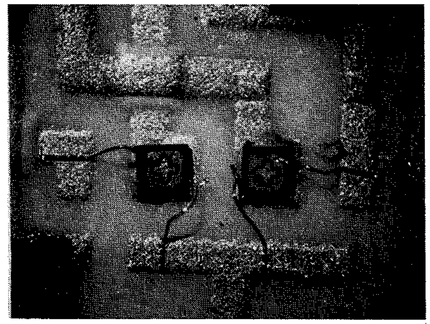
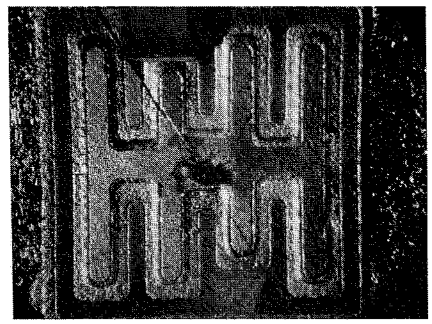


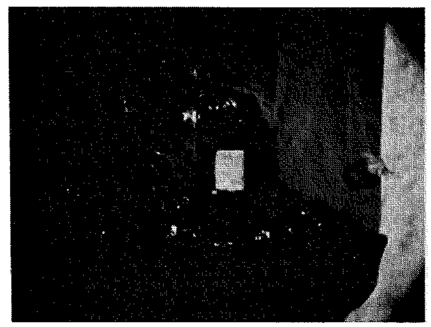
a) Au leaching of lead contacts on Al substrate.



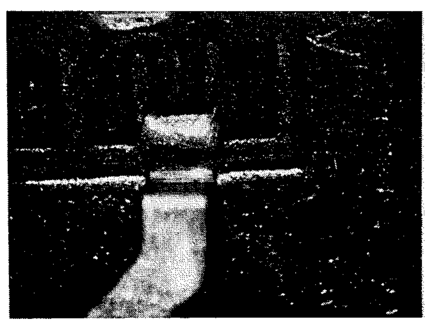
b) E bond wires lifted Q1 and Q2; Al metallization fractured under bonds.



c) Cracked pellet.



d) Epoxy mold in contact with areas of Q3 and Q4 and D1 and D2.



e) Q11 open base contact due to clip mounted off center at edge pellet.

Fig. 4 — Region I failures.

Parametric solutions for nonlinear equations (the catenary)

Dr. R. D. Scott

The equation of the catenary is studied for a set of values of various parameters that describe its geometry and tension level. As the locations of the supports change, the catenary undergoes changes in relative sag, tension level, and location of the axis of symmetry. Although all pertinent values are readily obtainable by standard methods if any two independent parameters are given, it is instructive to be able to follow the behavior graphically as certain parameters are held constant. A set of design curves is included as an example. The correspondence between a simple hanging cable and a neutrally buoyant cable towed through a fluid is also investigated, and techniques of computer solution and plotting of the equations are discussed.

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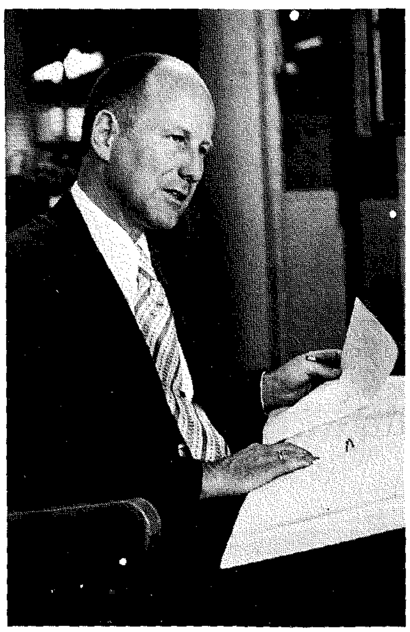

THE CATENARY is characteristic of a class of problems wherein a fundamental situation or process can be described by an explicit relationship, or set of relationships, but the study of the behavior of the process requires detailed consideration of a number of auxiliary derived functions. The generation of the auxiliary relationships may or may not be difficult, but often the complete understanding of their behavior (and the implications for the underlying process) rests on the appropriate choice of a set of functions. The study of the various functions is frequently done by graphical means; the method of generating and plotting the functions is therefore of significant interest, especially when a large number of curves is involved.

Where full computing service is available, especially with the availability of a sophisticated X-Y plotter, the problem of constructing a set of curves is simple, if not inexpensive. On the other hand, for the investigator with limited facilities (*e.g.*, a time-sharing terminal without a continuous-line plotter), there is nevertheless a procedure that will yield excellent results with high accuracy.

The following study of the catenary illustrates what can be done, using relatively simple tools, and its methods can be applied to a wide variety of problems.

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Flexible hanging string

The equation of a perfectly flexible string, hanging from two supports under its own weight is familiar[1]:

$$\eta = \cosh \xi - 1 \qquad \eta \equiv Y/a \qquad \xi \equiv X/a \tag{1}$$

Fig. 1 shows the relevant parameters to be studied. The angles θ_1 and θ_2 uniquely determine the shape and tension of the catenary since:

$$T_1 \sin \theta_1 - T_2 \sin \theta_2 = 0$$
$$T_1 \cos \theta_1 - T_2 \cos \theta_2 = W \tag{2}$$

from equilibrium considerations.

Eqs. 2 have a unique solution for T_1 and T_2 if $\theta_1 \neq \theta_2$. Also:

$$\text{ctn } \theta_1 = \left.\frac{d\eta}{d\xi}\right|_{\xi = \xi_1} = \sinh \xi_1$$
$$\text{ctn } \theta_2 = \left.\frac{d\eta}{d\xi}\right|_{\xi = \xi_2} = \sinh \xi_2 \tag{3}$$

which determines the geometry as a function of slope.

Now ξ (hence η) is a unique function of θ, if $0 < \theta < \pi$. Therefore, it is possible to represent the equilibrium configuration (or geometry) of any catenary as a single point: $P = P(\theta_1, \theta_2)$. Conversely, if $P(\theta_1 \theta_2)$ is given, the values of all other dimensionless parameters are implied because the geometry is fixed.

For example, the independent parameters $L/L^*, \theta^*, a/L^*, T_1/W, T_1/T_2, X_0/X^*, Y_0/Y^*$ are all functions of θ_1 and θ_2. The parameter T_1/W can be obtained from Eq. 2, and the remaining parameters by construction from Eq. 1. Specifically:

$$\frac{L}{L^*} = \frac{\text{ctn } \theta_1 - \text{ctn } \theta_2}{\left\{\left[\ln \dfrac{\tan \theta_2/2}{\tan \theta_1/2}\right]^2 + (\csc \theta_1 - \csc \theta_2)^2\right\}^{1/2}} \tag{4}$$

$$\frac{a}{L^*} = \left\{\left[\ln \frac{\tan \theta_2/2}{\tan \theta_1/2}\right]^2 + (\csc \theta_1 - \csc \theta_2)^2\right\}^{-1/2} \tag{5}$$

$$\frac{T_1}{W} = \frac{\sin \theta_2}{\sin (\theta_2 - \theta_1)} = \frac{\csc \theta_1}{(\text{ctn } \theta_1 - \text{ctn } \theta_2)} \tag{6}$$

$$\frac{T_1}{wX^*} = \frac{\csc \theta_1}{\ln \dfrac{\tan \theta_2/2}{\tan \theta_1/2}} \tag{7}$$

$$\theta^* = \tan^{-1} \left[\frac{\ln \dfrac{\tan \theta_2/2}{\tan \theta_1/2}}{(\csc \theta_1 - \csc \theta_2)}\right] \tag{8}$$

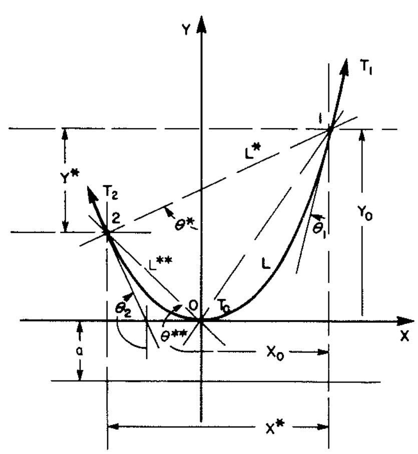

Fig. 1 — Standard catenary.

Symbols

- $\eta = Y/a$
- $\xi = X/a$
- a parameter of the catenary
- θ_1 angle between cable tangent and vertical (clockwise), at the right end
- θ_2 same as θ_1, but for left end
- θ angle between cable tangent and vertical
- θ^* slope of the line connecting the supporting locations (measured from vertical) $= \tan^{-1} (X^*/Y^*)$
- L^* length of line connecting the supporting locations
- T tension
- X^* horizontal distance between supporting locations
- Y^* vertical distance between supporting locations
- X_0 horizontal distance from lowest point to right support
- Y_0 vertical distance from lowest point to right support
- L cable length
- W cable weight (lb)
- w cable weight (lb/ft)
- ρ fluid density
- V velocity
- C_d drag coefficient
- d cable diameter
- $D^* = \frac{1}{2} \rho V^2 C_d d$ lb/ft (fluid normal form drag)

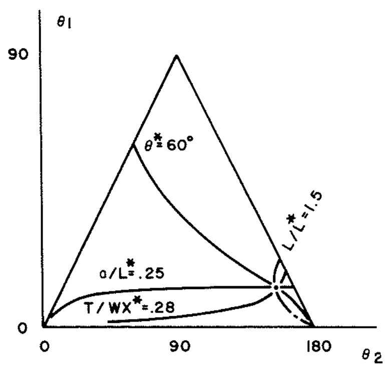

Fig. 2 — Equilibrium chart.

$$\frac{L}{X^*} = \frac{(\text{ctn } \theta_1 - \text{ctn } \theta_2)}{\ln \dfrac{\tan \theta_2/2}{\tan \theta_1/2}} \tag{9}$$

Fig. 2 shows the equilibrium state of the catenary of Fig. 1, as an example. The lines of constant parametric value for θ^*, L/L^*, a/L^*, and T_1/wX^* that pass through the equilibrium point are included. Note that the range $0<\theta_1<90°$, $0<\theta_2<180°$ covers all possible catenary geometry, since the roles of θ_1 and θ_2 can be reversed.

The significance of the lines of constant parametric value is that by following one of these lines, the behavior of θ_1 and θ_2 for hat value (and all other parameters) is noted directly. For example, all possible catenaries (equilibrium states) for which $T_1/wX^* = 0.28$ (constant tension) can be found by following the $T_1/wX^* = 0.28$ line shown in Fig. 2. All possible catenaries for which the slope of the line between the attachment points equals 60° can be found in the same way.

Although it is possible to write the equation of the catenary in physical X-Y coordinates (referred to one end), given θ_1 and θ_2 (from Eqs. 1 and 3), the items that are generally of interest are the tension, relative sag, location of the lowest point, *etc.* Given a full set of curves for each parameter, any catenary can readily be investigated by graphical means, once any two independent parameters are prescribed: the equilibrium point for the catenary lies at the intersection of the two corresponding parametric curves. Thus, in Fig. 2, it would be possible to determine the equilibrium point for the catenary by specifying (for example) any of the following pairs of values: $\theta_1 = 13.5°$, $\theta^* = 65°$; $\theta^* = 65°$, $T_1/W = 0.26$; $\theta_2 = 153.5°$, $a/L^* = 0.25$. The choice of a particular pair of parameters is determined by the formulation of the specific problem.

A further use of a set of parametric curves is the study of the sensitivity of certain variables to changes in others. For instance, suppose it is desired to find the influence of θ^* on tension as the catenary is extended at point 1. This requires the study of terms of the type

$$\left.\frac{\partial\,(T_1/W)}{\partial\theta^*}\right|_{\theta_2}$$

which can be evaluated graphically at any equilibrium point. This procedure is relatively difficult to accomplish algebraically (see Eqs. 6 and 8), but values suitable for survey purposes can easily be obtained from the chart.

Fig. 3 shows a comprehensive set of parametric curves for the catenary. Other families of curves are possible, but those included admit most calculations of interest.

The following two examples illustrate the use of the charts:

Example 1

Given the locations of the points of attachment (X=0, Y=0), (X=17.32, Y=10), what is the geometry and tension for a sag of 0.05? (Sag $\equiv L/L^*-1$)

First compute $\theta^*=\tan^{-1}(\Delta X/\Delta Y)$ = 60°, $L^*=(\Delta X^2+\Delta Y^2)^{1/2}$ = 20, $L/L^*=1+0.05=1.05$. Therefore the cable length is 1.05 L^*, and the equilibrium point is located at the intersection of L/L^*=1.05 and θ^*=60°. This results in values for θ_1=35.5°, θ_2=96°, T_1/wX^*=0.81, T_1/T_2=0.58.

Consider next the portion of the cable to the right of the lowest point (if such exists). This section of cable is itself in equilibrium, and has the parameters θ_1=35.5°, θ_2=90°. The new values of θ^*, L/L^* for this cable section are then read off, and the procedure is repeated for the other section of cable. From Fig. 1 it is seen that for the left section

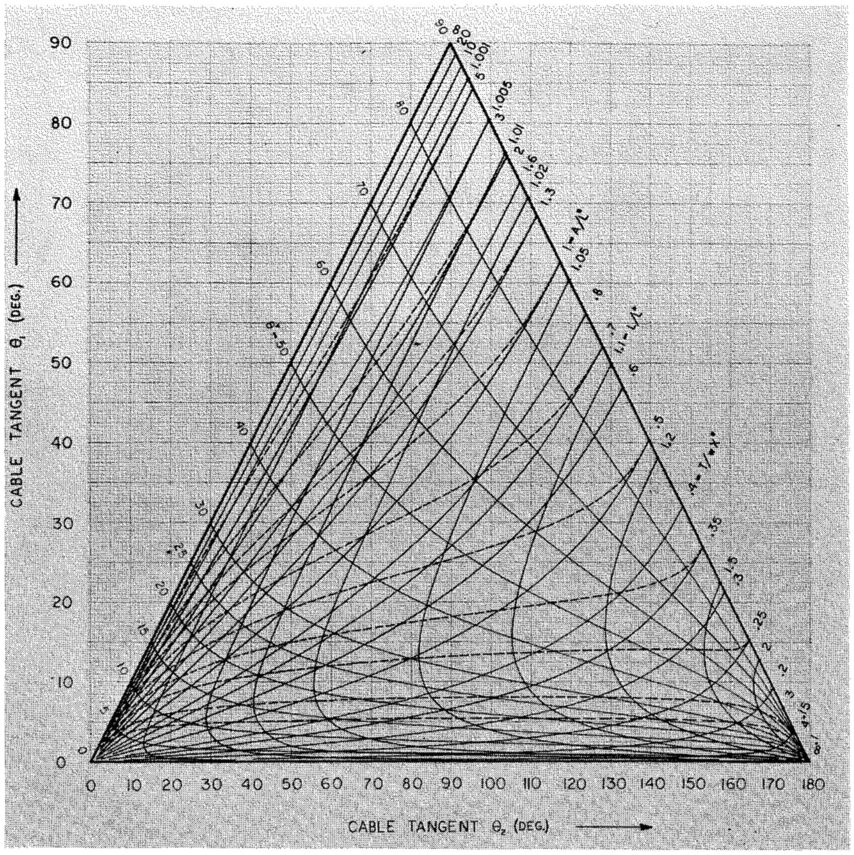

Fig. 3 — Parametric curves for the catenary (part 1).

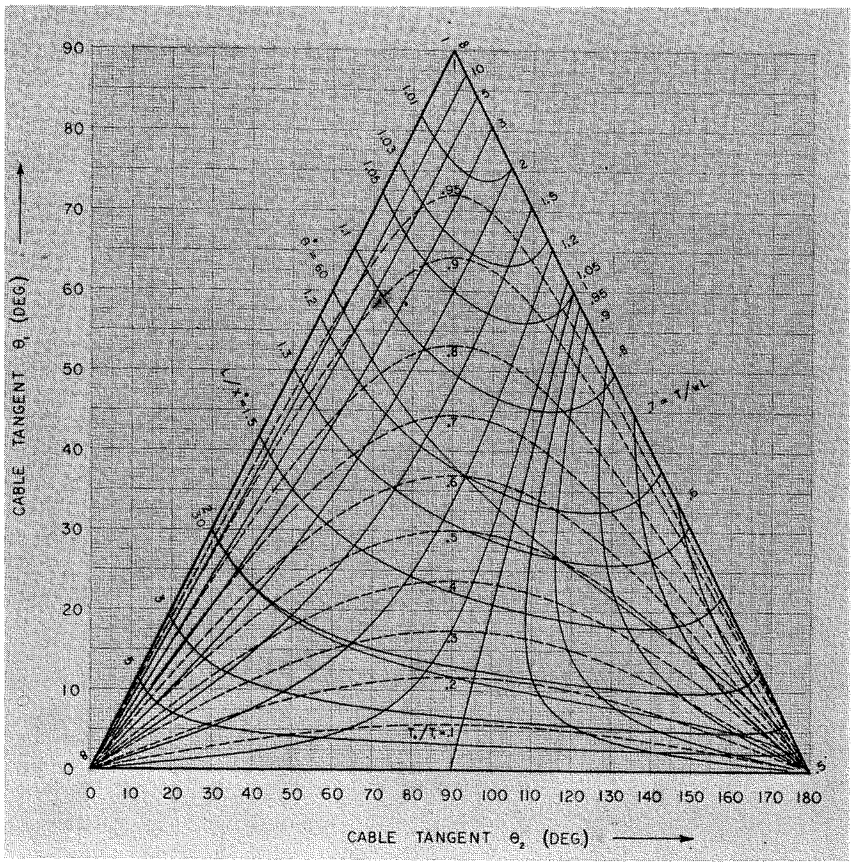

Fig. 3 — Parametric curves for the catenary (part 2).

$$L^{**} \sin \theta^{**} = X^{*} - X_0$$
$$L^{**} \cos \theta^{**} = Y^{*} - Y_0$$
or
$$X_0 = Y_0 \tan \theta^{*} \quad \text{(no minimum)} \qquad (10)$$

(X_0/X^* and Y_0/Y^* could thus be plotted by this approach.)

Example 2

Given a steel tape of length L (chaining problem) whose ends are inclined at an angle θ^* and whose upper tension is T_1, find the slant length, L^*.

Compute T_1/W, and find L/L^* at the intersection of the T_1/W and θ^* curves. Then $L^* = L/(L/L^*)$. If the tension at the lower end is known instead, compute T_2/W, and iterate to the proper value of T_1/W and θ^* via the T_1/T_2 curves. $[T_1/W = (T_2/W)(T_1/T_2)]$

Towed cable

When a fully submerged cable (neutrally buoyant) is towed in a fluid, as in Fig. 4, its shape as determined by form and skin drag is also a catenary.[2] The towed cable and the hanging chain have identical geometry if the direction of fluid flow is the same as the direction of gravity, *i.e.*, parallel to the axis of symmetry. Therefore, the parameters representing geometry (a/L^*, L/L^*, *etc.*) also have the same values as a function of θ_1, θ_2. The only change required is to replace W with $\frac{1}{2}\rho V^2 C_d dL$ (D^*L) for finding tension values. Furthermore, the tension (T_1) computed for the towed cable under this formulation applies only to the case of zero skin friction, is constant along the cable, and equals (T_1/W).

For non-zero skin friction, the tension curves must be modified. However, the tension at the end of the cable opposite the tow point can usually be determined by the attached devices (such as a buoy, hydrofoil, and the like), and for relatively short cables at $\theta_1 > 10°$, the error from neglecting the skin friction is small. The application of constant L/X^* curves is of particular interest here, since X^* plays the role of depth; θ_1 is the tangent angle at the tow point, and θ_2 the tangent angle at the opposite end.

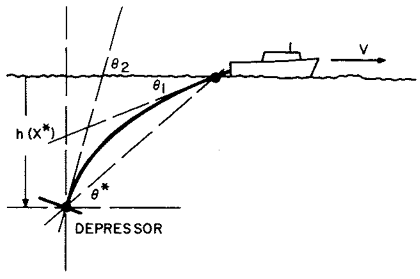

Fig. 4 — Towed cable.

Example 3

Suppose a submarine at 100-ft depth is towing a buoy at the surface at 10 knots, such that the lift/drag ratio of the buoy is 1. If D^* is 20, and the maximum strength of the cable is 5000 lb., what is the shortest permissible cable?

Since the lift/drag ratio of the buoy is 1, equilibrium requires that $\theta_2 = 45°$ [$\tan^{-1}(1)$]. The quantity T_1/wX^* is equivalent to T_1/D^*h, and the values chosen give $T_1/D^*h = 2.5$. The intersection of $\theta_2 = 45°$, and $T_1/wX^* = 2.5$ gives $\theta^* = 37°$, and $L/L^* = 1.003$. From this, we find $L = 167$ ft.

Plotting constant-parameter curves

Although accurate and fast plotting can be done by computer, this service is not always available on a convenient basis. Moreover, computer plotting in any event usually demands a nontrivial amount of programming: the equations to be solved and the solution method must be accounted for, as well as the plot format, interval, and scale. The alternative is manual plotting from tabulated data. It is instructive to compare the attainable accuracy of the two methods.

Graphic display of data implies a limit of accuracy of retrievable data that depends on several factors:

- The width of the line;
- The ability of the observer to interpolate values between grid lines;
- The accuracy of original plotting; the spacing of parametric values; and
- The accuracy of calculation or observation of source data.

In plotting data points by hand (without a magnifying glass) it is reasonable to be able to locate a point within 0.02 mm (± 0.01 mm) on a one-line/mm grid with a pin prick, and to construct a curve through that point so that the apparent center of the line represents the theoretical value to within ± 0.02 mm (± 0.005 in.). On an original format of 20 cm overall, this implies an achievable locating accuracty of ± 0.1% (full scale), with a careful manipulation. X-Y plotters can better this figure only with careful setup, in general, on a point-by-point basis.

Interpolation seriously degrades this figure, however, whether the plot is constructed manually or mechanically. Proper interpolation between curves requires a cross-plot, which involves the judgment of the observer; the remaining impact on accuracy depends on the spacing reference curves. Ideally, the spacing should be close enough to allow linear interpolation to the same degree of accuracy as the plot, but this is seldom possible from both graphic and economic considerations. The best service that the author of a set of curves can render is to include a useful set of data, plotted with an accuracy consistent with the intended use, with a spacing that allows smooth and consistent interpolation. Under these guidelines, hand plotting is essentially as accurate as machine plotting, in terms of useful information that can be extracted.

The data in Fig. 3 were plotted manually, using a template constructed for each curve. The accuracy was held generally to ½ of a minor scale division for each point plotted (1 part in 720 along the θ_2-axis). After plotting sets of curves for θ_1 *vs.* θ_2, the locations of a selected number of intersections of sets of curves were checked to verify accuracy. The method of calculating these intersections in instructive, since it is quite general and can be applied to many situations. The same basic computer program was used each time, in this case, with minor modifications.

Consider the intersection, $P(X^*, Y^*)$, of two functions in X and Y with a parameter:

$$f_1(X^*, Y^*, \alpha^*) = f_2(X^*, Y^*, \beta^*) \qquad (11)$$

where

$$f_1(x, y, \alpha) = 0$$
$$f_2(x, y, \beta) = 0 \qquad (12)$$

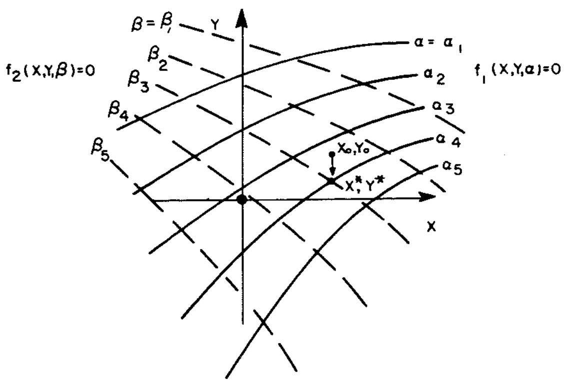

Fig. 5 — Intersection of two functions.

Fig. 5 illustrates this case graphically. If the point $P(X_0, Y_0)$ is chosen in the neighborhood of $P(X^*, Y^*)$ then the values of $f_1(X_0, Y_0, \alpha^*)$ and $f_2(X_0, Y_0, \beta^*)$ are generally non-zero, and can be considered as errors arising from the choice of X_0 and Y_0. Moreover, they can be approximated as a linear function of X_0, Y_0:

$$\epsilon_1 \equiv f_1(X_0,\ Y_0,\ \alpha^*) \approx \frac{\partial \epsilon_1}{\partial X^*}\bigg|_{Y^*,\ \alpha^*} (X_0 - X^*) + \frac{\partial \epsilon_1}{\partial Y^*}\bigg|_{X^*,\ \alpha^*} (Y_0 - Y^*)$$

$$\epsilon_2 \equiv f_2(X_0,\ Y_0,\ \beta^*) \approx \frac{\partial \epsilon_2}{\partial X^*}\bigg|_{Y^*,\ \beta^*} (X_0 - X^*) + \frac{\partial \epsilon_2}{\partial Y^*}\bigg|_{X^*,\ \beta^*} (Y_0 - Y^*) \tag{13}$$

where

$$\frac{\partial \epsilon_1}{\partial X^*}\bigg|_{Y_0,\ \alpha^*} \equiv \frac{\partial}{\partial X} f_1(X^*, Y^*, \alpha^*);\ etc.$$

As $P(X_0,\ Y_0) \rightarrow P(X^*,\ Y^*)$,

$$\frac{\partial f_1}{\partial X_0}\bigg|_{Y_0,\ \alpha_0} \rightarrow \frac{\partial f_1}{\partial X^*}\bigg|_{Y^*,\ \alpha^*} \quad etc.$$

and the pair of Eqs. 13 can be solved for the quantities $(X_0 - X^*)$, $(Y_0 - Y^*)$ as follows:

$$J = \begin{vmatrix} \dfrac{\partial f_1}{\partial X_0} & \dfrac{\partial f_1}{\partial Y_0} \\ \dfrac{\partial f_2}{\partial X_0} & \dfrac{\partial f_2}{\partial Y_0} \end{vmatrix} \neq 0$$

$$(X_0 - X^*) \approx \frac{1}{J} \begin{vmatrix} \epsilon_1 & \dfrac{\partial f_1}{\partial Y_0} \\ \epsilon_2 & \dfrac{\partial f_2}{\partial Y_0} \end{vmatrix} \tag{14}$$

$$(Y_0 - Y^*) \approx \frac{1}{J} \begin{vmatrix} \dfrac{\partial f_1}{\partial X_0} & \epsilon_1 \\ \dfrac{\partial f_2}{\partial X_0} & \epsilon_2 \end{vmatrix}$$

The estimate for X^*, Y^*, given values at X_0, Y_0 is then:

$$X^* \approx X_0 - (X_0 - X^*)$$
$$Y^* \approx Y_0 - (Y_o - Y^*) \tag{15}$$

If the initial value of X_0 and Y_0 is sufficiently close to X^* and Y^*, then iteration of the process implied in Eqs. 14 and 15 will yield a converged solution (Newton - Raphson method). In many cases, the initial estimate may be quite far removed from the solution, without compromising the process.

As an example, consider the problem of verifying the intersections of the set of curves for $a/L^* = 0.25$, and $\theta^* = 5°, 10°, \cdots, 85°$. The equations that determine a/L^* and θ^* were given in Eqs. 5 and 8. It is convenient to define $f_1(\theta_1, \theta_2, \theta^*)$ and $f_2(\theta_1, \theta_2, a/L^*)$ as follows:

$$f_1(\theta_1, \theta_2, \theta^*) = \frac{\ln \dfrac{\tan \theta_2/2}{\tan \theta_1/2}}{(\csc \theta_1 - \csc \theta_2)} - \tan \theta^* \equiv \epsilon_1 \tag{16}$$

$$f_1(\theta_1,\ \theta_2,\ a/L^*) = [\ \ln^2 \frac{\tan \theta_2/2}{\tan \theta_1/2} + (\csc \theta_1 - \csc \theta_2)^2] - (a/L^*)^2 \equiv \epsilon^2$$

The derivatives, as functions of θ_1 and θ_2, are found to be:

$$\frac{\partial \epsilon_1}{\partial \theta_1} = -\csc \theta_1 \sec \theta^* (a/L^*)\ [1 - \tan \theta^* \operatorname{ctn} \theta_1]$$
$$\frac{\partial \epsilon_1}{\partial \theta_2} = \csc \theta_2 \sec \theta^* (a/L^*)\ [1 - \tan \theta^* \operatorname{ctn} \theta_2]$$
$$\frac{\partial \epsilon_2}{\partial \theta_1} = 2 \csc \theta_1 \sin \theta^* (a/L^*)^3\ [1 + \operatorname{ctn} \theta^* \operatorname{ctn} \theta_1]$$
$$\frac{\partial \epsilon_2}{\partial \theta_2} = -2 \csc \theta_2 \sin \theta^* (a/L^*)^3 [1 + \operatorname{ctn} \theta^* \operatorname{ctn} \theta_2] \tag{17}$$

(θ^* and a/L^* are evaluated at the particular values of θ_1 and θ_2)

Computer analysis

A computer run, listing values of θ_1 and θ_2 that give the computed values of θ^* and a/L^* shown, demonstrates the effectiveness and accuracy of the method:

θ_1	θ_2	a/L^*	θ^*
14.81	163.71	2.500-01	85.00
14.41	162.49	2.500-01	80.00
14.18	160.84	2.500-01	75.00
14.08	158.62	2.500-01	70.00
14.07	155.63	2.500-01	65.00
14.11	151.61	2.500-01	60.00
14.15	146.16	2.500-01	55.00
14.16	138.79	2.502-01	50.00
14.07	128.83	2.500-01	45.00
13.86	115.67	2.500-01	40.00
13.48	99.07	2.500-01	35.00
12.90	79.73	2.500-01	30.00
12.07	59.62	2.500-01	25.00
10.92	41.16	2.500-01	20.00
9.36	25.97	2.500-01	15.00
7.22	14.39	2.500-01	10.00
4.22	5.98	2.500-01	5.00

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Recent advances in wideband recording systems

J.S. Griffin

The RCA Airborne Dual-channel Video Severe Environment Recorder (ADVISER) family of airborne/portable recorder/reproducers represents a new concept in design and construction of video tape recorders. The total functional modularity of the equipment results in high reliability and high maintainability, with state-of-the-art performance.

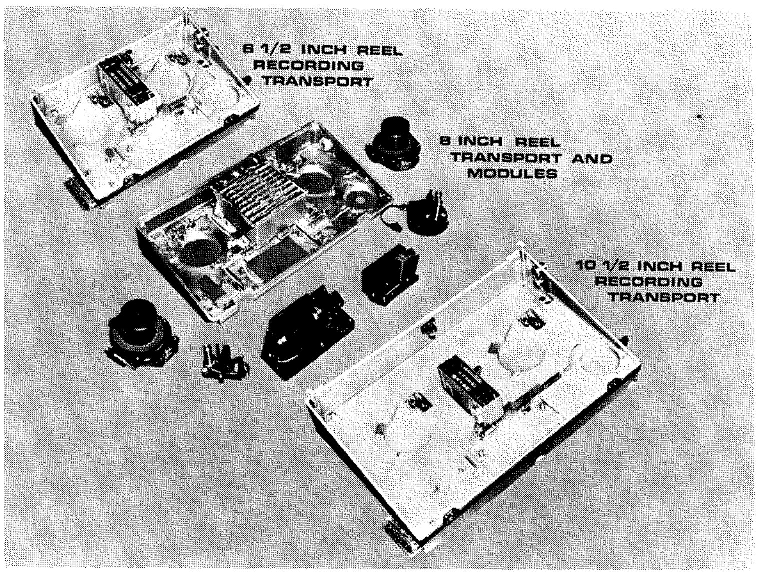

Fig. 1 — ADVISER transport modularity.

THE FUNDAMENTAL CONCEPT for design of the ADVISER family is to provide each function, such as a tape reel drive or capstan drive, in an easily inserted module. The system of functional modules then becomes flexible enough to provide a wide variety of equipment configurations with maximum commonality of component parts. Fig. 1 illustrates the module breakout for a typical transport unit. Only one element, the base casting, is changed to provide various tape capacities.

The functional modularity is, of course, extended to cover electronic modules contained in the package shown in Fig. 2. In addition, a greater flexibility is embodied in effective subsystem modularity as shown by the electronics subsystem nest-modules in Fig. 3, allowing the designer/owner to form a wide variety of recording systems. A partial list of available subsystem modules is shown in Tables I through V.

Drawing from the module complement, a configuration can be made of single or dual wideband channel equipment for instrumentation, electro-optical sensors, digital data, or any combination of the two. In addition, SMPTE compatible modules are available for NTSC/PAL color recording. Depending on size restrictions and record time requirements, there are three alternatives for single-channel and two alternatives for a two-channel configuration.

Thus, the ADVISER family represents a true advance in state-of-the-art recording systems, resulting in maximum efficiency both in manufacture and use.

Future developments

Present design efforts for extended bandwidth recording to 15 MHz are expected to be available as options in the near future. A two-channel recorder with 15-MHz per channel capabilities is shown in Fig. 4. Such a recorder will have dramatic performance advances in a configuration for the recording of the "I", "Y", and "Q" color television channels. The two-channel recorder will have maximum capability when recording the wideband "Y" (or white) information in one channel and the "I" and "Q" (or blue and red) multiplexed into the second wideband channel.

The increased bandwidth will provide excellent fidelity for recording and

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reproducing high-resolution imagery such as 655-or 875-line systems. Companion portable systems for remote location are also being developed for the wide-band applications.

Ultimately, requirements for editing and signal enhancement will lead to the digital encoding of television signals for maximum flexibility in processing. The signal will be reduced to analog form for transmission purposes only, though in the future, digital transmission may also occur. Recognizing this trend, RCA has been addressing the recording of digital data for long-term storage at data rates in excess of 10^8 bits per second. Borrowing from the video head/tape interface technology, RCA has developed multitrack, unitized head structures with track densities in excess of 80 tracks per inch as shown in Fig. 5. The high track density yields excellent efficiency for the recording of digital data. Fig. 6 indicates the data rate *vs* record time possible at various in-track bit densities. A recording system comprised of 160 channels requires the simplest of record/reproduce circuits to keep the component count low. Record/reproduce circuitry has been reduced to 18 active elements per channel, resulting in maximum reliability for the high-density multitrack recorder (HDMR). Through the use of error detection and correction encoding, a bit error rate of less than one error in 10^6 bits has been obtained. A unique

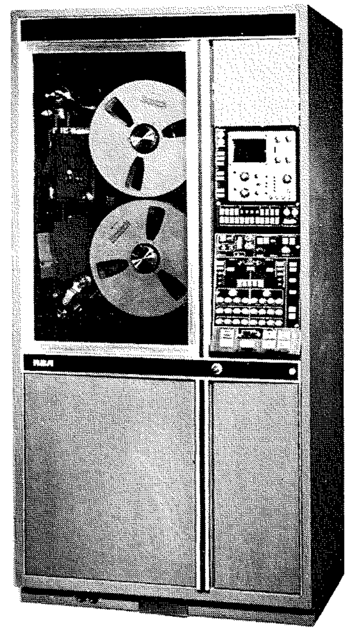

Fig. 4 — CVR 152, two-channel recorder.

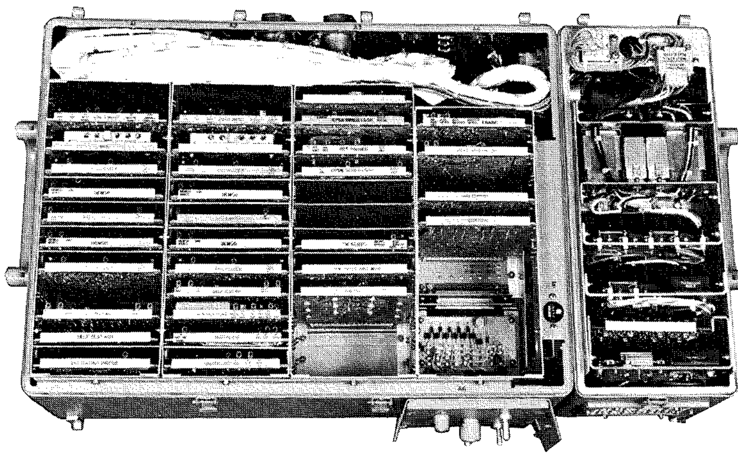

Fig. 2 — ADVISER electronics.

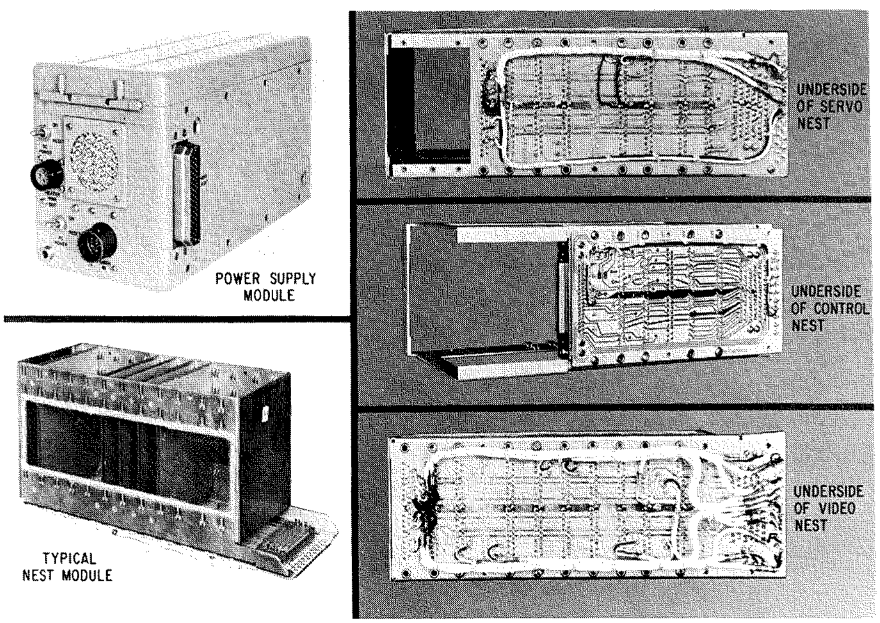

Fig. 3 — ADVISER electronics modularity.

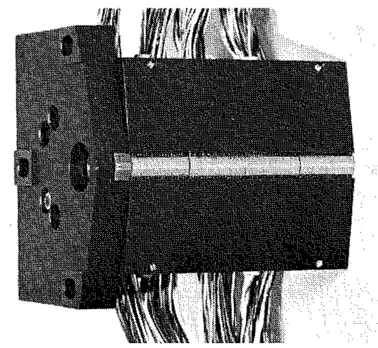

Fig. 5 — Multichannel head.

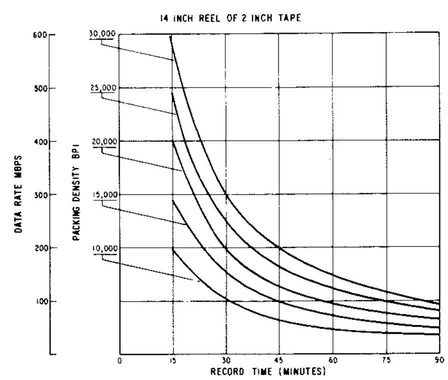

Fig. 6 — Data rate/record time tradeoffs.

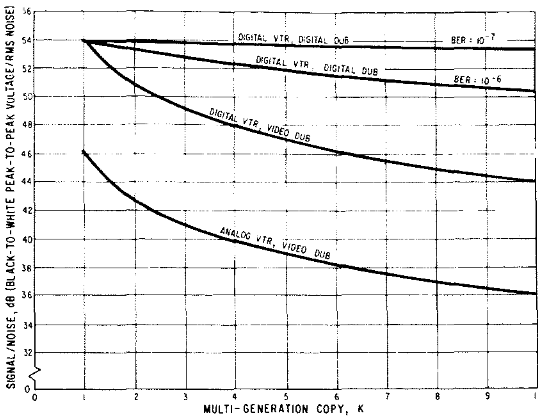

Fig. 7 — Multiple generation comparisons.

demultiplex-multiplex scheme which spreads drop-out errors across an entire field results in minimum image deterioration. Fig. 7 indicates the resultant signal degradation over multiple generation dubs to be expected for a digital system compared to present analog methods.

Conclusion

High-density digital magnetic recording has progressed to the point of successful laboratory operation of 1/4-, 1/2-, and 2-in. tape transports at bit densities as high as 10^6 bits per in^2. These laboratory tests are highly encouraging, and clearly point the way to the next steps in the application of HDMR to broadcast television recording.

First will come the task of setting the optimum digital format for color television. Since this format will have to be an international standard to permit economical interfacing among digital equipment of various manufacturers, the selection of the format must not be undertaken lightly nor pursued superficially. A careful selection here will pay dividends for decades.

Second must come the construction of hardware to achieve the ends permitted by the format. Now the laboratory accomplishments will be translated into producible equipment designed for customers to use in the field.

Both steps must be carried out with the care that is required when one is laying the foundation for a totally new phase of an industry. However, the careful execution of both steps will provide the television industry of the future with unparalleled performance from video recorders.

Table I — Wideband channel A and B module options.

Designation	*Description*
Wideband A (I) Wideband B (I)	Instrumentation recording, video/fm channels
Wideband A (D) Wideband B (D)	Digital recording, video/fm channels
Wideband A (EOS) Wideband B (EOS)	Television format recording, video/fm channels
SMPTE	NTSC/PAL highband color recording, video/fm (SMPTE) channel

Table II — Power supply module options.

Designation	*Description*
400-1	110-volt, 400-Hz, single phase
400-3	110-volt, 400-Hz, three phase "WYE"
50-60	110-volt, 50 to 60-Hz, single phase
50-60/2	220-volt, 50 to 60-Hz, single phase

Table III — Servo module options.

Designation	*Description*
Servo type 1	Asynchronous, rec./repr., power servo
Servo type 2	Asynchronous, rec./repr., hyst. synch. drive
Servo type 3	Synchronous, SMPTE, power servo
Servo type 4	Synchronous, line, hyst. synch. drive.

Table IV — Auxiliary channel module options.

Designation	*Description*
Aux type 1	15-kHz analog rec./repr.
Aux type 2	10-KBPS digital rec./repr.
Aux type 3	Multiplexed time code/voice rec./repr.

Table V — Transport module options.

Designation	*Description*
Transport type 1	Tape handling with 6-in. NAB reels
Transport type 2	Tape handling with 8-in. NAB reels
Transport type 3	Tape handling with 10-in. NAB reels

Hi-speed SOS COS/MOS random-access memories

A. Dingwall | W. R. Lile | J. H. Scott, Jr.

RCA has been a pioneer in COS/MOS technology, developing circuits for use in logic arrays and memories. The major virtues of COS/MOS devices are low power, high noise immunity (30%), and a wide range of operating voltages and temperatures (typically 3 to 15 volts and -55° C to +125° C). Memories, in particular, provide an excellent market potential for COS/MOS technology. With the recent achievements by using sapphire substrates instead of silicon, a new technology is added to the COS/MOS effort resulting in lower parasitic capacitances and higher speeds.

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Joseph H. Scott, Group Head, Integrated Circuit Technology and Applications Group RCA Laboratories Princeton, N.J. received the AB in Chemistry from Lincoln University, Pennsylvania, in 1957. In 1958, he attended the Graduate School of Chemistry at Howard University, Washington, D.C., followed by work in electrical engineering at Newark College of Engineering, while employed by RCA. In 1959, he joined RCA Electronic Components and Devices, Somerville, N.J., where he was engaged in research and development of semiconductor devices. This included work on gallium arsenide solar cells, silicon p-n junction devices, insulated gate MOS transistors, the NMOS memory device, and solid diffusion techniques. In 1967, he transferred to the David Sarnoff Research Center, Princeton, N.J., where he pursued work in integration of complementary devices (both MOS and bipolar) in bulk silicon, and in MOS on insulating substrates. For work in these areas, he was awarded three RCA Laboratories Achievement Awards. Mr. Scott has been issued 12 U.S. patents and has others pending. He is a member of Sigma Xi and the Electrochemical Society. Mr. Scott is also a Senior Member of the IEEE.

Dr. Andrew G. F. Dingwall, COS/MOS Memory Array Advanced Techniques, Solid State Technology Center, Solid State Division, Somerville, New Jersey, received the PhD in Glass and Ceramic Technology while a Fulbright exchange scholar at the University of Sheffield, England, and holds Masters Degrees in Mathematics as well as Electrical Engineering from the Polytechnic Institute of Brooklyn. He has participated both in the design and processing of LSI arrays at RCA. He is a member of the engineering team which received the 1971 David Sarnoff Outstanding Achievement Award in Science for use of LSI arrays. Dr. Dingwall is a member of Tau Beta Pi, the IEEE, the American Mathematical Society, the Institute of Physics, and the American Ceramic Society. He has 25 issued or pending patents in the semiconductor field and has written for numerous publications.

THERE ARE MANY advantages in using COS/MOS technology. COS/MOS memory arrays fabricated on either bulk silicon or on insulating substrates have outstanding features not achieved by either single-channel MOS or bipolar technology. One of the advantages of COS/MOS circuitry is its capability to operate over wide extremes of temperature and power-supply variations, yet maintaining a high noise immunity. For example, a typical operating range for memories is 3 to 15 volts. COS/MOS memory-array devices can be designed to be compatible with the popular bipolar T-L circuits without additional interfacing circuits.

Economies in COS/MOS system design are achieved because the system can operate on a single, noncritical, power-supply voltage. Furthermore, static memories, *i.e.*, those that require no refresh memory also require no high-power clock or preconditioning pulses for proper operation. System noise problems are diminished because large transient currents are not developed. System complexity can be substantially less than that required with other semiconductor technologies.

For most users, another outstanding advantage of COS/MOS memories is the low power dissipation, which is typically 0.4μW/bit at 10 volts for a 256-bit memory. Low power dissipation permits COS/MOS chips to be assembled in

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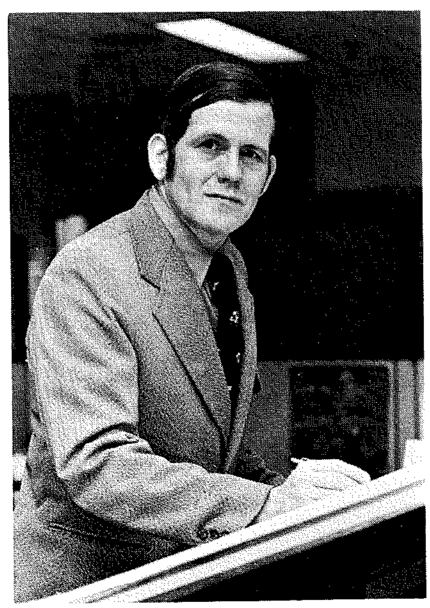
W. R. Lile, left, and J. H. Scott.



Dr. A. G. Dingwall

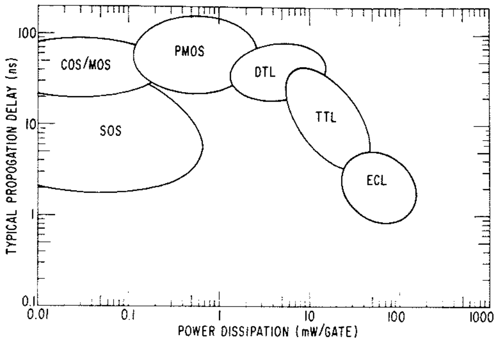
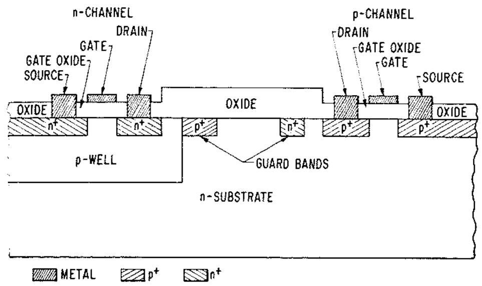


Fig. 1 — Comparison of propagation delay and power dissipation for digital devices using several technologies.



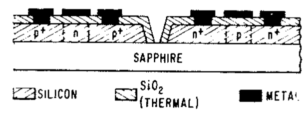


Fig. 2a — COS/MOS transistor on bulk silicon showing guard bands. Fig. 2b — COS/MOS transistor on sapphire.

compact arrays without heat sinking, and the low heat generation ensures extremely high levels of reliability and trouble-free operation. This latter feature makes their use ideal in remote locations. The low levels of power dissipation enable a simple, standby battery power supply to guard against memory loss in case of power failure in those applications where a nonvolatile memory is important.

Two technologies used in COS/MOS are concerned with bulk silicon and silicon-on-sapphire (SOS). The bulk process has been in use for several years and is well entrenched. The more recently introduced SOS process has already proven to be of great value because the substrate is an insulator, thereby eliminating parasitic capacitances that decrease speed and increase power consumption. Memory systems based on COS/MOS technology can be designed to operate with access times in the range 50 to 500 ns. Because SOS memory devices typically have 3 to 5 times the speed of corresponding bulk silicon devices, their use would be mandatory in the highest-speed systems.

Performance of different logic devices

Logic circuits have experienced upgrading within various technology groups such as transistor-transistor logic (TTL), emitter-coupled logic (ECL) and COS/MOS. Delay times and rise times have decreased; however, this increase in speed often results in an increase in power dissipation. The ECL circuits exhibit this type of trade-off.

A technological advance in the COS/MOS type of logic circuits is the use of silicon-on-sapphire. Fig. 1 illustrates the approximate power-stage delay boundaries for several circuit types. The circuit designer considering power, speed, and cost, must pick the optimum type and subgroup. SOS processing is not intended to replace the bulk process, but is an extension of the technology and must be regarded in such a light. The design trade-off will probably be between cost and speed in the initial production stages.

Bulk vs. SOS design trade-offs

The rapidly emerging SOS technology has provided the designer with features not available in bulk COS/MOS technology. For applications where maximum speed and lowest dynamic power or transient radiation burst environments are critical, the SOS technology now provides a clearly superior design choice. On the other hand, the mature, low-cost bulk COS/MOS technology offers significant potential for continuing improvement. Development projects now being pursued should result in significant increases of packing density and speed. Such projects include better design rules to increase compactness, ion implantation techniques to result in lower and more closely controlled thresholds, and the introduction of self-aligned silicon-gate structures. COS/MOS memories provide an example of an application where speed is frequently important and where the impact of current and projected designs in these two technologies can be assessed.

Profile comparison of process layers

Fig. 2a illustrates the process profile of a complementary-symmetry pair. Bulk processing involves a diffusion around the periphery of each separate transistor, diode, well, or tunnel to provide electrical isolation. Electrical isolation is necessary because the substrate is a conductor. Although leakage has been reduced to a minimum, parasitic capacitors introduced in the processing must be charged and discharged, thus requiring additional transient power, which results in decreased speed.

The SOS technology uses a sapphire substrate with an orientation of its crystal lattice to approximate that of pure silicon. Silicon is grown epitaxially on the sapphire, circuits are defined by photoresist techniques, and diffusions are completed. All unused material is stripped down to the insulating sapphire substrate. Fig. 2b illustrates the cross section of a COS/MOS inverter on sapphire. Leakage is reduced since the area between devices acts as an insulator rather than as a reverse-biased diode. With the elimination of the guard rings the total area available for circuits has been increased, thereby affording potentially higher packing densities. A major advantage of the SOS structure is the low parasitic capacitance losses that result from fabricating MOS devices on an insulating substrate. This process eliminates the capacitance caused by reverse-biased junctions. In bulk silicon devices the drain-source junctions are an important source of parasitic capacitance; smaller but still significant sources of parasitic loss in bulk devices are the areas where gate and interconnect lines cross over the substrate space between MOS devices. These sources of loss are absent in SOS structures in which the space between devices is insulator material. Table I summarizes typical drain substrate parasitic capacitance for aluminum and silicon-gate COS/MOS transistors.

Other parasitic capacitance losses in both bulk and SOS devices are associated with gate-to-substrate and gate-to-drain capacitance. These include the geometric gate-substrate capacitance and the non-negligible gate-drain capacitance whose effect is enhanced by the gain of the MOS transistor. As noted in Table I, this latter effect will be substantially reduced for

Table I — Typical sources of parasitic capacitance in SOS and bulk COS/MOS transistors.

Parasitic capacitance source	Capacitance (pF)			
	Aluminum Gate		*Silicon Gate*	
	SOS	*Bulk*	*SOS*	Bulk
Drain-substrate	—	0.10	—	0.07
Gate-substrate	0.03	0.06	0.03	0.03
Effective Miller gate-drain	0.08	0.12	0.02	0.02
Interconnect	0.01	0.02	0.01	0.03
Total capacitance	0.12	0.30	0.06	0.15
Relative speed	3	1	5	2

both SOS and bulk technologies as self-aligned silicon gate COS/MOS process and tighter layout rules are employed. Device speed in COS/MOS circuits is intimately related to parasitic capacitive loads resulting from device structure. The high degree of correlation between parasitic capacitance and relative speeds is shown in Table I.

The introduction of a self-aligned silicon-gate process will increase the speed performance of both SOS and bulk structures, although the same relative SOS advantage of approximately 3:1 will be maintained.

The memory cell

The development of MOS memories has resulted in several varieties of either static or dynamic cells. Static memory cells retain their information as long as power is applied to the device without the necessity of pulsing to "refresh" the data. This "refresh" term implies that some memories slowly lose information because of charge leakage, and this lost charge must be periodically replaced. A type of memory that requires a refresh function or cycle is known generally as a dynamic memory. Random-access memories (RAM's), shift registers, and binary counters can be designed to be either static or dynamic.

The choice of a static or dynamic memory for use in a system is dictated by many interrelated factors. The design engineer must, therefore, know the requirements of his system and the advantages and disadvantages of each type of memory. A type of dynamic cell is shown in Fig. 3a. This three-transistor cell stores the information (charge) at the node containing parasitic capacitor C.

Fig. 3b illustrates a basic six-transistor static cell. The memory cell itself comprises two COS/MOS inverters cross-coupled to form a flip-flop. Two additional devices are needed to provide access to the cell. This is the cell configuration chosen for the first SOS memory array to go into production.

Although it requires more surface area than the dynamic cell, the static cell is quite stable, requiring no refresh cycle. Further, it lends itself well to computer-aided design techniques. The static cell operates at a high confidence level as a nondestructive readout device.

The typical COS/MOS inverter is fast and has low power consumption. Since the p-channel and n-channel devices do not conduct simultaneously in the static state, the only power-supply current flowing during the static state is that required to supply the leakage current in the "off" devices. This is typically of the order of one nA per flip-flop. The data stored is not degraded by this leakage as is the data stored in the dynamic cell. The leakages can vary from cell to cell without resulting in system timing problems or device yield.

When the cell is changing state, both inverters conduct for a few nanoseconds. This switching current is, therefore, duty-cycle sensitive. In the language of the design engineer, a memory said to have a dissipation of so many microwatts per bit is usually a static memory. Dynamic power consumption in a static memory is sometimes specified for a given cycle time. Power consumption for COS/MOS arrays is several orders of magnitude less than that for bipolar devices such as TTL or ECL. Bulk process COS/MOS circuits are not as fast as the SOS process COS/MOS circuits. SOS memories will compete directly for the TTL device market in the realm of speed and maintain a decided advantage in terms of power requirements. These comparisons are discussed later.

The memory array

Since SOS circuits are not new types of circuits, but rather improvements stemming from technology of COS/MOS devices, it is reasonable that during the early new-product stages, SOS products should be an enhancement of an existing bulk-process device. Typical of this philosophy is the TA6140, which is a SOS version of the CD4024, a 7-stage binary counter. The TA6366 is an example of a 256-bit SOS RAM that can be used to achieve improved performance over the bulk TA5974. These SOS devices are pin-for-pin compatible with their bulk-process counterparts.

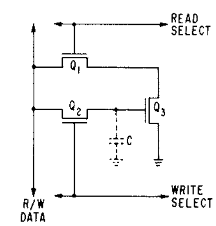


Fig. 3a — Dynamic memory cell.

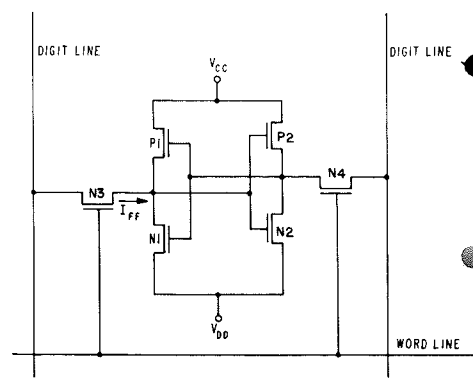


Fig. 3b — Static memory cell.

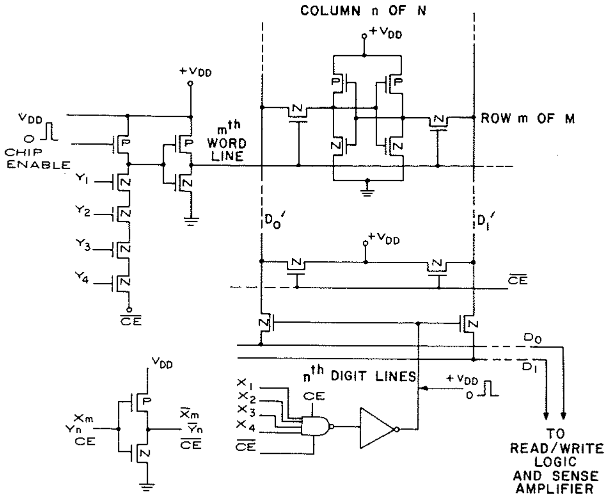


Fig. 4 — Single cell in an M x N matrix.

The memories described here have a sense amplifier integrated on the memory chip. This amplifier has three output states: logical "1", logical "0", and high impedance. It is this last property that enables many memories to have their outputs OR-tied together. Only a capacitive load will be presented to the selected memory output. Since the COS/MOS circuits, including memory arrays, can operate over a voltage range of 3 to 15 volts, the memory should be capable of interfacing directly with TTL logic circuits as well as with COS/MOS circuits. To provide for proper TTL drive capabilities in the sense-amplifier output, this interfacing advantage must be recognized in the design stages.

In the TA6366 (the SOS version of the 256-bit RAM) all decoding, read-write logic, and sense amplifier circuitry is COS/MOS. This provides consistency in the design with low power dissipation. The static six-transistor memory cell illustrated in Fig. 3b is repeated and interconnected to form a memory matrix of M x N words, one bit long. This matrix, with a single representative cell and associated decoding, is shown in Fig. 4.

The operation of this word-organized SOS memory can be understood by referring to Fig. 3b. The bulk unit differs from the SOS unit primarily in the design of the decoder. Inverters P_1-N_1 and P_2-N_2 are cross-coupled to make the flip-flop storage cell. The cell is accessed via N_3 and N_4. This is the same cell as shown in Fig. 3b.

The selection of a cell (word) is determined by selecting one of the M-word lines (this selects a particular row) and one pair of the N sets of digit lines (this selects the column). The m^{th} row is selected when a particular address decodes so as to make the m^{th} word line a "high" voltage (V_{DD}) or a logical "1".

This has the effect of turning on N_3 and N_4 for every cell in the row. However, since only one column-pair of the N columns is selected, the selection of the cell is unique.

Data is entered into the cell in the digit lines D'0, D'1 and read from the cell on these same lines. Transistors Q_7-Q_8 are also repeated in every column. Their function is to keep the digit lines charged to V_{DD} when the memory is not being accessed.

For a memory of M x N bits where M x N is a power of two, the required number of bits in an *address* register may be expressed by the formula

$$2 \times \log_2 \quad (M \times N)^{1/2}$$

In order to decode any address in binary, both the function (in this case an *address* bit) and its complement must be made available to the memory decoder. It is more practical to generate these complements internally than to require large packages with sufficient pin requirements.

To save socket pins and wiring complexity, only the *address* lines are brought into the chip. For a 256-bit memory, eight address lines are needed. In Fig. 4, each input to decoder, Y_1 to Y_4 or X_1 to X_4, is a "1" or "0" giving 16 combinations for X or Y. A particular decoder is activated if, and only if, all inputs are "1". *Chip Enable* (CE) is considered a fifth input to each decoder and this pulse controls the interrogation of the memory for *reading* or *writing*.

The sense amplifier for the SOS memory is a completely new design meeting the requirements of a tri-state output, all COS/MOS circuitry, and high speed. This amplifier is shown in Fig. 5.

System timing

A memory system consisting of many packaged memory chips needs these basic timing pulses and address information:
— *address* of the word
— *chip enable* (or select)

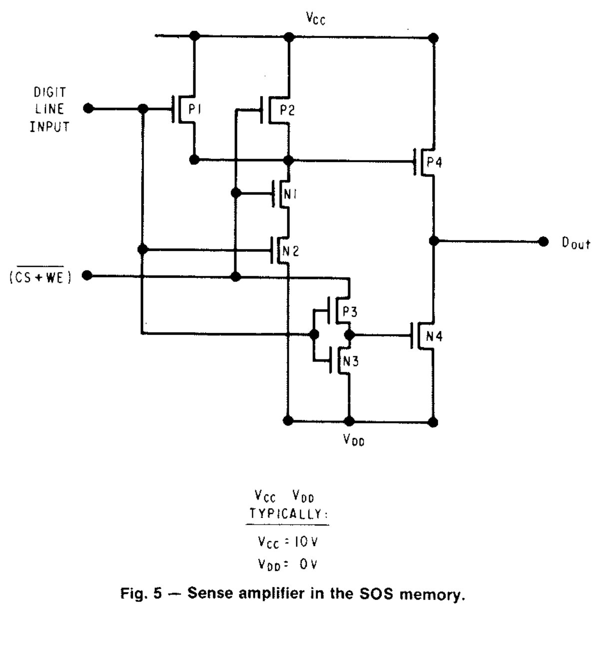
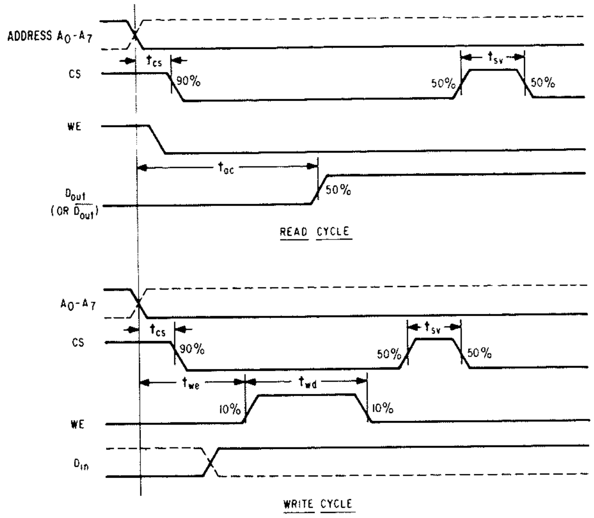


Fig. 5 — Sense amplifier in the SOS memory.



	SOS TA6366	BULK TA5974
t_{oc}	70 NS	400
t_{we}	30 NS	50
t_{wd}	40 NS	150
t_{sv}	30 NS	200

CONDITIONS: TEMPERATURE = 55°C
CAPACITIVE LOADING TO OUTPUTS = 20 pF
P-TYPE THRESHOLD VOLTAGE = 1.5 V
N-TYPE THRESHOLD VOLTAGE = 2.0 V

Fig. 6 — Memory system timing.

— *write enable*
—*data input*

The function of the *address* lines is to decode one cell in the M x N matrix of Fig. 4.

The *chip enable* pulse selects the chip to be interrogated when a multitude of chips have their outputs OR tied together More basic though, is that the CE pulse establishes the time during which a memory is to be *read* from or *written* into

The *write enable* pulse is fairly self explanatory in that when it appears, the selected cell will store that data appearing on the *data-in* line. The absence of this level will automatically imply a *read* condition.

The *data-in* line is used only during the *write* cycle and is inactivated during the *read* cycle. Before the appearance of the *write enable* pulse, the *data-in* line must be stabilized with the information that is to be written into the memory. The data must remain long enough — as must the *write enable* pulse — to guarantee storage of this new information.

Table II — Comparison between bulk and SOS memories.

	Bulk (TA5974)	SOS (TA6366)
Access time	400 ns	70 ns
Write setup time	50	30
Write enable width	150	40
Chip select setup time	200	30

V_{CC} = 10 V for both memories.

Not listed above, but appearing on the timing diagram in Fig. 6, is the *data-out* line. The output delay time is dependent upon the design of the system including capacitive loading.

The relative speeds of a general class of SOS and bulk silicon COS/MOS circuits were shown in Table I. Table II shows a comparison between the two memories discussed here.

The times that engineers usually quote are access time and cycle time; for the bulk memory, cycle time is about 600 ns, and for the SOS memory, about 130 ns. These numbers further validate the approximate speed ratio of 3:1 between the two technologies as illustrated in Table I. As usual, if the designer wants to upgrade his system, he must be willing to pay a higher price as SOS technology cannot as yet produce COS/MOS circuits as cheaply as the older, more-established bulk silicon process.

A timing diagram with those pulses discussed previously as necessary for memory system operation is shown in Fig. 6.

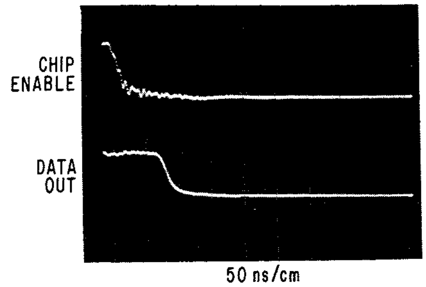


Fig. 7 — SOS memory waveforms showing access time of 70 ns.

Actual waveforms for the SOS memory showing *chip enable* and *data out* are illustrated in Fig. 7. The access time is 70 ns, which was calculated to be the nominal value. Memory output transition time is observed to be about 20 ns with a 20-pF load.

An overall view of the SOS chip is shown in Fig. 8. The scribe dimensions of the chip are 126x132 mils. The sapphire substrate is 10 mils thick.

Acknowledgments

The authors wish to thank R.J.Hollingsworth for his contributions in the design of the SOS memory and J. C. Sarace for SOS process innovations.

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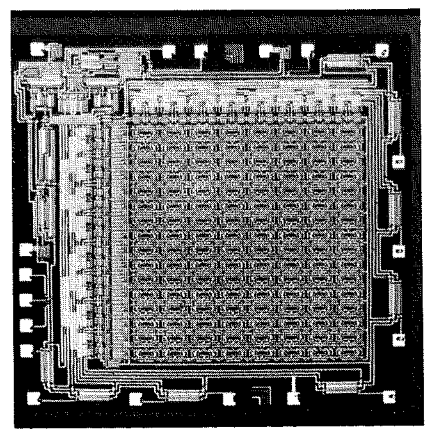

Fig. 8 — Photograph of TA-6366 SOS 256-bit memory.

The RCA VHF ranging system for Apollo

E.J. Nossen

Redundancy of functions on manned space flights has been an important concept for crew safety. However, a redundant system generally implies doubled weight — a luxury that cannot easily be afforded on a spacecraft. Thus, to achieve a backup for the ranging function for the Apollo Command Module-Lunar Module rendezvous mission, RCA developed the VHF Ranging System which permitted the voice radios to be adapted for this important function. This technique kept the weight increase to less than one-eighth of what would have been required for a redundant rendezvous radar system.

As the Apollo program proceeded, NASA became increasingly concerned for the safety of its crews on manned space flights. Redundancy became a requirement for all crew safety functions. One critical period of the Apollo missions was the rendezvous of the Command Module and the Lunar Module. The RCA-developed rendezvous radar provided the critical range, range rate, and angle measurements necessary to complete the rendezvous. Use of a redundant radar for backup was out of the question because of its 80-lb weight. Angle measurements could be obtained from the navigation sextant, range rate could be obtained by differentiation of range data in the spacecraft computer. However, redundant range data was not available.

After investigating the requirements and possible solutions to the problem, RCA proposed to Frank Borman, the commander of the first Apollo flight to the moon, that the voice radios be adapted to perform the ranging function. Slight modifications of the RCA-built VHF voice radios and the addition of a ranging interrogator and transponder at a weight of less than 10 lb total would provide an accuracy of 100-ft rms at several hundred miles.

The Apollo VHF ranging system development was authorized in the Fall of 1967. Slightly more than a year later, the system was successfully flight-tested at the White Sands Proving Grounds, and the first space-qualified flight hardware was delivered. Since it has performed flawlessly on every Apollo Lunar rendezvous mission, it was the sole rendezvous ranging sensor on the Skylab mission and will be on the upcoming Apollo-Soyuz earth orbital mission.

VHF duplex system

The basic VHF ranging system, as illustrated in Fig. 1, uses a full duplex communications system. The Command Module (CM) VHF transmitter is modulated by a voice signal or by a ranging signal, or both functions can be carried out simultaneously. The signal is transmitted via a diplexer and antenna for reception by the Lunar Module (LM) VHF receiver. The voice information signal is obtained by conventional envelope detection; the ranging signal is demodulated and applied to the transmitter. In some modes it is fed

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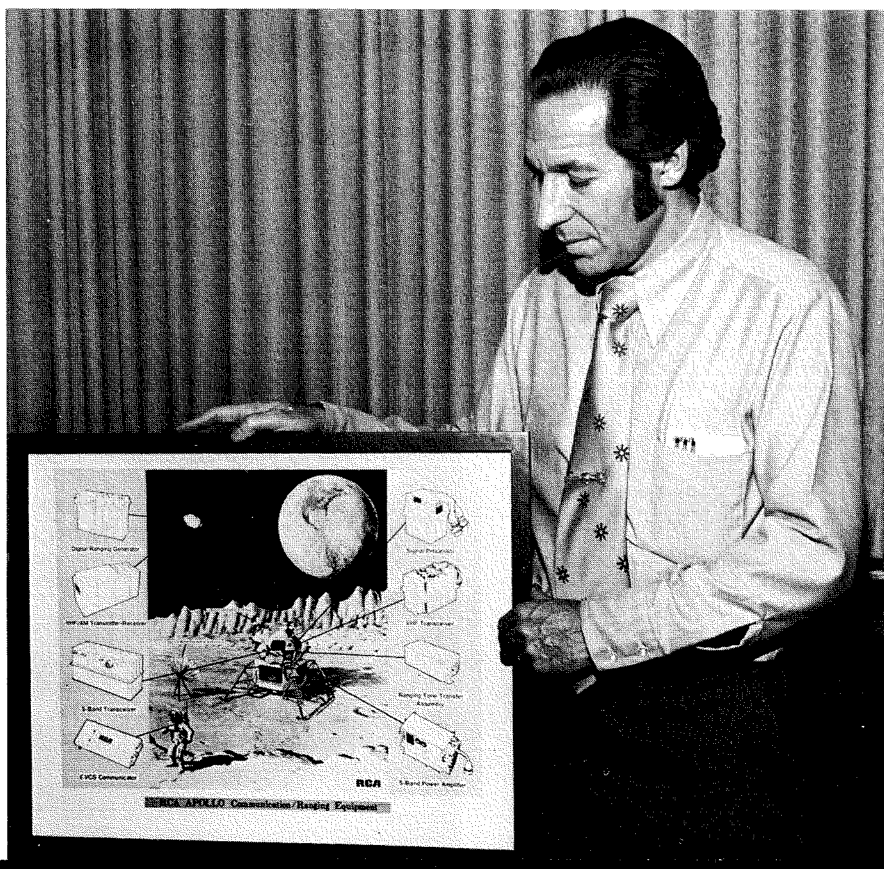


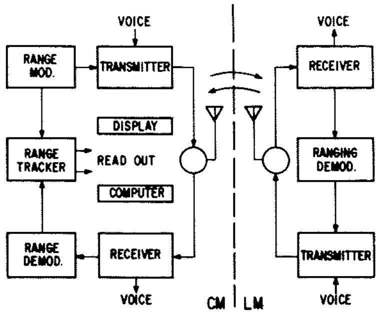


Fig. 1 — VHF ranging system.

directly from the envelope detector to the transmitter without synchronization. The LM transmitter and antenna radiate a voice and/or ranging signal which is picked up by the CM receiver. The voice and ranging modulation are fed to separate circuits. The range information is demodulated and causes a range tracker to follow the transmission path delay. Comparison of the time position of the received ranging waveform with respect to the transmitted ranging waveform at the CM, yields the range readout.

Transmitter configuration

The VHF transmitters in the CM and LM use speech clipping for the conversion of the analog voice signal to a bi-level waveform. The bi-level voice signal amplitude modulates the rf carrier in a binary fashion (*on-off*) by means of a keyer. The modulated carrier is further amplified and filtered before transmission. In the receiver, the bi-level waveform is filtered and a very intelligible voice signal is recovered.

The CM and LM VHF transmitter configuration is shown in Fig. 2. The rf carrier is derived from a crystal-controlled oscillator, which drives a multiplier and an amplifier chain. The voice signal is processed by successive clipping and appears as a bi-level waveform at the input of the keyer. In some units a data input also drives the keyer. For minimum impact on the communications system, the most suitable ranging waveforms are

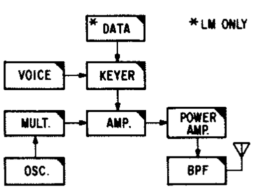


Fig. 2 — VHF transmitter.

square waves or a combination of square waves. Thus the ranging signal, the bi-level voice signal, or the combination signal drives the keyer and causes *on-off* modulation of the rf carrier to take place in the amplifier chain. The transmitter is capable of handling signals with a bandwidth of several MHz so that it does not severly limit the range measurment accuracy capability.

Receiver configuration

The VHF receivers in the CM and the LM, shown in Fig. 3, are fixed-tuned receivers designed to operate over a wide dynamic range of signal levels. A received signal at 259.7 MHz or 296.8 MHz is applied to a broadband gain-controlled rf amplifier and then translated to a 30-MHz i.f. signal. A crystal-controlled oscillator, frequency multiplier, and mixer perform the heterodyning function. The i.f. channel consists of relatively broadband i.f. amplifiers and a narrowband crystal filter. The filter's transmission bandwidth of approximately 60 kHz determines the receiver's selectivity. The i.f. amplifier preceding the envelope detector is also gain controlled to maintain a relatively uniform output level. The filter characteristics are shown in Fig. 4.

Although the i.f. filter bandwidth is about 60 kHz, the frequency stability of the transmitter and receiver oscillators may result in nearly ± 15 kHz of drift of the carrier frequency. Due to the steep skirt selectivity of the crystal filter near its band edge, it is not recommended to pass signals with frequency components in excess of 15 kHz through the receiver. Ranging signals much below 15 kHz may be passed through the i.f. amplifier. However, certain factors must be considered, such as the delay through the receiver and the variation of this delay with temperature, signal level, and from one unit to the next. The fixed delay for a typical receiver through the detector output is approximately 21 μs.

Most of this delay is attributable to the crystal filter and the i.f. amplifier. The delay varies about ± 0.6 μs due to temperature changes and about ± 0.9 μs due to signal level variation. The variation between different sets can be as much as ± 3 μs, in addition. Frequency offsets between the transmitter and the receiver local oscillators will also add several microseconds to the total delay uncertainty.

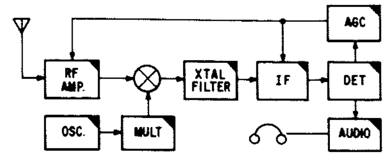


Fig. 3 — VHF receiver.

Ranging considerations

The variable delay determines the limit of measurement accuracy achievable with a given system regardless of the ranging waveform or signal-to-noise ratio. The variable delay is usually some fraction of the fixed delay and can be therefore minimized by also reducing the fixed delay. This requires the use of the widest bandwidth circuits available relative to the frequency spectrum of the ranging waveform. Better performance can thus be expected from a ranging signal which does not have to pass through the band-limited i.f. amplifiers and filter. It must therefore be correlated before the filter, so that only an error signal is passed. Since the error signal is usually heavily filtered, a narrowband i.f. channel is adequate to pass it. The ranging signal may have frequency components of the order of 100 kHz or more, because the transmitter and the rf amplifier in the receiver can handle several megahertz.

A combination ranging approach lends itself to the reception of two or more ranging tones, where only the highest fundamental frequency can not be passed by the receiver i.f. The highest frequency tone is demodulated to preserve system accuracy; the lower frequency tones needed for ambiguity elimination are not demodulated in the transponder. This is acceptable, even from an accuracy point of view, because the range measurement accuracy is not influenced by tolerable errors in the ambiguity-resolving waveforms.

Ranging implementation

To obtain measurements by means of the VHF radio equipment, a three-tone ranging technique is used. To be compatible with the *on-off* modulation of present transmitters, a similar modulation is used for ranging purposes. To avoid reducing the transmitter duty cycle, and thereby reducing transmitted power, a time sequential transmission of tones is used. Fine range is measured with a 31.6-kHz square-wave tone. Range ambiguity is resolved with a mid-frequency of 3.95-

kHz and a low frequency tone, which is a modulo 2 combination of a 3.95-kHz and a 247-Hz square wave. This combination has the advantage of a maximum unambiguous range of about 327 nm while the signal is narrowband and centered about 3.95 kHz. Since normal tracking provides range measurements, the mid-and coarse-tone signals are only transmitted when range tracking is initiated, when an interruption of tracking has occurred, or when the range data is to be checked. A manually initiated operation provides for automatic acquisition and tracking of the mid-and coarse-range signals for an 8-second period. Thereafter, automatic switching to the fine-ranging signal occurs. At the CM, both the transmitter and receiver tracker are sequenced through the appropriate mode. At the LM transponder, the presence of narrowband modulations, either mid-or coarse-range tones, is sensed and the mode of operation is automatically changed. The ranging tones are shown in Fig. 5.

Transponder operation

The primary goal of the ranging system development was to minimize changes to existing equipment. The VHF set and its interfaces with the ranging transponder unit are shown in Fig. 6. In the coarse-ranging mode, the VHF receiver operates in its normal fashion. A composite ranging tone centered about 3.95 kHz is received, clipped to produce a bi-level signal, and then applied to the transmitter to key the modulator as is otherwise done with voice signals. A coarse-tone signal sensor inhibits the fine-tone tracker from degrading the signal and selects the appropriate signal for application to the transmitter input.

In the fine-ranging mode, the received signal is *on-off* gated by interrupting the signal path preceding the crystal filter at a 31.6-kHz rate. The phase of the incoming square wave is correlated with the signal generated by the fine-tone tracker. By accurately tracking the received signal with a locally generated waveform, the latter may be used to key the transmitter with a noiseless signal. Smoothing in the tracking loop reduces the phase jitter to a relatively small value.

In the fine-ranging mode the received signal is the 31.6-kHz square wave, which is gated before it reaches the narrowband i.f. filter. The gating waveform is derived by counting down from 2.022 MHz which is generated by a voltage-controlled crystal oscillator (VCXO). The countdown chain also produces a square wave at 5.27 kHz which is used to shift the phase of the 31.6 kHz signal by $\pm 2\ \mu$s. By shifting the phase of the reference signal, an early and a late version of the waveform are produced suitable to provide a tracking-error signal. The early/late switching of the reference signal is essential to the tracking operation, and it is performed at a rate slow enough to be passed by the i.f. amplifier.

After correlation of the received signal with the reference signal, i.f. filtering, and detection, the remaining fine tone is attenuated and only the carrier components remain. If a tracking error exists, a switching-frequency component at 5.27 kHz will also be present. This is filtered, and then an "early" minus "late" signal substraction is accomplished by means of a synchronous detector. The latter lets the detected and correlated ranging signal pass directly into a low-pass filter during the early portion of the switching cycle; the ranging signal is inverted. The filter thus performs the subtraction and yields the average value of the early minus late ranging signal. The presence of an error voltage causes the VCXO to change its frequency in an attempt to reduce the error. The VCXO to change its frequency in an attempt to reduce the error. The VCXO drives the waveform generator and therefore controls the phase of the ranging waveform. For good performance, the loop filter and the VCXO form a second-order phase-lock loop of less than 30-Hz bandwidth. This assures an adequate signal-to-noise ratio and tracking accuracy with negligible dynamic error.

In the LM transponder, the late response waveform for receiver gating is also applied to the transmitter. Since it is used directly to key the transmitter, the total transponder delay consists of the 2-μs late delay, the transmitter delay, and the delay encountered up to the receiver's mixer. Since the ranging code has been demodulated before the i.f., the i.f. delay does not influence the time position or static range accuracy.

Range tracker operation

The CM VHF radio equipment implemented for the ranging function is illustrated in Fig. 7 . The appropriate ranging waveform, which is generated by

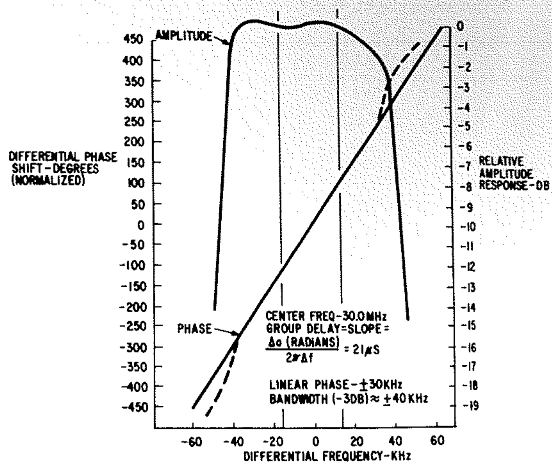


Fig. 4 — VHF receiver crystal filter characteristic.

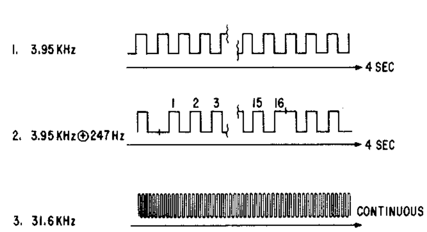


Fig. 5 — Ranging tones generated.

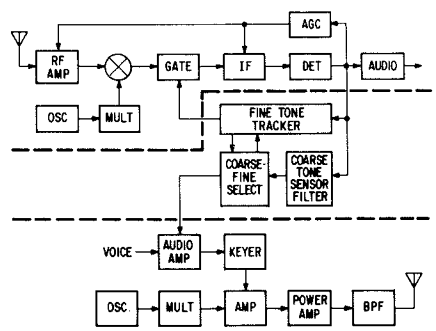


Fig. 6 — LM transponder (ranging function).

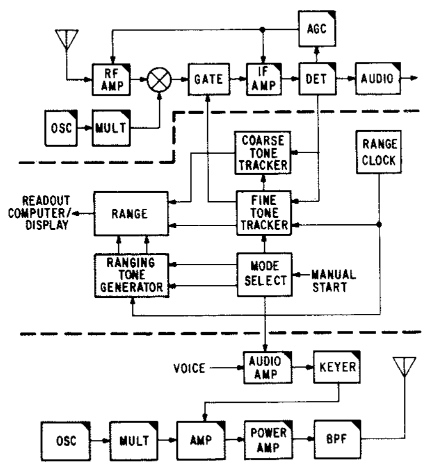
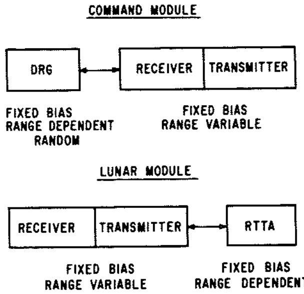


Fig. 7 — CM VHF radio (ranging function).



FIXED	– FIXED FOR EACH SET OF EQUIPMENTS-TRANSMITTER & RECEIVER COMBINED
RANGE VARIABLE	– VARIES WITH SIGNAL LEVEL IN AN UNPREDICTABLE FASHION ALTHOUGH KNOWN FOR EACH TRANSCEIVER
RANGE DEPENDENT	– VARIES WITH SIGNAL LEVEL SMALLER ERRORS AT HIGHER SIGNAL LEVEL
NOISE	– VERY SMALL IN VHF RANGING
GRANULARITY	– DUE TO DIGITAL TRACKER AND DISPLAY
VOICE	– NOISE LEVEL INCREASES AND LOOP BANDWIDTH CHANGES

Table I - Range error types.

	200 N.M. RANGING	200 N.M. R/VOICE	10 N.M. RANGING	10 N.M. R/VOICE
RANGE VARIABLE BIAS ERRORS	283	283	283	283
RANGE DEPENDENT BIAS ERRORS	196	330	30	30
FIXED BIAS ERRORS	562	562	562	562
TOTAL BIAS ERRORS	659	710	631	631
RANDOM ERRORS	68	151	61	209
TOTAL ERROR	663	726	634	665

Table II - Accuracy model (3-sigma) in nanoseconds.

SYSTEM	APOLLO 10	APOLLO 11	APOLLO 12	APOLLO 13
DELAY, µs	3.008	2.706	3.047	2.675
CALIBRATION, µs	2.835	2.835	2.835	2.835
ERROR, µs	0.173	-0.129	0.212	-0.160
ERROR, FEET	86	-64	106	-80

Table III - Actual Apollo range errors (—86dB).

	DIGITAL RANGING GENERATOR (DRG)	RANGING TONE TRANSFER ASSEMBLY (RTTA)
• WEIGHT	6.2 LBS.	2.9 LBS.
• SIZE	8-1/2 X 4 X 6 INCHES	8 X 4 X 3-3/4 INCHES
• POWER	19.7 WATTS	4.3 WATTS

- USE OF EXISTING VHF EQUIPMENTS (259.7 MHz AND 296.8 MHz) WITH APPLIQUE BOXES (DRG & RTTA)
- THREE FULL DUPLEX SYSTEM OPERATING MODES
  - a) RANGING OR
  - b) VOICE OR
  - c) VOICE / RANGING COMBINED
- THREE TONE SYSTEM FOR ACCURACY AND UNAMBIGUOUS RANGE (247 Hz, 3.95 KHz AND 31.6 KHz)
- SQUARE WAVE TONES - COMPATIBLE WITH APOLLO TRANSMITTER MODULATION
- FULLY QUALIFIED FOR SPACECRAFT ENVIRONMENT
- UNAMBIGUOUS RANGE READOUT TO 327.68 N.M.
- RANGE ACCURACY (3σ) ± 350 FEET TO 200 N.M.
- DISPLAY READOUT RESOLUTION - 0.01 N.M.
- COMPUTER DATA RESOLUTION - 0.01 N.M.
- ACQUISITION TIME 12-14 SECONDS (THREE TONES)
- MINOR CHANGES IN SPACECRAFT WIRING
- FLIGHT HARDWARE DELIVERED IN 14 MONTHS

Table IV - Apollo VHF ranging system characteristics.

means of the range clock and the ranging tone generator, is selected to provide either coarse or fine ranging. It is then applied to the transmitter where the keyer *on-off* modulates the rf carrier. The rf signal modulated with either the 31.6-kHz or the 3.95-kHz ranging tone, is then transmitted to the LM receiver.

The LM VHF equipment acts as a transponder and replies with the same signal it has received. The reply signal is received by the CM receiver, which is also modified to allow gating ahead of the i.f. filter to generate a tracking error signal when the fine-tone ranging mode is used. The range tracker may be shown functionally as a coarse-tone tracker and a fine-tone tracker. The range clock drives the fine-tone tracker, which in turn drives the coarse-tone tracker. Both trackers generate a waveform which is correlated with the received signal, before the i.f. filter for the fine tone, and after the detector for the coarse tone. The selection of the coarse- or fine-tone tracker is made concurrently with the selection of the transmitted ranging waveform. In the coarse ranging mode the receiver gate will pass the signal without interruption so that the maximum available signal-to-noise ratio will be realized.

A subtraction of the system delays is also performed before the data is transferred in serial form to the 5-decimal digit display and the spacecraft computer. The output data is available and displayed with a resolution of 0.01 nmi up to a maximum range of 327.67 nm.

Voice/ranging transmission

The Apollo VHF radio transmitter can be operated in either of two modes. In the non-ranging mode, the input audio signal is amplified, added to the 30-kHz sawtooth waveform from a noise suppression oscillator (NSO) and clipped, to produce a pulse-width-modulated signal. Strong audio signals will override the sawtooth waveform and will result in a clipped audio signal. In the absence of an audio signal, a 30-kHz square wave is transmitted, so that the amplitude modulated transmitter is always operating at a 50% duty cycle.

In the ranging mode, the NSO is disabled, and the 31.6-kHz ranging square wave is substituted. During acquisition by the ranging system, the lower frequency tones will, of course, be transmitted.

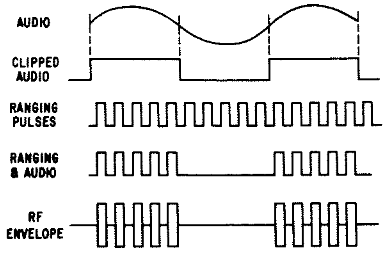


Fig. 8 — Ranging and audio waveforms.

Without the presence of voice or other audio signals, the transmission duty cycle is still 50%. Audio signals applied to the transmitter follow the identical path, but due to the absence of the NSO sawtooth waveform they are always clipped . In the ranging mode, voice signals therefore appear as clipped speech. This is combined in a logic AND function with the 31.6-kHz ranging square wave, so that the duty cycle of the transmitted signal can drop to as little as 25%. The rf envelope derived in Fig. 8 consists of the clipped audio waveform, which is further amplitude modulated (in *on-off* fashion) by the ranging square wave.

In the receiver, the high-frequency components due to either the NSO or the ranging square wave are filtered out by the audio amplifier. The audio signal is thus recovered.

Apollo VHF ranging accuracy

The range errors fall into a number of categories as shown in Table I; the major types are bias errors and random errors. The total range error includes the bias errors in the radio receiver and transmitter due to delay variations. In the Apollo system, the CM transceivers had 424 ns peak range error, while the LM transceivers had 395 ns peak range error. The ranging tone transfer assembly (RTTA) and the digital ranging generator (DRG) bias errors are due to offsets in product detectors, filter amplifiers, fixed and voltage controlled oscillators, and delay variations in interface circuits.

The combination of all range errors amounted to a 3 sigma value of about 660 ns or 330 ft as shown in Table II. These are three sigma errors allowing for all units under any of the specified spacecraft environmental conditions and for ranging or ranging/voice modes. The rms range error is therefore about 100 ft at maximum range. Actual measurements on four Apollo systems are shown in Table III, which indicate good agreement between predicted and actual range errors.

Apollo ranging system characteristics

The Apollo ranging system characteristics are summarized in Table IV. For details of how range tracking occurs through use of receiver gating on the Apollo VHF ranging system, refer to Appendix A.

Appendix A — range tracking through receiver gating

In the Apollo VHF Ranging System, the range measurement is accomplished by accurately tracking a 31.6-kHz square wave. Because this square wave is not passed by the receiver i.f. filter and because of the large delay variation in the i.f. filter-amplifier, a scheme of receiver gating is used. Since this gating or correlation takes place before the bandwidth limiting components, the system measurement accuracy is greatly enhanced. The principle of the gating operation is explained below.

Consider a dual i.f. tracking receiver as shown in Fig. A-1. It has a broadband rf section and mixer which do not impair the high-frequency components of the ranging signal. After the mixer, the receiver is split into an "early" channel and a "late" channel, where each consists of an rf gate, an i.f. filter-amplifier, and a detector. The gate allows the incoming signal to pass only for half the time under control of a reference square wave. The exact time interval is a function of the "early" and "late" reference signal phasing, which for illustration purposes will be taken as 1/8 cycle advanced for the "early" signal and 1/8 cycle retarded for the "late" signal. These reference square waves are derived by digital countdown logic which is driven by a voltage-controlled oscillator (VCO).

Fig. A-2 shows the input signal which will be assumed to agree in phase with the local reference in the tracking system. The relative time positions of the "early" and the "late" reference signals into the respective rf gates are also shown. The gate outputs shown at the bottom indicate that equal amounts of signal energy will reach the i.f. filter-amplifiers and detectors. Subtraction of the late output signal from the early output signal and filtering therefore produces no error signal to drive the VCO from its current phase position.

In Fig. A-3, the input rf signal is assumed to be delayed by 1/8 cycle with respect to the tracking system. The "early" gate disagrees in time position by a 1/4 cycle so that the gated output signal is only half the width of the incoming signal. However, the "late" gate agrees in time position so that the entire rf signal is passed by the gate. There is now an obvious difference between the detector outputs of the early and late channels, which will cause the VCO to change phase in an effort to minimize the error signal.

The error discriminator curve can be derived simply as shown in Fig. A-4. The baseband waveform is the square wave shown at the top. When it is multiplied by a replica of itself at all phase delays, the triangular autocorrelation function results. It reaches a maximum when the square wave and the reference are aligned; it is zero when they are out of phase. Autocorrelation functions for an "early" and "late" signal can also be drawn. They are also triangular but displaced in phase. A point on these correlation functions will exist for the "early" and "late" receiver channels for a particular phase of the incoming square-wave-envelope rf signal. The subtraction of receiver outputs may be represented by subtraction of the respective autocorrelation functions. This produces the time discriminator curve shown at the bottom of Fig. A-4. One of the zero crossings is the null around which the VCO tracks the signal phase.

The use of a dual i.f. channel receiver has several disadvantages. First, it requires a certain amount of equipment duplication which is seldom available in existing voice or data radio transceivers. Second, it is difficult and expensive to build two channels of identical bandwidth and gain. For these practical reasons, it is therefore advisable to time share a single i.f. filter-amplifier channel as shown in Fig. A-5. The identical reference signals are produced, but they are applied to the gate in sequence. The input to the differential amplifiers, which performs the "early/late" subtraction, is switched in synchronism. After filtering, the appropriate time error signal is obtained the drive the VCO and synchronize with the incoming waveform. The VCO output and its subharmonic frequencies may then be used for retransmission or for comparison with a transmitted signal to extract range measurements.

In the Apollo VHF Ranging System the "early/late" switching rate is 5.27 kHz so that three "early" gating pulses are followed by three "late" gating pulses. This frequency passes through the i.f. filter-amplifier without difficulty. The actual early and late displacements are 1/16 cycle to maximize the voice signal which is also obtained at the output of the envelope detector. The actual waveforms are illustrated in Fig. A-6.

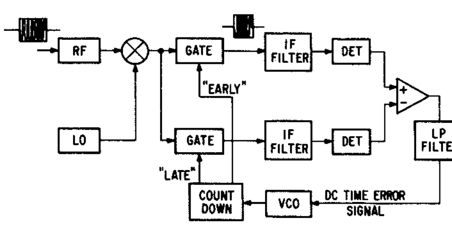


Fig. A1 — Dual i.f. tracking receiver.

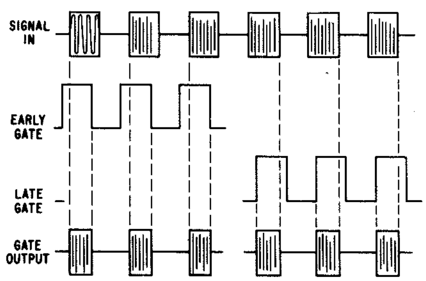


Fig. A2 — Tracking waveforms, input signal locked.

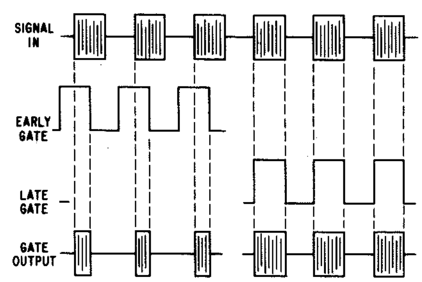


Fig. A3 — Tracking waveforms, input signal delayed 1/8 cycle.

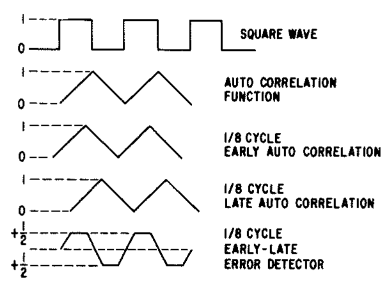


Fig. A4 — Early/late error detection diagram.

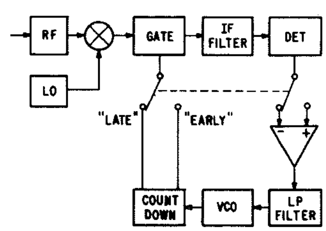


Fig. A5 — Single i.f. tracking receiver.

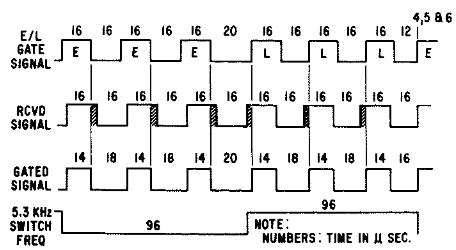


Fig. A6 — Early/late gating diagram.

Minimum-shift-keying modem for digitized-voice communications

E.J. Sass | J.R. Hannum

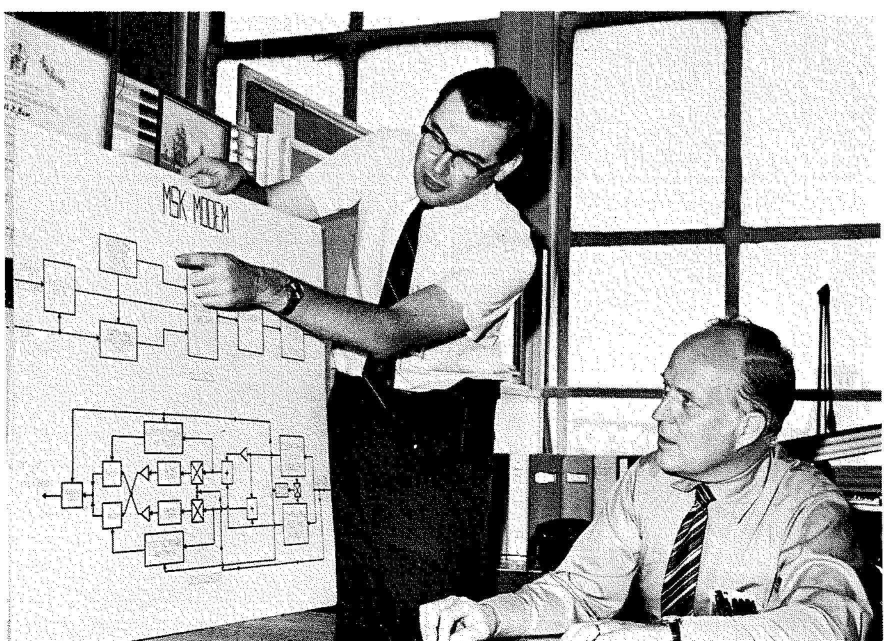
Since digitized-voice communications can enhance the security of communications among many units of a modern military force, a digital modulation scheme was developed to provide a 19.2-kilobaud rate in a channel spacing of 25 kHz. The minimum-shift-keying modem central to this modulation scheme provides the narrow spectrum and low out-of-channel splatter of FSK and the bit-error-rate performance of PSK. Modulation is accomplished by frequency shift between two phase-locked signals separated from the modem center frequency by plus or minus one-quarter of the clock rate. Demodulation is accomplished by a phase-locked-loop, clock-and-carrier-reference extraction circuit that provides reference for a phase demodulator and level detector unit.

THE SUITABILITY of a digital modulation technique for military tactical radio depends on a large number of performance and cost factors. These factors must be considered in total, since a modulation technique which seems attractive based on theoretical communications efficiency may not be feasible due to circuit complexity, system incompatibility, sensitivity to signal distortion, excessive power consumption, or many other factors.

The goal of reducing the channel spacing for tactical radios affects the modem design by constraining the modulation waveform spectrum width. This requirement affects the system in two ways. First, since the optimum rf bandwidth minimizes the signal-to-noise ratio required to achieve a specified channel bit-error rate, a departure from this optimum will lead to less than optimum performance, although in many cases, the reduction in performance is slight. Second, low adjacent-channel splatter is achieved by a combination of appropriate modulation technique and filtering. The filtering can be either pre- or post-modulation. The choice of a modulating technique with inherently low sideband splatter is of benefit in reducing the filtering complexity and possible attendant delay distortion problems.

Various digital modulation techniques such as FSK, PSK, MSK, PEK and others have been developed for a number of different applications. Each has its own advantages and problems. Consideration of the various factors led to the choice of MSK as the modulation method to be implemented as a working model. MSK is essentially an FSK signal with peak-to-peak deviation equal to one-half the symbol rate and with modulation coded so that the signal can be demodulated as a coherent four-phase signal. MSK provides the narrow spectrum and low out-of-channel splatter of FSK, and the power advantage available from correlation detection. This improved performance is achieved at the

Authors Hannum (left) and Sass



Earl J. Sass, Ldr., Advanced Development, Communications Equipment Engineering, Government Communications Systems, Communications Systems Division, Camden, N.J., received the BSEE in 1945 from the University of Nebraska and the MSEE in 1954 from the University of Pennsylvania. He joined RCA after receiving his undergraduate degree and worked on rf and i.f. coils and transformers and the first RCA printed-circuit television tuner. He was then assigned responsibility for product design of picture and sound i.f. amplifiers for RCA's first commercial color television receivers. Next, he supervised half of the electrical design of commercial color television receivers and remote control receivers. He then transferred to RCA's Government Communications Systems Division where he was responsible for the design 1) of the ground receiving equipment for the DYNA-SOAR Project and 2) a part of the GRC and 744 project. Later he was responsible for the design and development of the AN/TRC-97 exciter, receiver, and shelter. At present, he is a group leader responsible for a portion of the advanced development of military communication and electronic warfare equipments. He holds three US patents and is the co-author of six published articles. He is a Senior Member of the IEEE, and a registered Professional Engineer in the State of New Jersey. He is a member of the Armed Forces Communicaitons and Electronics Association, Sigma Tau, and Pi Mu Epsilon. Mr Sass received an RCA Professional Excellence Award in 1967.

John R. Hannum, Advanced Development, Communications Equipment Engineering, Government Communications Systems, Communications Systems Division, Camden, NJ, received the BSEE in 1969 from Syracuse University and the MSEE in January, 1972, also from Syracuse University. While at Syracuse, he worked as a technician and engineer in the Medical Electronics Laboratory where he assisted in the development of electrocardiographic and magnetocardiographic instrumentation. He joined RCA in 1970 and worked on i.f. filters for the RCA MIC up-down converter and also assisted in the development of the RCA digital synthesizer for a high performance hf receiver. He was next given responsibility for the development of an MSK modem as part of the Tactical Radio Communications Study (TRCS). Following this, he assisted in the development of a compressive receiver. Because of his background in computer theory and application, he was also made responsible for engineering support of the Automatic Test Facility (ATF). In this capacity, he has developed a number of programs to enhance the ATF measurement capability. At present, he is part of the group working on a portion of the advanced development of military communication and electronic warfare equipments.

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expense of more complex circuits than are required for simpler systems, such as FSK.

Although very little comparative experimental data is available, theoretical analysis indicates that the whole class of quaternary PSK modulation methods, of which MSK is one method, rate lower in performance for impulse noise and multipath than straight FSK systems.

MSK theory

Although MSK is continuous-phase, coherently-detected FSK with a deviation of 0.5, it is more instructive to think of MSK as a continuous-phase PSK signal. This is done in the following discussion. At the end of the discusison, it will be shown how the MSK signal can be generated by an FSK method.

Suppose that we have a bit stream such as Fig. 1a. The bits in this stream can be encoded into the continuous phase shift of a carrier frequency as shown in Fig. 1b. In this case, a 0 is coded as a −90° phase change in the bit interval in which the 0 occurs, and 1 is encoded as a +90° phase change in its corresponding bit interval. A transmitted carrier having the phase shifts shown in Fig. 1b is given in Fig. 1e. Such a carrier can be generated by the addition of two signals — an I and a Q signal. The I signal, Fig. 1c, is a sinusoidally amplitude-modulated carrier signal with a phase that is either 0° or 180°. The Q signal, Fig. 1d, is similar, but its phase is either +90° or −90°. One form of MSK modulator works by generating the I and Q signals as shown and adding them together at its output.

Regardless of how the MSK signal is generated, it apperars as shown in Fig. 1e — a constant-amplitude, continuous-phase, PSK, signal. For a bit stream containing 1's and 0's, the spectrum of the transmitted carrier contains two components: one at carrier frequency plus one-quarter the clock frequency and the other at the carrier frequency minus one-quarter clock frequency. These components can be used to phase lock two corresponding VCO's in the demodulator. Appropriate combination of the VCO outputs results in the I-ref. signal, Fig. 1f, and Q-ref. signal, Fig. 1i.

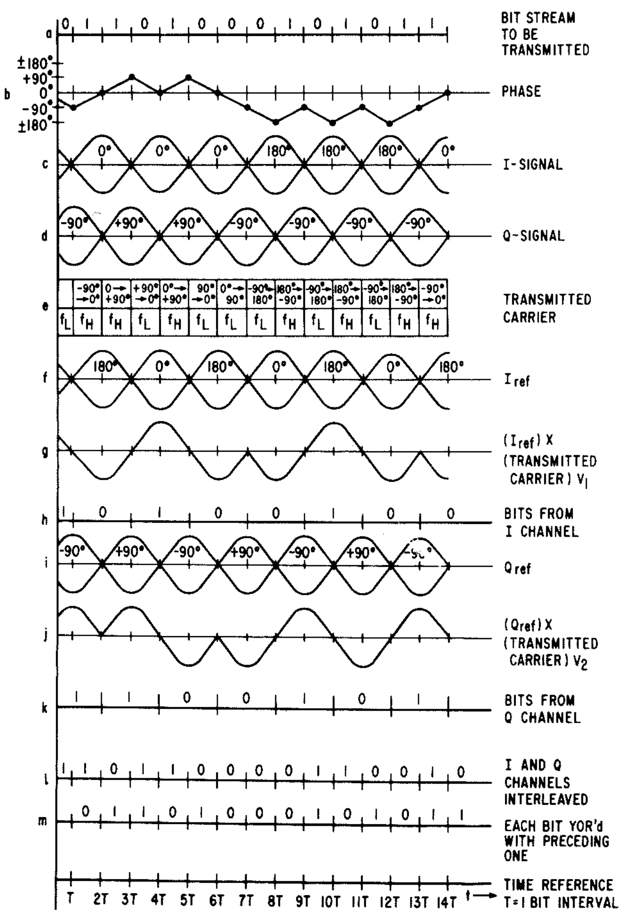


Fig. 1 — Signal phase and timing for minimum shift keying (MSK).

The transmitted carrier signal is given by

$$V_{carrier} = \sin(\omega_c t + \phi) \qquad (1)$$

where ϕ is the variable carrier phase. The I-ref. signal is given by

$$V_{I(ref)} = \sin[(\pi t/2T) + (\pi/2)] \sin(\omega_c t) \qquad (2)$$

where $T = 1$ bit period.

Mixing these two signals and eliminating terms at twice the carrier frequency gives

$$V_{carrier} \times V_{I(ref)} = \tfrac{1}{2}\sin[(\pi/2T) + (\pi/2)] \cos\phi = V_1 \qquad (3)$$

voltage V_1 is shown in Fig. 1g. In a similar fashion, the Q-ref. signal is given by

$$V_{Q(ref)} = \sin(\pi t/2T) \sin[(\omega_c t) - (\pi/2)] \qquad (4)$$

with T as in Eq. 2.

Mixing $V_{q(ref)}$ and $V_{carrier}$ and eliminating terms at twice the carrier frequency gives

$$V_{carrier} \times V_{q(ref)} = \tfrac{1}{2}\sin(\pi t/2T) \cos[\phi + (\pi/2)] = V_2$$

Voltage V_2 is shown in Fig. 1j.

Voltages V_1 and V_2 constitute I- and Q-channel signals respectively. They are run through separate level detectors. The resulting logical values are given in Figs. 1h and 1k. The level detectors are set up so that a positive level of the input signal (V_1 or V_2) gives a 1 and a negative level gives a 0. Note that in each channel, each bit lasts for two bit times ($2T$).

The bit streams from the I and Q channels are now interleaved. In the bit interval ending at T, we sample a 1 from the I channel. In the interval T–$2T$, we sample a 1 from the Q channel. In the interval $2T$–$3T$, we sample a 0 from the I channel, and so on, alternating channels each bit interval. The resulting interleaved stream is shown in Fig. 1*l*. Now, starting in the bit interval T–$2T$, each bit is *exclusive-or*'ed with the preceding bit, giving the reconstructed bit stream of Fig. 1m. Thus, the original bit stream is obtained, delayed by 1 bit interval from the transmitted stream.

It has been stated that an MSK signal can be generated by a frequency-shift method rather than by a "brute force" construction of the I signal and Q signal. Suppose at the beginning of a bit interval, the carrier signal is given as

$$V_{carrier} = \sin(\omega_c t + \theta) \qquad (6)$$

where θ is the carrier phase. If a 1 is transmitted during the bit interval; then at the end of the interval, the carrier signal is given by

$$V_{carrier} = \sin[\omega_c t + \theta + (\pi/2)] \qquad (7)$$

Thus, the phase change through one bit interval T is $\pi/2$. This phase change can be accomplished by adding an angular frequency of $\pi/2T$ to the carrier frequency during the bit interval so that the carrier signal becomes

$$V'_{carrier} = \sin[\omega_c t + \theta + (\pi t/2T)] \qquad (8)$$

$$= \sin\{2\pi[f_c + (1/4T)]t + 0\} \qquad (9)$$

for the bit interval. Since $1/T = f_D$ = clock frequency, the frequency transmitted for a 1 is

$$f_H = f_C + f_D/4 \qquad (10)$$

Similarly, if a 0 is transmitted during a bit interval, then the carrier signal at the end of the interval is

$$V_{carrier} = \sin[\omega_c t + \theta - (\pi/2)] \qquad (11)$$

This phase shift can be accomplished by subtracting an angular frequency of $\pi/2T$ from the carrier frequency during the bit interval so that the carrier signal becomes

$$V'_{carrier} = \sin[\omega_c t + \theta - (\pi t/2T)] \qquad (12)$$

$$= \sin\{2\pi[f_c - (1/4T)]\, t + \theta\} \qquad (13)$$

for the bit interval. Again, $1/T = f_D$ = clock frequency, so the frequency transmitted for a 0 is

$$f_L = f_C - (f_D/4) \qquad (14)$$

Since the bit stream consists of 1's and 0's, it can be transmitted by appropriate shifts between these two frequencies as shown in Fig. 1e. Note that because of the requirement for phase continuity at switching point, these two oscillators must be phase-locked to an appropriate standard.

The signal resulting from doubling of the transmitted signal has frequency components $2f_H$ and $2f_L$ contained in it if the source bit stream has both 1's and 0's. These two frequencies can be used to phase lock demodulator oscillators at f_H and f_L. Thus, there are two oscillators in the demodulator:

$$V_H = \sin\{2\pi[fc + (1/4T)]t\} \qquad (15)$$

and

$$V_L = \sin\{2\pi[fc - (1/4T)]t\} \qquad (16)$$

Upon expansion, these two give

$$V_H = \sin(2\pi f_C t)\cos(\pi t/2T) + \cos(2\pi f_C t)\sin(\pi t/2T) \qquad (17)$$

$$V_L = \sin(2\pi f_C t)\cos(\pi t/2T) - \cos(2\pi f_C t)\sin(\pi t/2T) \qquad (18)$$

If V_H and V_L are added, the result is

$$V_H + V_L = 2\sin 2\pi f_C t\cos(\pi t/2T) \qquad (19)$$

$$= 2\sin[(\pi t/2T) + (\pi/2)]\sin(\omega_c t) \qquad (20)$$

which after appropriate amplitude scaling becomes

$$\tfrac{1}{2}[V_H + V_L] = \sin[(\pi\tau/2T) + (\pi/2)]\sin(w_c t) = V_{I(ref)} \qquad (21)$$

If V_H is subtracted from V_L, the result is

$$V_L - V_H = -2\cos(2\pi f_C t)\sin(\pi t/2T) \qquad (22)$$

$$= -\sin(\pi t/2T)\sin[\omega_c t + (\pi/2)] \qquad (23)$$

$$= \sin(\pi t/2T)\sin[\omega_C t - (\pi/2)] \qquad (24)$$

which after appropriate amplitude scaling becomes

$$\tfrac{1}{2}[V_L - V_H] = \sin(\pi t/2T)\sin[w_c t - (\pi/2)] = V_{q(ref)} \qquad (25)$$

It is important to note that the above explanation of MSK is based upon one

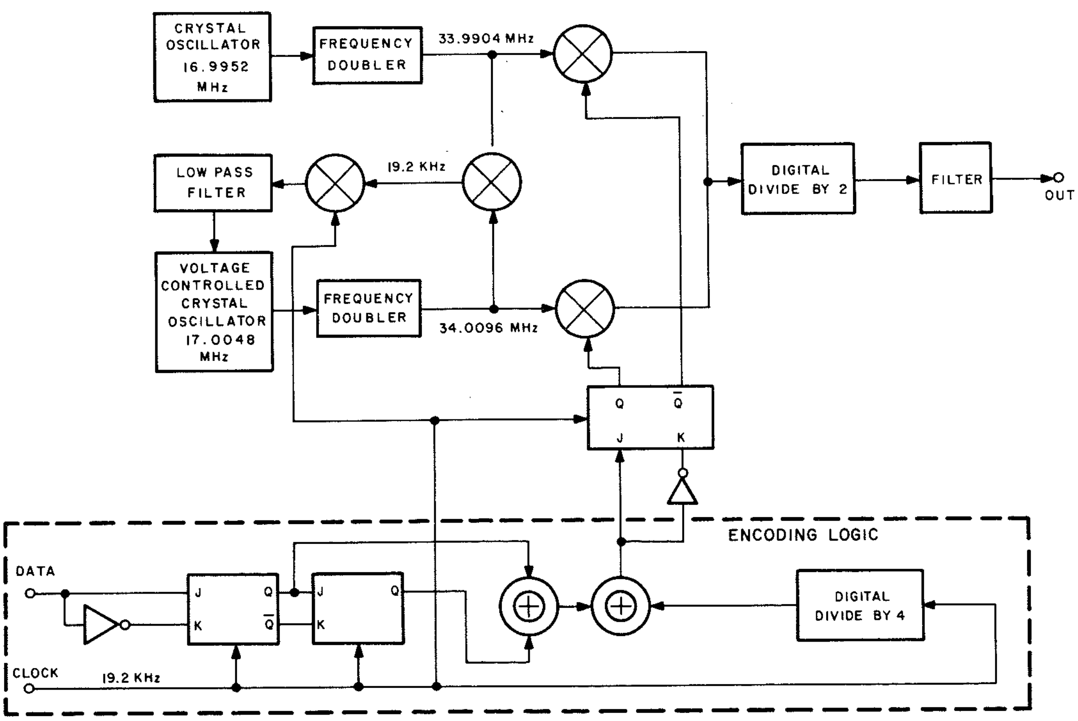


Fig. 2 — MSK modulator circuit.

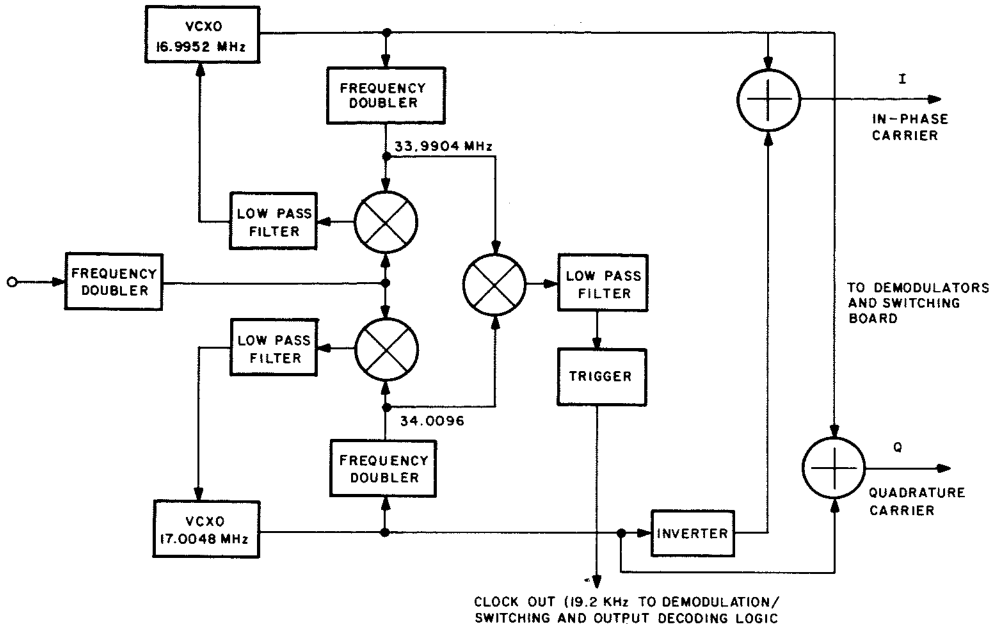


Fig. 3 — Clock and carrier-extraction board provides two reference carriers to the demodulation and switching.

particular set of phase and timing relationships. It would be equally valid to permit a +90° carrier phase change to represent a 0 and a −90° phase change to represent a 1. Also, the polarities and timings of the *I*- and *Q*-reference signals have been chosen for convenience. Other polarities and timings are possible, the only restriction being that the two signals must be in quadrature. Choice of other polarities and timings for the *I*- and *Q*-reference signals may, however, result in more complex interleaving and combining logic.

Since proper polarity and timing of the *I*- and *Q*-reference signals are necessary, some form of synchronization is necessary for the demodulator. For this experimental model, a preamble was transmitted, and a simple manual procedure was used to synchronize the demodulator.

Modulator circuit

The MSK modulator circuit, the block diagram of which is shown in Fig. 2, is based on the MSK demodulator circuit described by DeBuda[1]. The output of an 16.9952-MHz crystal oscillator is doubled to 33.9904 MHz and mixed with a 34.0096-MHz signal which is the doubled output of a voltage-controlled crystal oscillator (VCXO). The difference frequency of 19.2 kHz is mixed with the incoming clock frequency of 19.2 kHz, and the resulting error signal is used to phase lock the VCXO. The outputs of the two frequency doublers are also brought out to electronic switches which select one of the two signals to be switched to the output divide-by-2 and a filter. The selection is made by the control signals generated by the encoding logic (at the bottom of Fig. 2).

Encoding logic

The encoding logic processes the incoming bits in such a way as to prevent sustained transmission of a single frequency which could result if a long string of 1's or 0's is to be transmitted. Sustained transmission of a single frequency could result in loss of phase-lock in the demodulator VCO corresponding to the untransmitted frequency.

The encoding logic stores successive symbols in the incoming bit stream and combines them logically to provide a differential signal. The combined output bit stream is further combined logically with a 4800-Hz signal locked to the clock in such a way as to reverse successive pairs of symbols. The resulting signal causes frequency-shift keying between two frequencies separated by 19.2 kHz. When these two frequencies are halved, reducing the frequency difference to 9.6 kHz, an MSK signal results. In this signal, each successive input bit corresponds to a ±90° change in the phase of the output signal. There are two quadrature reference phases which define the "one" and "zero" phases for successive bits, so that coherent detection can be employed.

Demodulator circuit

The demodulator circuit is divided into three parts: the demodulation and switching board, the clock and carrier-extraction board, and the output-decoding logic board. The first two boards make up a demodulator circuit based on that of DeBuda[1]. The last board, the output decoding logic board, receives the encoded output of the demodulator and converts it to a bit stream that is a replica of the input bit stream but delayed in time and synchronous with the clock generated on the clock and carrier-extraction board.

Clock-and-carrier-extraction board

Fourier analysis of the doubled MSK wave shows that it contains both high and low frequency components of sufficient amplitude to maintain phase lock of both loops shown in Fig. 3. The outputs of the frequency doublers in the loops are mixed to provide the clock reference signal. The two phase-locked oscillator signals are added to obtain an amplitude modulated quadrature carrier (*Q*). The higher-frequency signal is subtracted from the lower-frequency signal to obtain an amplitude modulated in-phase carrier reference (*I*). These two carriers are used as reference signals on the demodulation-and-switching board (Fig. 4).

Demodulation and switching board

The incoming MSK signal is mixed with the *I* and *Q* carrier reference signals to give the *I* and *Q* demodulation channels. The mixer outputs in each channel pass through low-pass filters with sufficiently long time constant to bypass rf coming

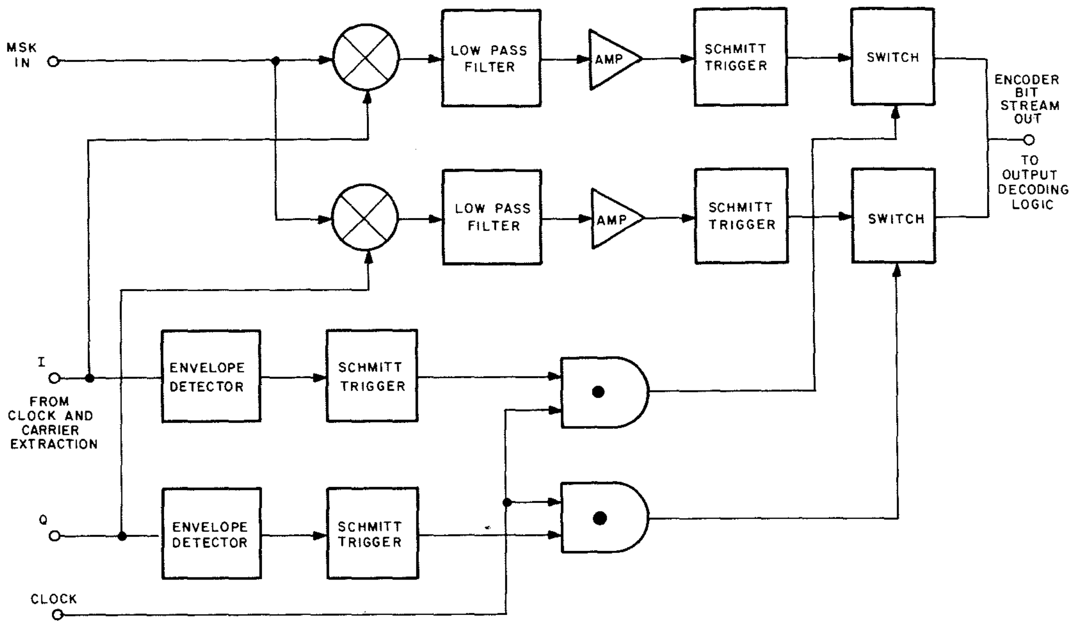


Fig. 4 — Demodulation and switching board.

from the mixer and to pass the desired mixer output with minimum of noise. The filter output is amplified and applied to a Schmitt trigger. Each filtered mixer output provides approximately a matched-filter output for the successive input symbols. Because of the method of derivation of the references, the sign of every other symbol from each mixer is reversed. After amplification, each signal changes the output level of the succeeding Schmitt trigger. This serves to convert each signal output to a rectangular wave with transitions determined by the input signals.

The amplitude-modulated I and Q reference carriers are detected, and the peaks of the detected envelopes generate gating signals through Schmitt triggers. The trigger circuit outputs are gated by the clock to provide clock-synchronized switching signals to activate the switches on the outputs of I and Q demodulation channels which sample the two channel outputs at the proper time. The overall timing is such that the output is switched from one demodulation channel to the other for each successive clock pulse. The resulting output bit stream is an encoded version of the original bit stream put into the modulator.

Output-decoding logic board

The output decoding logic is shown in Fig. 5. This logic essentially reverses the effect of the encoding logic at the modulator input when the "divide-by-four" has the proper phase. The *and* gate at the output gates the output bit stream in synchronism with the local clock signal.

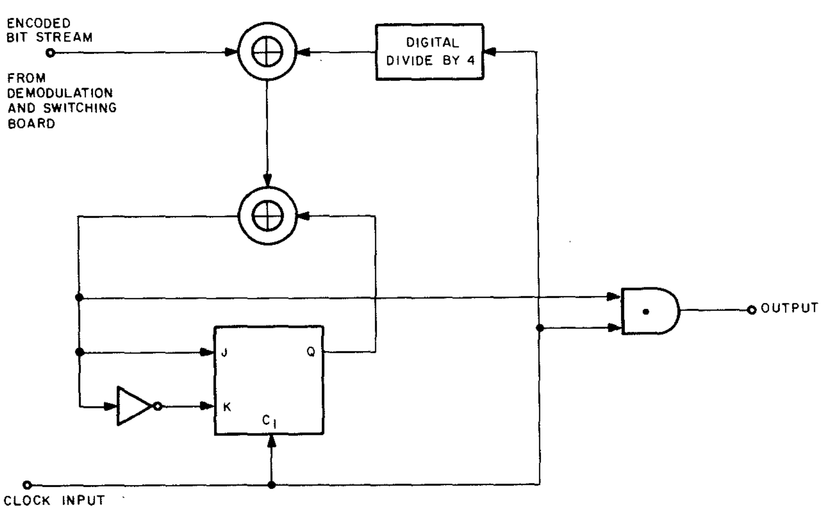


Fig. 5 — Output-decoding logic.

Implementation

Modulator and demodulator breadboards have been constructed at RCA using discrete components for the analog circuits, CMOS logic for the low frequency digital circuits, and TTL logic for the high frequency digital circuits. All the boards operated without problems.

The breadboard circuits were reviewed to determine what could be achieved by hybridization, using a standard size of 1.5×1.1×0.13 inch for the hybrid modules. The following table is an estimate of what can be achieved.

Function	*Hybrid modules*	*Discrete components*
Modulator	3	3 mixers*
Clock and carrier extractor	3	3 mixers
Demodulator and decoder	3	2 mixers

*Four mixers required; however, one can be active and part of the hybrid module.

Bit-error rate

The bit-error-rate performance of the

MSK modem was measured as a function of the received-signal noise bandwidth. The noise bandwidths of the filters were 24.7 kHz, 16.0 kHz, and 14.7 kHz. The roll-off of the filters was Gaussian to 12 dB to minimize intersymbol interference. Fig. 6 shows how the bit-error rate changes as a function of E_b/N_o and noise bandwidth. The best performance is achieved at a noise bandwidth of 16 kHz, where the amount of sideband attenuation is not sufficient to cause greater distortion than the restriction of the noise.

MSK signal spectrum

Fig. 7 shows the actual MSK spectrum at the modulator output for a pseudo-random input bit stream. Note the important spectral components at carrier frequency ±one-quarter of clock frequency (≅5 kHz). Comparison of Figs. 7 and 8 shows that the actual MSK spectrum is quite close to theoretical. The general shape is the same, and specific points show excellent agreement. For example, at $X = 1$ (19.2 kHz away from the center frequency), $X = 3$ and $X = 5$, the actual spectrum is within a few tenths of a dB of theoretical.

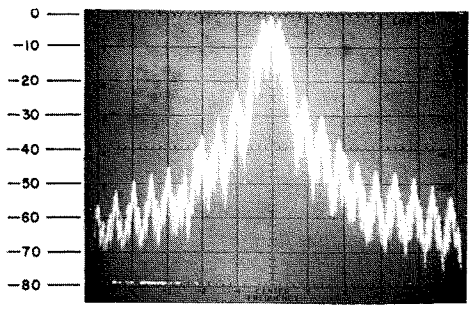
Conclusion

The RCA-developed modem described in this paper has been tested in the laboratory with interesting results. In a system with a noise bandwidth of 16 kHz, a bit-error rate of 1×10^{-3} was obtained with an E_b/N_o level of 6.9 dB, and a bit-error rate of 1×10^{-5} was obtained with an E_b/N_o of 9.8 dB. These results are within 0.1 dB of the values for an ideal PSK system.

Measurement of the MSK signal spectrum shows it to be within a few tenths of a dB of theoretical. The 99% energy bandwidth was measured as 23.0 kHz or 1.2 times the bit rate (which is the theoretical value).

Acknowledgment

The authors acknowledge the help of Dr. T.T.N. Bucher, RCA Staff Engineer, who guided on the Tactical Radio Communications Study (TRCS) program; R.R. Rugarber, who was the Contracting Officer's Technical Representative on the TRCS program; and Lee Gregory, RCA engineer in training, who assisted in the test of the modem.



Scan width: 20 kHz/div.
Scan time: 0.2 sec./div.
Bandwidth: 1 kHz

Fig. 7 — Actual MSK spectrum.

Reference

1. DeBuda, Rudi; "The Fast FSK Modulation System," IEEE International Conference on Communications, *Conference Record*, Montreal, Canada, (June 14-16, 1971).

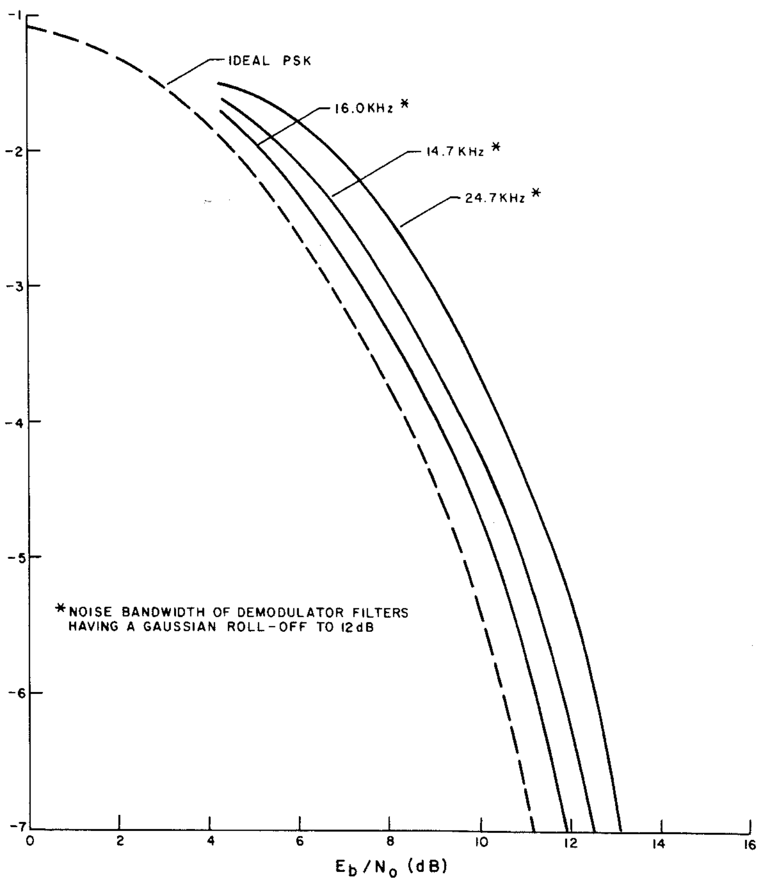


Fig. 6 — MSK bit-error rate versus Eb/No for noise bandwidths of 16.0 kHz, 14.7 kHz, and 24.7 kHz. [Eb is the signal energy as measured over the duration of one bit of information; No is the noise power as measured in a 1-Hz bandwidth.]

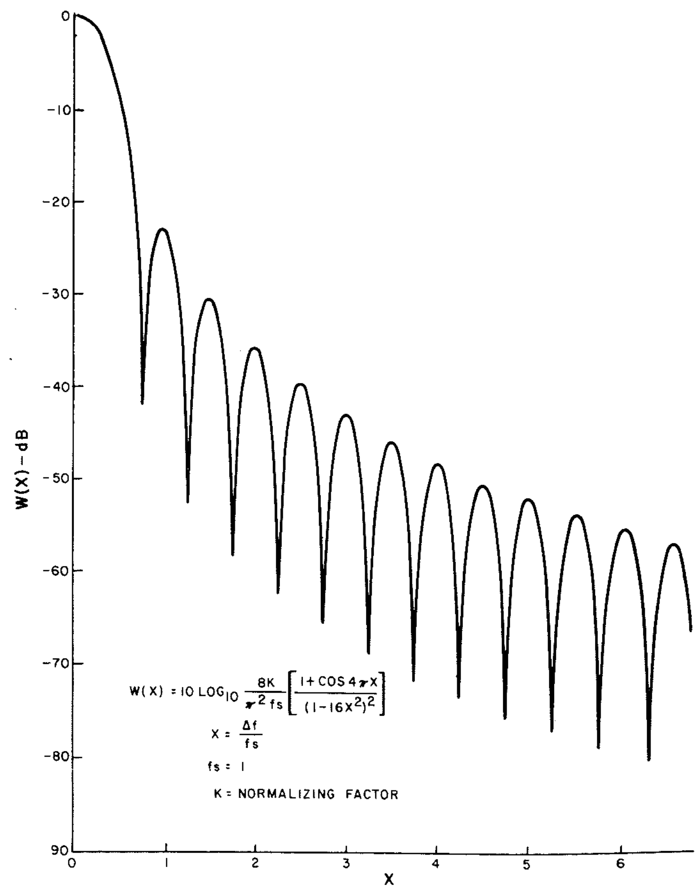
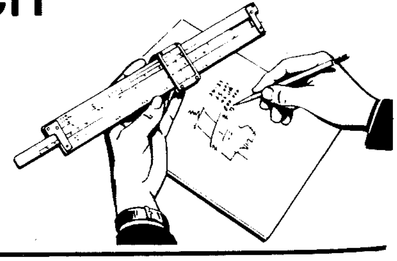


Fig. 8 — Theoretical MSK power spectrum.

Engineering and Research Notes



Apparatus for measuring cable elongation

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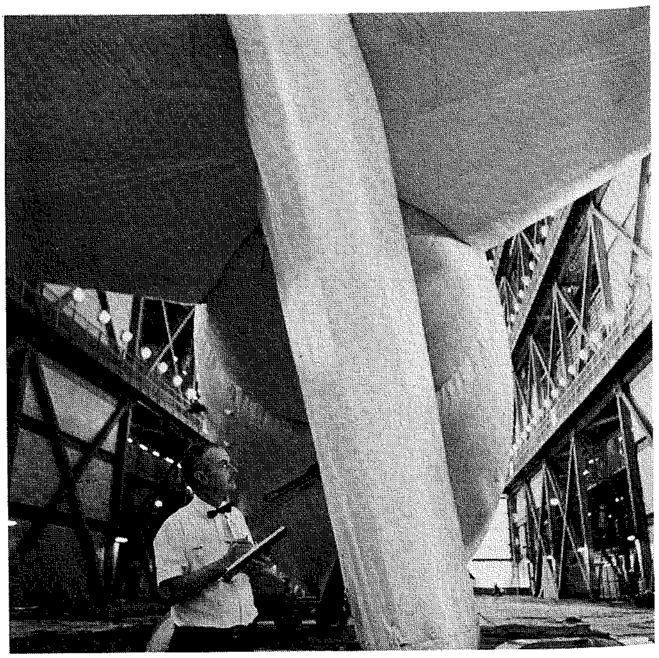

Author Edward L. Crosby monitoring a fin pressurization test on T/N 204. This is one of several 200,000 cu. ft. tethered balloons being developed for the Defense Advanced Research Projects Agency, DOD, by the Range Measurements Laboratory, AFETR. Technical support in the categories of development, engineering, operations and analysis is provided to the RML by RCA's TELTA Project, managed by R.P.Murkshe. In the photo, the balloon is air inflated inside the Vehicle Assembly Bldg., VAB, with the cooperation of NASA KSC, for equipment installation, checkout and aerostatic calibration.

Progress in applied polymer chemistry in the past two decades has provided industry with synthetic fibers that are far superior to natural fibers (*e.g.*, manila used in rope) and surpass even steel in their strength-to-weight ratios. These new synthetics, which are thermoplastic elastomers, exhibit mechanical properties that deviate from the classical principles of mechanics in several respects. Besides departure from Hooke's Law, they display *stress-creep* and *stress-memory*. This latter term means that when a cable of this type is stressed in tension, and thus stretched, it will not relax completely to its original length when the tension is removed. The magnitude of this permanent elongation depends upon several factors of which the magnitude of stress is most important. Subsequent stress and relaxation leaves the cable with increased permanent elongation.

For a given composition and construction, the ultimate yield (i.e., break) will occur at a fixed percent elongation which is typical. It has been observed that when a cable has accumulated permanent elongation of this magnitude from repeated and/or prolonged stress, it is liable to break. Therefore, knowledge of the permanent elongation and deformation of a cable produced with thermoplastic materials is believed to be a reliable method of predicting, and therefore preventing, cable failure.

However, when monitoring cable elongation it is necessary to mark the cable so that its permanent elongation after the application of stresses may be measured. Typically, cables using thermoplastic fibers as the tensile member are jacketed with a tough, soft material such as a polyolefin, *e.g.*, extruded polyethylene. To so mark a cable or otherwise identify a given length thereof, before and after applied stresses, has been a difficult problem. Any markings applied to the outer surface of the cable jacket have proven to be unreliable because the jacket material is usually extremely "slick" and difficult to mark. Moreover, when the cable is utilized with sheaves and capstan grooves on a winch, the outer jacket of the cable is cold-worked which tends to destroy any markings or identification placed on the outer jacket thereby defying any effort to permanently mark the outer jacket. Additionally, differential tension present between the jacket and the cable core results in displacement of the jacket with respect to the inner strength member such that any physical tagging of the jacket does not effectively tag or locate a given point on the inner strength member. Further, the markings on the jacket cannot be such as to degrade the strength of the cable in any way, or the integrity of the jacket.

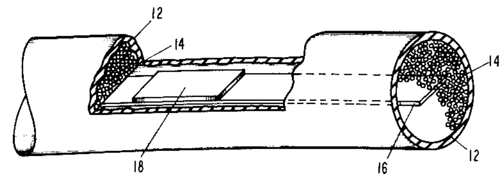

Fig. 1 — Perspective fragmented view of a cable constructed of synthetic fibers.

In Fig. 1, a cable fabricated of synthetic materials includes an outer jacket (12) made of polyolefin such as polyethylene, or the like, and inner strength core (14) made of a large number of strands of polyester fibers which are strung together as known in synthetic cable manufacturing art. During the manufacture of the cable, a plastic tape (16), such as polyster film, having a length the same as cable, is disposed centrally among the fibers of core along the entire length of cable. The tape may be formed within the cable core by any suitable technique such as feeding the tape into a core-forming die (not shown) which arranges the fibers of the core.

At equally spaced intervals on the tape are strips of conductive foil (18). These foil strips may be pieces of conductive material of uniform dimensions which are bonded to the tape (16), or they may be formed of a conductive material which is electro-deposited on the tape by any suitable method. The finished cable then comprises an outer jacket (12), the inner core (14) and tape (16) having deposited, at regular intervals, identical conductive foil strips (18). It has been found that the tape, being of substantially the same material as the polyster fibers of the core, does not diminish cable performance in strength or durability and, moreover, stretches with the cable. On the other hand, the presence of the spaced strips of foil permits detection of each of the foils by a suitable rf detector (Fig. 2). The rf detector is placed adjacent to the cable, and when the foil passes in the proximity of the detector, a signal is generated indicating this condition. The spacing of the foils is measured by noting the cable length between the detected indications and comparing the lengths before and after stress is applied to the cable.

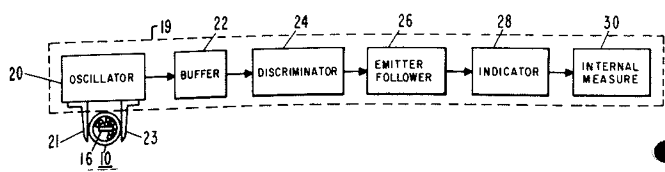

Fig. 2 — Block diagram of a suitable detector utilized in measuring the amount of cable elongation.

In Fig. 2, the rf detector includes an oscillator (20) coupled to a pair of probes (21) and (23); a buffer circuit (22); and a discriminator circuit (24) which converts frequency to voltage, the output voltage having a level calibrated to the input frequency. An emitter follower circuit (26) provides coupling to indicator device (28). The indicator includes a meter for measuring the voltage amplitude at the output of the emitter follower and an alarm or light circuit to provide an audible or visible indication of the presence of a foil strip between the probes. The output on the indicator device may be applied to an interval-measuring circuit. Alternatively, the foil spacing may be measured with a hand-held detector and a tape measure.

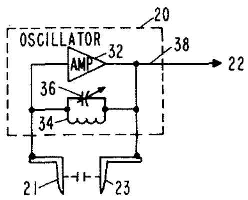

Fig. 3 — Suitable oscillator used in the circuit of Fig. 2.

In Fig. 3, a suitable oscillator includes an amplifier and a resonant circuit having a variable capacitor. The output of oscillator is applied to a buffer. The output is in the rf range. The variable capacitor is adjusted to tune the output frequency of oscillator to provide a given voltage at the output of the discriminator (24, Fig. 2). This zeros the meter of the indicator. Coupled to each side of the resonant circuit (as shown in Figs. 2, 3 and 4) is a short length of a conventional aluminum angle, or the like, which forms the pickup probes (21 and 23). The spacing of the probes from each other provides a capacitance across the resonant circuit which tunes the oscillator to a given frequency. A shift in the capacitance between the probes will shift the frequency of the oscillator.

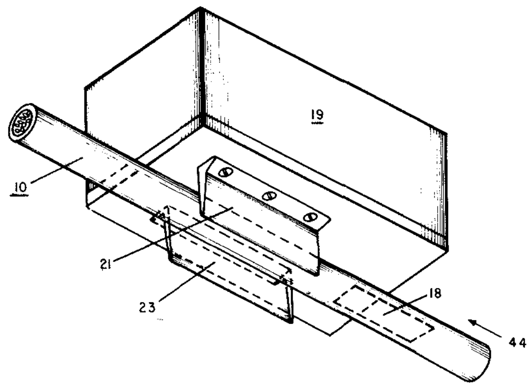

Fig. 4 — Perspective of the RF detector assembly and a cable under test.

In operation, the probes (21 and 23) are placed astride the cable as shown in Fig. 2 and 4. The cable is then moved in the direction of the arrow (44, Fig. 4) between probes 21 and 23. However, when a foil strip (18) in the cable passes between the probes, the capacitance between the probes is increased in value. This detunes the oscillator, providing a shift in frequency, which results in the discriminator producing a different voltage level at the output. This voltage level is detected to provide an audible or visible alarm and provide the meter reading for calibration purposes during initial setup. The detected signal is applied to the interval-measuring device which measures the interval between the time of occurrence of these shifts in oscillator output frequency.

To provide precise measurement, the cable should be passed in both directions between the probes and the cable marked upon the occurrence of the signal from the indicator. The midpoint of these two markings, should they not coincide, provides the exact center of the foil. This center can then by precisely measured to the center of the next adjacent foil. Of course, it is to be understood that a reference point on the cable, such as one end, or a winch calibrated with a footage meter or the like, provides a reference indication as to the identification of the respective foils during the initial marking of the distance between foils before and after cable use.

Since the elongation will occur in the portion of cable that is stressed the greatest, only the most heavily stressed portion of the cable need be monitored. Cables made of thermoplastic materials have found widespread use as tether cables for communication balloons, helicopter lifting and towing, power-line rigging, power logging, submarine and oceanographic work, rescue operations, and construction work. By providing means for measuring cable elongation, such cables may be utilized for these desirable applications with the greater assurance of safety provided.

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J - K' bistable multivibrator

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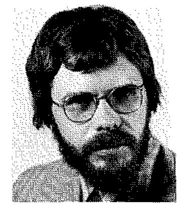


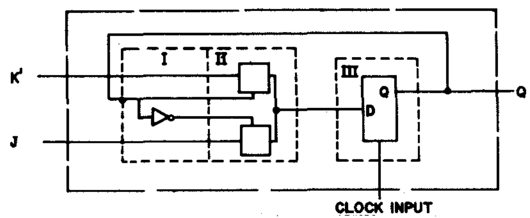


Fig. 1 — J-K' bistable multivibrator.

Fig. 1 is a schematic of a *J-K'* bistable multivibrator useful in any sequential binary logic circuit, particularly in digital control sequencers wherein data-type flip-flops are primarily used, and it is desired to perform the *J-K* flip-flop function.

The *J-K'* bistable multivibrator of Fig. 1 is logically equivalent to a *J-K* bistable multivibrator but is responsive to the *J-K'* logic inputs. This multivibrator is a combination of three logic blocks which are commonly found in MOS logic families. In some families, such as RCA ATL standard cell LSI logic, Blocks I and II actually are contained in a single block, or cell. Block I is an inverter, Block II contains two transmission gates, and Block III is a clocked data-type flip-flop (D flip-flop), all of which are standard devices whose operations are well known. It should be noted that if the data flip-flop has a Q' output, so will the *J-K'* flip-flop.

J	K'	Q	D
0	0	0	0
0	0	1	0
0	1	0	0
0	1	1	1
1	0	0	1
1	0	1	0
1	1	0	1
1	1	1	1

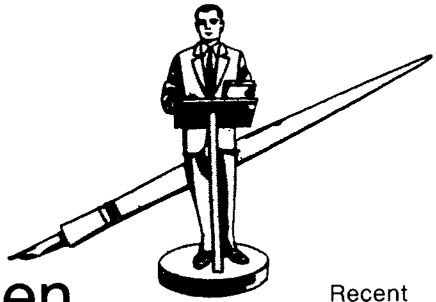
Table I — Truth table of "D" of Block III.

J	K'	K	$Q(t_{n+1}) = D(t_n)$
0	0	1	0
0	1	0	$Q(t_n)$
1	0	1	$Q'(t_n)$
1	1	0	1

Table II — Transition table J-K' for bistable multivibrator.

The logic input at the "D" terminal of Block III is readily determined to be $QK' + Q'J = D$. The truth table for this function is given in Table I. Based on the known operation of the data-type flip-flop, Table II may be derived. Table II shows the Q output during time frame t_{n+1} (equal to the Q output during time frame t_n) as a function of the inputs J and K'. The time frames are defined by one cycle of the clock input. Also shown in Table II for reference is K (the inverse of K'). Thus, the composite unit acts as a J-K flip-flop with inputs J and K.

Reprint RE-4-25| Final manuscript received April 26, 1973.



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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.

SERIES 100 BASIC THEORY & METHODOLOGY

170 Manufacturing and Fabrication

production techniques for materials, devices and equipment.

INTERCONNECTIONS — B.R. Schwartz (MSRD,Mrstn) *Electromechanical Design*; 10/73

175 Reliability, Quality Control and Standardization

value analysis, reliability analysis, standards for design and production.

CMOS PROCESSOR, Survivable — S. Teitelbaum (AED,Pr) Hardened Guidance and Weapon Delivery Tech. Conf., Orlando, Fla.; 10/16-18/73

HARDNESS TESTING, MNOS/SOS — G. Brucker (AED,Pr) Hardness Guidance and Weapon Delivery Tech. Conf., Orlando, Fla.; 10/16/73

PROCESS, Communications and the — E.S. Shecter (MSRD,Mrstn) Joint Engineering Management Conf., St. Petersburg, Fla.; 10/25/73; *Conf. Proc.*

180 Management and Business Operations

organization, scheduling, marketing, personnel.

COST CONTROL on a large program in today's competitive environment, The growing importance of costs and ways to maintain — J. Newman (AED,Pr) 21st Joint Engineering Management Conf.; *Proc.*; 10/25/73

DESIGN-TO-COST CONTRACTING guidelines, Summary of American Preparedness Association — C.D. Fisher (CSD,Cam) V.E. Division of ADPA Mtg., Ft. Belvoir, Virginia; 10/15-16/73

PROJECT MANAGEMENT—future developments — M.W. Buckley (MSRD,Mrstn) Meeting Chairman and Lecturer, AMA Seminar, Los Angeles, Calif.; 10/8/73

PURCHASING problems, Blanket orders: a solution to — J. Lyons (EC,Har) 65th Annual SLA Conf., Toronto, Canada; 6/9-13/73

SERIES 200 MATERIALS, DEVICES, & COMPONENTS

205 Materials (Electronic)

preparation and properties of conductors, semi-conductors, dielectrics, magnetic, electro-optical, recording, and electro-magnetic materials.

GETTER for vacuum devices — J.C. Turnbull, W.H. Silvers (EC,Har) Sixth Int'l Vacuum Congress, Kyoto, Japan; 3/25-29/73

215 Circuit and Network Designs

analog and digital functions in electronic equipment; amplifiers, filters, modulators, microwave circuits, A-D converters, encoders, oscillators, switches, masers, logic networks, timing and control functions, fluidic circuits.

INTERPOLATIVE COMPRESSOR, Two-dimentional — M. Hecht, S. Henry (AED,Pr) Management and Utilization of Remote Sensing Data; American Society of Photogrammetry, Sioux Falls, S.D.; 10/30/73

220 Energy and Power Sources

batteries, solar cells, generators, reactors, power supplies.

SOLAR CELLS, Temperature, illumination, and fluence dependence of current and voltage in electron irradiated — T.J. Faith, A.F. Obsenschain (AED,Pr) IEEE Photovoltaic Specialists Conf., *Proc.*; Palo Alto, Calif.; 11/11-13/73

225 Antennas and Propagation

antenna design and performance; feeds and couplers, phased arrays; randomes and antenna structures; electromagnetic wave propagation, scatter, effects of noise.

ANTENNA ARRAY, A wide-band, short-backfire uhf — W. Maxwell, H.H. Rickert (AED,Pr) 23rd Annual USAF Antenna Symp.; Monticello, Ill.; 10/11/73

CONFORMAL ARRAYS come of age — B.F. Bogner (MSRD,Mrstn) *Microwave J.*; 10/73

REFLECTIONS, Another look at — W. Maxwell (AED,Pr) First Army M.A.R.S., N.J. Division; Ft. Monmouth, N.J.; 9/23/73. A.R.R.L. Convention, New England Div., Hyannis, Mass.; 9/30/73. E. Brunswick Amateur Radio Club, E. Brunswick, N.J.; 10/10/73. Mt. Airy VHF Society, Southampton, P.A.; 11/15/73

RF MANIFOLD, Ultra lightweight stripline — B.F. Bogner (MSRD,Mrstn) Eleventh Electrical/Electronic Insulation Conf., Chicago, Ill.; 10/3/73; *Conf. Proc.*

240 Lasers, Electro-Optical and Optical Devices

design and characteristics of lasers, components used with lasers, electro-optical systems, lenses, etc. (excludes: masers).

RADIOMETER, Very high resolution — A.I. Aronson (AED,Pr) Delaware Valley Section of the Optical Society of American, Hightstown, N.J.; 9/25/73

250 Recording Components and Equipment

disk, drum, tape, film, holographic and other assemblies for audio, image, and data systems.

RECORDING system, User-oriented image data cartridge — C.R. Horton (CSD,Cam) Int'l Telemetering Conf., Washington, D.C.; 10/9-11/73; *Proc.* Vol. 9, page 86

TAPE RECORDER for earth observation satellites, A 40MHz multitrack — C.R. Thompson (AED,Pr) Int'l Telemetry Conf.; Washington, D.C.; 10/9-11/73

TAPE RECORDER for spacecraft applications, An ultra high rate digital — C.R. Thompson (CSD,Cam) Int'l Telemetering Conf., Washington, D.C.; 10/9-11/73' *Proc.* Vol. 9, p. 220

TAPE RECORDING system, A high capacity, high data rate instrumentation — O.E. Bessette (CSD,Cam) *Int'l Telemetering Conf. Proc.* Washington, D.C.; Vol. 9, p. 74; 10/9-11/73

SERIES 300 SYSTEMS, EQUIPMENT, & APPLICATIONS

305 Aircraft and Ground Support

airborne instruments, flight control systems, air traffic control, etc.

AIR TRAFFIC-CONTROL facilities, alternatives for fixed VFR terminal — J.N. Ostis (ASD,Burl) Int'l Aerospace Show, Sao Paulo, Brazil; 9/14-23/73

310 Spacecraft and Ground Support

spacecraft and satellite design, launch vehicles, payloads, space missions, space navigation.

ATMOSPHERE EXPLORER C,D, and E, Some design aspects of the low perigee — B. Steward (AED,Pr) IAF 24th Int'l Congress, *IAF PROC.*; Baky, USSR; 10/7-13/73

ERTS (Earth Resources Technology Satellite) data be previewed? Can tape — A.F. Martz (AED,Pr) Management and Utilization of Remote Sensing Data, American Society of Photogrammetry; Sioux Falls, S.D.; 10/29-11/2/73

GEOMETRIC CORRECTION—constant resolution processor — J. Danko, M. Hecht, T. Altman (AED,Pr) Management and Utilization of Remote Sensing Data; American Society of Photogrammetry, Sioux Falls, S.D.; l0/30/73

INFORMATION and less data, The need for more — H.M. Gurk (AED,Pr) Management and Utilization of Remote Sensing Data, American Society of Photogrammetry; Sioux Falls, S.D.; 10/29-11/2/73

PROPULSION system, Block-5D ascent phase — D.L. Balzer (AED,Pr) 1973 JANNAF Propulsion Mtg.; *Chemical Propulsion Information Agency J.*, Las Vegas, Nev.; 11/6-8/73

315 Military Weapons and Logistics

missiles, command and control.

WEAPON SYSTEM, The AEGIS — F.G. Adams (MSR,Mrstn) Philadelphia Chapter-American Society Naval Engineers/Society Naval Architects and Marine Engineers, Moorestown, NJ; 10/17/73

325 Checkout, Maintenance, and User Support

automatic test equipment, (ATE), maintenance and repair methods.

TESTING, Latest methods of functional and diagnostic — F. Pfifferling, C.A. Schmidt (MSRD,Mrstn) IEEE Wescon: Producibility Workshop; 9/13/73

345 Television and Broadcast

television and radio broadcasting, receivers, transmitters, and systems, television cameras, recorders, studio equipment.

VIDEO RECORDING format, Possibilities for a new compatible quadruplex — J.L. Grever (CSD,Cam) 114th Conf. of SMPTE, New York City; 10/18/73

TV TRANSMITTER operation, Design considerations for unattended — T.M. Gluyas (CSD,Cam) RCA Broadcast Consultants Seminar, Washington, D.C.; 11/9/73

BROADCAST VIDEO RECORDING, The effect of the choice of tape format on — K. Sadashige (CSD,Cam) 23rd Annual IEEE Broadcast Symp. : Washington, D.C.; 9/20-21/73

TELEVISION FILM, A cartridge system for 16mm — A.H. Lind, A.E. Jackson (CSD,Cam) IEEE Broadcast Symp.; Washington, D.C.; 9/20-21/73

380 Graphic Arts and Documentation

printing, photography, and typesetting; writing, editing, and publishing; information storage, retrieval, and library science.

VISUAL AIDS, Communication with — C.W. Asbrand (ASD,Burl) *J. of Society for Technical Communications*; 8-9/73

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Subject listed opposite each author's name indicates where complete citation to his paper may be found in the subject index. An author may have more than one paper for each subject category.

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Simulated load for internal combustion engines — R. T. Cowley, L.R. Hulls (ASD, Burl.) U.S. Pat. 3757570, September 11, 1973.

Simulated load for internal combustion engines — G.A. Chamberas (ASD, Burl.) U.S. Pat. 3757571, September 11, 1973.

Holographic readout system employing predispersion diffraction grating — B.R. Clay and D.A. Gore (ASD, Burl.) U.S. Pat. 3754808, August 28, 1973.

Electromagnetic and Aviation Systems Division

Communications between central unit and peripheral units — R.A. Ito (EASD, Van Nuys) U.S. Pat. 3766530, October 16, 1973.

Astro-Electronics Division

Cutting pliers — E.J. Von Makkay and J.J. Pesce (AED, Pr.) U.S. Pat. 3763560, October 9, 1973.

Phase measuring system — F. B. Griswold (AED, Pr.) U.S. Pat. 3764865, October 9, 1973.

Method of preventing inoperation of a component by loose particles of materials — G.S. Gordon (AED, Pr.) U.S. Pat. 3765941, October 16, 1973.

Analog phase tracker — F.B. Griswold (AED, Pr.) U.S. Pat. 3769589, October 30, 1973.

Rate-aided ranging and time-dissemination receiver — R.F. Buntschuh and H.G. Rouland (AED, Pr.) U.S. Pat. 3769589, October 30, 1973.

Apparatus for remotely measuring the temperature and emissivity of a radiant object independent of the object's radiant surface property — W.E. Martin (AED, Pr.) U.S. Pat. 3770973, November 6, 1973.

Missile & Surface Radar Division

Waveguide assembly including a toroid having dielectric material therein — R.J. Mason and N.R. Landry (MSRD, Moorestown) U.S. Pat. 3768040, October 23, 1973.

Advanced Technology Laboratories

Optical data transmission system employing polarization-shift, multiple-cavity laser — D.G. Herzog and C.W. Reno (ATL, Cam.) U.S. Pat. 3766393, October 16, 1973.

Multifunction full adder — R. L. Pryor (ATL, Cam.) U.S. Pat. 3767906, October 23, 1973.

Article identification apparatus — J.F. Schanne, F.G. Nickl (ATL, Cam.) U.S. Pat. 3770944, November 6, 1973.

Communications Systems Division

Circularly polarized antenna — O. Ben-Dov (G&CS Comm. Sys, Gibbsboro) U.S. Pat. 3765023, October 9, 1973.

Film handling apparatus — B.F. Floden (G&CS Comm. Sys., Cam.) U.S. Pat. 3765586, October 16, 1973.

Automatic iris control — S.L. Bendell and C.A. Johnson (G&CS Comm. Sys., Cam.).

Universal data quality — J.G. Butler and A. Mack (GCS, Cam.) U.S. Pat. 3771059, November 6, 1973.

Palm Beach Division

Self-cleaning read head — E.H. Del Rio and D.F. Pothier (PBD, Palm Beach Gardens) U.S. Pat. 3763352, October 2, 1973.

Laboratories

Redundant Fraunhofer Recording System— T. E. J. Gayeski (Labs, Pr.) U.S. Pat. 3749469, July 31, 1973

Fabrication of Semiconductor Devices — E.J. Boleky, 111, J. H. Scott, Jr. (Labs., Pr.) U.S. Pat. 3749614, July 31, 1973

Method of Fabricating Transparent Conductors — J. L. Vossen, Jr. Labs., Pr.) U.S. Pat. 3749658, July 31, 1973

Electroacoustic Semiconductor Device Employing an IGFET — C. W. Benyon, Jr., D. Leibowitz (Labs, Pr.) U.S. Pat. 3749984, July 31, 1973

Semiconductor Assembly — G. A. Swartz (Labs., Pr.) U. S. Pat. 3753053, August 14, 1973

Double Pulse Bias Stabilization of a Microwave Oscillator Using an Avalanche Diode Operative in the Anomalous Mode — S. Liu, J. J. Risko (Labs., Pr.) U.S. Pat. 3753153, August 14, 1973

Array of Devices Responsive to Differential Light Signals — J. A. Rajchman (Labs., Pr.) U. S. Pat. 3753153, August 14, 1973

Array of Devices Responsive to Differential Light Signals — J. A. Rajchman (Labs., Pr.) U. S. Pat. 3753247, August 14, 1973

Chemical Polishing of Sapphire — P. H. Robinson, R. O. Wance (Labs., Pr.) U. S. Pat. 3753775, August 21, 1973

Method of Depositing Epitaxial Semiconductor Layers from the Liquid Phase — H. F. Lockwood, D. P. Marinelli (Labs., Pr.) U. S. Pat. 3753801, August 21, 1973

Method of Repairing or Depositing a Pattern of Metal Plated Areas on an Insulating Substrate — N. Feldstein, H. B. Law (Labs., Pr.) U. S. Pat. 3753816, August 21, 1973

Electronic Security System — Y. Matsumoto, Y. Kuwahara (Labs. Eng., Tokyo, Japan) U. S. Pat. 3754214, August 21, 1973

Method for Depositing an Epitaxial Semiconductive Layer from the Liquid Phase — H. P. Kleinknecht, A. H. Oberholzer (Labs., Zurich, Switz.) U. S. Pat. 3755011, August 28, 1973

Method of Providing a Semiconductor Body with Piezoelectric Properties — H. F. Lockwood (Labs., Pr.) U. S. Pat. 3755671, August 28, 1973

Glass Encapsulated Semiconductor Device — W. Kern (Labs., Pr.) U. S. Pat. 3755720, August 28, 1973

Coarse Pinhole Array for Recording Improved Redundant Holograms — E. C. Fox (Labs., Pr.) U. S. Pat 3756684, September 4, 1973

Method for Controlling the Composition of a Deposited Film — D. Leibowitz, D. M. Hoffman (Labs., Pr.) U. S. Pat. 3756847, September 4, 1973

Multilayer Circuit Board Techniques — R. J. Ryan (Labs., Pr.) U. S. Pat. 37566891, September 4, 1973

Input Circuits for Charged-Coupled Circuits — W. F. Kosonocky (Labs., Pr.) U. S. Pat. 3760202, September 18, 1973

Internally-Modulated Gas Laser — K. G. Hernqvist (Labs., Pr.) U. S. Pat. 3760296, September 18, 1973

Semiconductor Memory Using Variable Threshold Transistors — J. R. Burns (Labs., Pr.) U.S. Pat. 3760378

Surface strip transmission line and microwave devices using same — C.P. Wen (Labs, Pr.) U.S. Pat. RE27755, September 11, 1973.

Method of manufacturing holographic replicas — J.R. Frattarola (Labs, Pr.) U.S. Pat. 3758649, September 11, 1973.

Charge-coupled shift registers — W.F. Kosonocky (Labs, Pr.) U.S. Pat. 3758794 September 11, 1973.

Method of forming region of a desired conductivity type in the surface of a semiconductor body — H. Kressel, H. Nelson (Labs, Pr.) U.S. Pat. 3762968, October 2, 1973.

Reduction of non-linear spatial distortion in holographic system employing different wavelengths for recording and playback — R.A. Bartolini and D. Karlsons (Labs, Pr.) U.S. Pat. 3763311, October 2, 1973.

Linearity correction circuit for a cathode ray tube — H.E. White (Labs, Pr.) U.S. Pat. 3763393, October 2, 1973.

Digital and analog data handling devices — P.K. Weimer (Labs, Pr.) U.S. Pat. 3763480, October 2, 1973.

Semiconductor devices having closely spaced contacts — L.S. Napoli, W.F. Reichert (Labs, Pr.) U.S. Pat. 3764865, October 9, 1973.

Lattice network using distributed impedance transmission lines — T. H. Campbell, P. Schnitzler, L.J. West (Micro. Elec. Som.) U.S. Pat. 3768047, October 23, 1973.

Method of making MOS transistors — P.E. Norris, J.M. Shaw (Labs, Pr.) U.S. Pat. 3766637, October 23, 1973.

Enhanced readout of stored holograms — R.S. Mezrich (Labs, Pr.) U.S. Pat. 3767285, October 23, 1973.

Direct print-out photographic optical recording media comprising a rhodamine dye — S.E. Harrison, J.E. Goldmacher (Labs, Pr.) U.S. Pat. 3767408, October 23, 1973.

Method for fabrication of polycrystalline films — R.D. Larrabee (Labs, Pr.) U.S. Pat. 3767462, October 23, 1973.

Method for epitaxially growing layers of a semiconductor material from the liquid phase — M. Ettenberg and V.M. Cannuli (Labs, Pr.) U.S. Pat. 3767481, October 23, 1973.

Radiation resistant lithium ferrite cores — P.K. Baltzer (Labs, Pr.) U.S. Pat. 3767581, October 23, 1973.

Transmission line using a pair of staggered broad metal strips — L.J. West (Labs, Som.) U.S. Pat. 3769617, October 30, 1973.

Electronic Components

Method for photoexposing a coated sheet prior to etching — J.J. Moscony and R.L. Kennard (EC, Lanc.) U.S. Pat. 3751250, August 7, 1973.

Cathode-ray tube including a glass envelope with two spaced external conductive coatings and a connecting strip of a third external conductive coating thereon — A.J. Torre (EC, Lanc.) U.S. Pat. 3746904, July 17, 1973.

Coating molybdenum with pure gold — R.L. Buttle (EC, Hrsn.) U.S. Pat. 3741735, June 26, 1973; Assigned to U.S. Government.

Cathode-ray-tube-yoke combination with at least two spaced bodies of organic thermoplastic material therebetween and a method of making said combination — S.B. Deal (EC, Lanc.) U.S. Pat. 3764740, October 9, 1973.

Temperature compensation of transferred electron amplifiers — C.L. Upadhyayula, B.S. Perlman (EC, Pr.) U.S. Pat. 3768029, October 23, 1973.

Method of making an electrically-insulating seal between a metal body and a semiconductor device — S.W. Kessler, Jr. and R.F. Keller (EC, Lanc.) U.S. Pat. 3769688, November 6, 1973.

Method for making an image screen structure for an apertured-mask cathode-ray tube using a mask having temporary apertures — H.B. Law (EC, Pr.) U.S. Pat. 3770434, November 6, 1974.

Cathode-ray tube with shadow mask having random web distribution — R.L. Barbin (EC, Lanc.) U.S. Pat. 3766419, October 16, 1973.

Color image reproducing apparatus — T.F. Simpson (EC, Lanc.) U.S. Pat. 3767845, October 23, 1973.

Solid State Division

Peak detector circuit — R.C. Heuner, G.W. Steudel (SSD, Som.) U.S. Pat. 3758792, September 11, 1973.

Push-pull darlington amplifier with turn-off compensation — J.F. Alves, III (SSD, Som.) U.S. Pat. 3764929, October 9, 1973.

Method of making beam leads for semiconductor devices — T.G. Athanas and A.A. Anastasio (SSD, Som.) U.S. Pat. 3765970, October 16, 1973.

Consumer Electronics

Wide-angle deflection system — P.C. Tang (CE, Indpls.) U.S. Pat. 3758814, September 11, 1973.

Automatic record changer — J.A. Tourtellot (CE, Indpls.) U.S. Pat. 3762723, October 2, 1973.

Blanking circuits for television receivers — M.N. Norman (CE, Indpls.) U.S. Pat. 3763315.

Circuit for transmitting digital signals to conventional television receiver — D.J. Carlson and J.B. George (CE, Indpls.) U.S. Pat. 3766313, October 16, 1973.

Slant track rotating head recorder-reproducer system for selective retention of special information — H.R. Warren (CE, Indpls.) U.S. Pat. 3766328, October 16, 1973.

High voltage hold-down circuit — J.J. Mcardle, R.L. Rauck (CE, Indpls.) U.S. Pat. 3767963, October 23, 1973.

Instant-on circuitry for ac/dc television receivers — D. W. Luz (CE, Indpls.) U.S. Pat. 3767967, October 23, 1973.

Delay of video amplifier dc bias change to accomodate rise/fall of kinescope high voltage after turn on/off of receiver — J. Stark, Jr. and R.J. Gries (CE, Indpls.) U.S. Pat. 3767854, October 23, 1973.

Direct-coupled triggered flip-flop — S.A. Streckler (CE, Som.) U.S. Pat. 3767943, October 23, 1973.

High voltage regulator — P.R. Ahrens (CE, Indpls.) U.S. Pat. 3767960, October 23, 1973.

Blanking circuits for television receivers — T.W. Burrus (CE, Indpls.) U.S. Pat. RE27793, October 30, 1973.

Remote control system for a television receiver — L.B. Juroff (CE, Indpls.) U.S. Pat. 3769588, October 30, 1973.

Metal mask screen for screen-printing — P.J. Griffin (CE, Indpls.) U.S. Pat. 3769908, November 6, 1973.

RCA Limited

Interference suppression circuits — M.J. Shilling, E. Peak (Ltd., England) U.S. Pat. 3763395, October 2, 1973.

Inteference suppression circuits — M.J. Shilling (Ltd., England) U.S. Pat. 3763396, October 2, 1973.

Wideband hybrid system — P. Foldes (Ltd., Montreal, Canada) U.S. Pat. 3768043, October 23, 1973.

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APRIL 28-MAY 1, 1974 — **Cost Effectiveness in the Environmental Sciences,** IES, Shoreham-Americana, Washington, DC. **Deadline info:** (ms) 2/15/74 **to:** Technical Program Committee, Institute of Environmental Sciences, 940 East Northwest Highway, Mt. Prospect, IL 60056.

JUNE 10-12, 1974 — **Power Electronics Specialists Conf.,** IEEE (S-AES), Bell Labs., Murray Hill, N.J. **Deadline info:** (A&S) 1/30/74 **to** W.E. Newell, Westinghouse Res. Labs., Pittsburgh, PA 15235.

JUNE 10-13, 1974 — **G&AP Symp & USNC/URSI Meeting,** G-AP, URSI, Ga. Inst. of Tech., Atlanta, GA. **Deadline info:** (ms) 2/4/74 **to** R. C. Johnson, Ga. Inst. of Tech., Atlanta, GA 30332.

JUNE 12-14, 1974 — **Int'l Microwave Symposium,** IEEE (G-MTT) Atlanta, Ga. **Deadline info:** (A&S) 2/4/74 **to** Dr. Richard C. Johnson, Engineering Experiment Station, Georgia Institute of Technology, Atlanta, GA 30332.

JUNE 11-13, 1974 — **USNC/URSI Meeting,** IEEE (G-MTT) Atlanta, Ga. **Deadline info:** (A&S) 2/4/74 **to** Dr. Richard C. Johnson, Engineering Experiment Station, Georgia Institute of Technology, Atlanta, GA 30332.

JULY 14-19, 1974 — **Power Engineering Society Summer Meeting and Energy Resources Conference,** IEEE (S-PE). **Deadline info:** (ms) 2/1/74 **to** IEEE Headquarters, Attn. S. H. Gold, Technical Program Chairman, 345 East 47th Street, New York, N.Y. 10017.

JULY 14-19, 1974 — **LATINCON 74** (IEEE Region 9 et al), Palace Anehmbi San Paulo, Brazil. **Deadline info:** (abst) 1/31/74 **to** Jose Americo Sampaio, Caixa Postal 20806, San Paulo, Brazil.

JULY 15-17, 1974 — **Conference on Computer Graphics and Interactive Techniques,** University of Colorado Computing Center and ACM/SIGGRAPH, Univ. of Colo., Boulder, CO 80302. **Deadline info:** (ms) 5/20/74 **to** Robert L. Schiffman, General Chairman, Computing Center, University of Colorado, Boulder, CO 80302.

AUG. 26-30, 1974 — **Intersociety Energy Conversion Engrg. Conf.,** (IEEE G-ED, S-AES) ASME et al, Jack Tar Hotel, San Francisco, CA **Deadline info:** (abst) 1/14/74 **to** Hsuan Yeh, Univ. of Pa., Sch. of Elec. & Mech. Eng., Phila., PA 19104.

SEPT. 8-12, 1974 — **Jt. Power Generation,** (IEEE S-PE) ASME, ASCE, Deauville Hotel, Miami Beach, FL. **Deadline info:** (ms) 4/26/74 **to** J.J. Heagerty, Gen'l Elec. Co., POB 2830, Los Angeles, CA 90051.

OCT. 27-31, 1974 — **1974 IEEE International Symposium on Information Theory,** Notre Dame, Indianapolis, IN. **Deadline info:** (ms) 6/1/74 **to** R.T. Chien, Coordinated Science Laboratory, University of Illinois, Urbana, IL 61801.

JAN. 75 — **"Integrated Optics and Optical Waveguides" Special Issue of the IEEE Transactions on Microwave Theory and Techniques,** G-MTT. **Deadline info:** (papers) in triplicate 4/1/74 **to:** D. Marcuse, Bell Laboratories, Crawford Hill Lab., Box 400, Holmdel, N.J. 07733.

Dates of upcoming meetings
—plan ahead

Ed. note: Meetings are listed chronologically. Listed after the meeting title (in bold type) are the sponsor(s), the location, and the person to contact for more information.

JAN. 27-FEB. 1, 1974 — **Power Engineering Society Winter Meeting,** IEEE (S-PE), Statler Hilton Hotel, New York, N.Y. **Prog info:** J. W. Bean, Am. Elec. Pwr., Service Corp., 2 Broadway, New York, N.Y. 10004.

JAN. 28-FEB. 2, 1974 — **IEEE India Section Conv. & Exhibition,** IEEE India Section, Patkar Hall, Sunderbai Hall, Bombay, India. **Prog. info:** G.V. Desai, 253, A-Z, Induatrial Estate, Fergusson Road, Bombay-400 013 India.

JAN. 29-31, 1974 — **Reliability & Maintainability Symp.,** IEEE (G-R) ASQC et al, Biltmore Hotel, Los Angeles, CA. **Prog info:** C.M. Bird, IBM Corp., 7900 N. Astronaut Blvd., Cape Canaveral, FL 32920.

FEB. 13-15, 1974 — **Int'l Solid State Circuits Conf.,** IEEE (SSC Council, Phila. Section) Univ. of Penna., Marriott Hotel, Phila., PA. **Prog info:** H. Sobol, RCA Corp., Princeton, N.J. 08540.

FEB. 20-21, 1974 — **National Conference on Standards for Environmental Improvement,** American National Standards Institute & American Society for Testing and Materials, ASME. **Prog info:** Claude H. Burns, American National Standards Institute, 1430 Broadway, New York, N.Y. 10018.

FEB. 26-28, 1974 — **Computer Conf. (COMPCON),** IEEE (S-C), Jack Tar Hotel, San Francisco, CA **Prog info:** A.F. Hartung, Commercial Sys. Div., 2500 Colorado Ave., Santa Monica, CA 90406.

MAR. 30-APRIL 4, 1974 — **Gas Turbine Conference & Products Show,** ASME, Zurich, Switzerland. **Prog info:** Marion Churchill, ASME, 345 E. 47th Street, New York, N.Y. 10017.

MARCH 12-14, 1974 — **Aerospace & Elec. Sys. Winter Conv. (WINCON),** IEEE (S-ASE) L.A. Council, Marriott Hotel, Los Angeles, CA. **Prog info:** D.A. Hicks, Res. & Tech., 1800 Century Park East, Century City, CA 90067.

MARCH 12-15, 1974 — **Zurich Digital Communications Int'l Seminar,** IEEE (Switzerland Section, G-AE, S-C, S-COMM, et al). **Prog info:** W. Guggenguehl, Institut fur Fernmeldetechnik ETH, Sternwartstrasse 7 CH-8006 Zurich, Switzerland.

MARCH 25-29, 1974 — **IEEE International Convention (INTERCON),** IEEE, Coliseum & Statler Hilton Hotel, New York, N.Y. **Prog info:** J.H. Schumacher, IEEE, 345 E. 47th St., New York, N.Y. 10017.

APRIL 1-4, 1974 — **Design Engineering Conference & Show,** ASME, Chicago, Ill. **Prog info:** A.B. Conlin, ASME, 345 E. 47th Street, New York, N.Y. 10017.

APRIL 2-4, 1974 — **Joint 1974 Railroad Conference (Rail Transportation—Energy Crisis Superstar),** Land Transportation Committee of the IEEE (Industry and Gen. App. Group), Rail Transportation Div. of the ASME, — Hilton Hotel, Pittsburgh, PA. **Prog info:** Paul Drummond, Mgr., Industry Dept. ASME, 345 E. 47th St., New York, N.Y. 10017 also E.K. Farrelly, Port Authority of N.Y. & N.J., World Trade Ctr., New York, N.Y. 10047.

APRIL 1-5, 1974 — **IEEE Power Engrg. Society Underground Transmission & Distribution Conf.,** IEEE (S-PE) Dallas Conv. Ctr., Dallas, TX. **Prog info:** N.E. Piccione, L.I. Lighting Co., 175 E. Old Country Rd., Hicksville, N.Y. 11801.

APRIL 2-4, 1974 — **Reliability Physics Symposium,** IEEE (G-ED, G-R), MGM Grand, Las Vegas, NV. **Prog info:** I.A. Lesk, Motorola Inc., 5005 E. McDowell Rd., Phoenix, AZ 85008.

APRIL 3, 1974 — **Minicomputers - Trends and Applications,** IEEE (S-C), Nat'l. Bureau of Standards, Gaithersburg, MD. **Prog info:** Harry Hayman, 738 Whitaker Terrace, Silver Spring, MD 20901.

APRIL 8-11, 1974 — **Computer Aided Design Int'l Conf. & Exhibition,** Inst. of Civil Engrs., IERE, IEEE UKRI Sec. et al., Univ. of Southampton, Southampton, England. **Prog. info:** Inst. of Civil Engrgs., Great George St., Westminster, London SW 1, UK.

APRIL 9-11, 1974 — **Optical Computing Symposium,** IEEE (S-C), Aurich, Switzerland. **Prog. info:** David Casasent, Carnegie-Mellon Univ., Dept. of EE, Pittsburgh, PA 15213.

APRIL 16-18, 1974 — **Optical & Acoustical Micro-Electronics,** IEEE (G-MTT, G-SU, PIB et al), Commodore Hotel, New York, N.Y. **Prog info:** PIB, MRI Symp. Comm., 333 Jay St., Brooklyn, N.Y. 11201.

APRIL 17-19, 1974 — **Structures, Structural Dynamics & Material Conference,** ASME, AIAA, Las Vegas, NV. **Prog info:** P. Drummond, ASME, 345 E. 47th St., New York, N.Y. 10017.

APRIL 16-18, 1974 — **"Optical and Acoustical Micro-Electronics"** MRI International Symposium XXIII, Hotel Commodore, New York, N.Y. **Prog info:** Jerome Fox, Executive Secretary, Polytechnic Institute of New York, MRI Symposium Committee, 333 Jay Street, Brooklyn, N.Y. 11201.

APRIL 21-24, 1974 — **Int'l Circuits & Systems Symp.,** IEEE (S-CAS), Sir Francis Drake Hotel, San Francisco, CA. **Prog info:** L.O. Chua, Dept. of EE, Univ. of Calif. at Berkeley, Ca 94720.

APRIL 22-24, 1974 — **Communications Satellite Sys. Conference,** IEEE (S-AES), AIAA, Int'l. Hotel, Los Angeles, CA **Prog info:** Dave Lipke, Comm. Satellite Corp., 950 L'Enfant Pl., S., S.W., Washington, DC 20024.

APRIL 28-MAY 2, 1974 — **Diesel & Gas Engine Power Conference & Exhibit,** ASME, Houston, TX. **Prog info:** Marion Churchill, ASME, 345 E. 47th St., New York, N.Y. 10017.

APRIL 29-MAY 1, 1974— **American Power Conference (IIT),** Chicago, IL. **Prog info:** ASME, 345 E. 47th St., New York, N.Y. 10017.

MAY 5-8, 1974 — **Offshore Technology Conference,** AIME, ASME, and others, Houston, TX. **Prog info:** ASME, 345 E. 47th St., New York, N.Y. 10017.

MAY 12-15, 1974 — **National Incinerator Conference and Exhibit,** ASME, Carillon Hotel, Miami Beach, FL. **Prog info:** Mr. Maurice Jones, Information Services, ASME, 345 E. 47th Street, New York, N.Y. 10017.

MAY 13-15, 1974 — **Joint Fluids Engineering & CSME Conference,** ASME, Montreal, Quebec, Canada. **Prog info:** ASME, 345 East 47th Street, New York, N.Y.

May 21-23, 1974 — **Intersociety Material Handling Symposium & MHI Show,** ASME, Detroit, MI. **Prog info:** ASME, 345 E. 47th St., New York, N.Y. 10017.

JUNE 3-7, 1974 — **7th US Congress of Theoretical & Applied Mechanics,** ASME, Boulder, CO. **Prog info:** ASME, 345th E. 47th St., New York, N.Y. 10017.

JUNE 10-13, 1974 — **Summer Annual Meeting,** ASME, New Orleans, LA, **Prog info:** ASME, 345th E. 47th St., New York, N.Y. 10017.

SEPT. 15-18, 1974 — **Petroleum Mechanical Engineering Conference,** ASME, Dallas, TX. **Prog info:** ASME, 345th E. 47th St., New York, N.Y. 10017.

SEPT. 22-27, 1974 — **9th World Energy Conference,** Detroit, MI. **Prog info:** ASME 345 E. 47th St., New York, N.Y. 10017.

SEPT. 23-25, 1974 — **4th Urban Technology Conference,** AIAA, ASME, & Others., Chicago, IL. **Prog info:** ASME, 345 E. 47th St., New York, N.Y. 10017.

OCT. 1-3, 1974 — **International Forum for Air Cargo,** SAE, ASME, San Diego, CA. **Prog info:** ASME, 345 E. 47th St., New York, N.Y. 10017.

JUNE 17-19, 1974 — **Lubrication Symposium,** ASME, Key Biscayne, FL. **Prog info:** ASME, 345 E. 47th St., New York, N.Y. 10017.

June 18-21, 1974 — **Joint Automatic Control Conference,** AIChE, ASME, IEEE, AIAA, ISA, Austin, TX. **Prog info:** ASME, 345 E. 47th St., New York, N.Y. 10017.

JUNE 24-28, 1974 — **Pressure Vessels & Piping Conference,** ASME, Miami Beach, FL. **Prog info:** ASME, 345 E. 47th St., New York, N.Y. 10017.

JULY 15-17, 1974 — **Thermophysics and Heat Transfer Conference,** AIAA, ASME, Boston, MA. **Prog info:** ASME, 345 E. 47th St., New York, N.Y. 10017.

JULY 30-AUG. 4, 1974 — **Intersociety Conference on Evnironmental Systems,** SAE, ASME, Seattle, WA. **Prog info:** ASME, 34t E. 47th St., New York, N.Y. 10017.

AUG. 5-10, 1974 — **IFIP Congress 74,** American Federation of Information Processing Societies, Stockholm, Sweden. **Prog info:** U.A. Committee for IFIP Congress 74, Box 426, New Canaan, CT 06840.

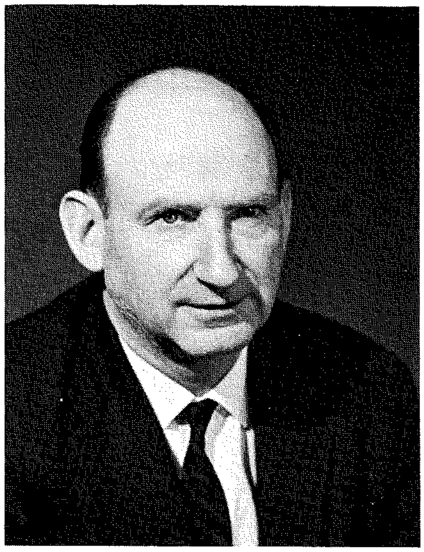
AUG. 21-23, 1974 — **Engineering in the Ocean Environment,** IEEE, Halifax, Nova Scotia, Canada. **Prog info:** Ocean '74, P.O. Box 1000, Halifax, Nova Scotia, Canada.

AUG. 25-30, 1974 — **Intersociety Energy Conversion Engineering Conference,** ASME & Others, San Francisco, CA. **Prog info:** ASME, 345 E. 47th St., New York, N.Y. 10017.

SEPT. 3-7, 1974 — **5th International Heat Transfer Conference,** ASME. **Prog info:** ASME, 345 E. 47th ST., New York, N.Y. 10017.

SEPT. 15-19, 1974 — **Jt. Power Generation,** IEEE, ASME, Miami Beach, FL. **Prog info:** ASME, 345 E. 47th St., New York, N.Y. 10017.

Engineering News and Highlights



Rajchman receives Edison Medal

Dr. Jan A. Rajchman, Staff Vice President, Information Sciences at RCA Laboratories was recently selected to be the recipient of the 1974 Edison Medal "for a creative career in the development of electronic devices and for pioneering work in computer memory systems." The Edison medal is presented by the IEEE in recognition of career achievements in electrical science or electrical engineering or electrical arts.

Dr. Rajchman joined RCA in 1935 and has been engaged in research since that time, first in Camden, N.J., then from 1942 in Princeton, N.J. with RCA Laboratories. In 1959 he became Associate Director, Systems Research Laboratories; in 1961 Director, Computer Research Laboratories; 1967, Staff Vice President, Data Processing Research; 1969, Staff Vice President, Information Sciences. In 1971 the supervision of RCA research laboratories in Zurich and Tokyo was added to his management role in Princeton, where he maintains the direction of an exploratory research group.

Professional Activities

Aerospace Systems Division

Burt Clay is a member of the Board of Governors of the Society of Motion Picture and Television Engineers. He was Chairman, and is now a member of the Program Committee. Mr. Clay is also a member of the council of the New England Section of the Optical Society of America and a member of the Program Committee. In addition, he is a member of the Executive Committee of the Research Society of America (ASD Branch).

Bill Gray has been a member of the Reliability Chapter, Boston Section IEEE for the past 13 years. He served as Technical Program Chairman three years, as Chairman of the Spring Reliability Seminar three years and as Chapter Chairman for one year. He participated in the National Reliability Symposium both as a Moderator for two years and as a member of the technical panel. He has been involved in the Government & Industry Data Exchange Program for the past ten years, during the last two years as RCA representative on the Industry Advisory Group.

Joe Vick Roy is a member of the National Administration Committee for the IEEE Engineering Management Society and is a past Chairman of the Boston Chapter of the IEEE Engineering Management Society. He is currently the chairman of the Management Society Technology Forecasting and Assessment Committee. This committee is currently preparing a forecast of engineering management techniques.

Electromagnetic and Aviation Systems Division

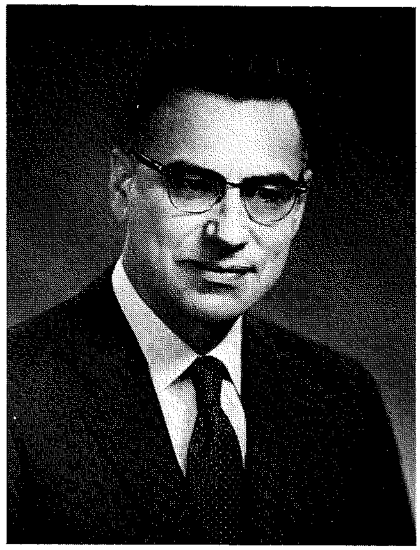
James R. Hall, Advanced Project Development Manager, is chairman of the San Fernando Valley Section of IEEE.

Engineering Staff Technical Advisor **Ramon H. Aires** recently accepted the IEEE Sixth Region Company-of-the-Year award on behalf of EASD. Presented at WESCON, the plaque acknowledged the Divison's contributions and services to the IEEE and to the Engineering profession.

Nominated by the San Fernando Valley Section for this award from the 16-western states region, EASD had already won recognition in June when Division Vice President and General Manager **Frederick H. Krantz** accepted the Company-of-the-Year award from IEEE's Los Angeles Council. The Los Angeles Council includes the area from Santa Barbara to San Diego.

Communication Systems Divison

The Government Communications Systems activity was well represented at the National Telemetering Conference held in late November 1973. **J.M. Osborne,** Divison Vice President, was co-organizer for two sessions; **A. Mack** was chairman of one session; and **E.D. Taylor, L. Kolodin, H. Hovagimyan, E.G. Tyndall,** and **P.J. Bird** presented papers.



H.J. Woll receives University of Pennsylvania Distinguished Alumnus Award

Dr. Harry J. Woll, Division Vice President Government Engineering, has been awarded a Moore School 50th year Gold Medal by the University of Pennsylvania as a distinguished alumnus. The University's Moore School of Electrical Engineering is celebrating its 50th anniversary this year, and special gold medals were struck in commemoration of the event.

Dr. Woll received the award in recognition of his "distinctive contribution to engineering and society as a Moore School Alumnus." The presentation was made at the annual Engineering Alumni Society Dinner held at the University. Dr. Woll received his Ph.D. from the University of Pennsylvania in 1953. He is a member of several major engineering societies and is Fellow of the Institute of Electrical and Electronics Engineers.

Brindley named coordinator for patents and new technology

D.M. Cottler, Chief Engineer, Missile and Surface Radar Divison recently appointed **C.M. Brindley** as coordinator of patents and new technology for MSRD. Anyone at MSRD who produces a patentable invention or is involved in development of new technology should forward disclosure material to Charlie Brindley (Mail Stop 108-213, PM 3978), who will arrange for logging the item and forwarding the material to RCA Patent Operations at Princeton.

Bachynski wins top Quebec award

Dr. Morrel P. Bachynski, Director, Research and Development, RCA Limited has been awarded the 1973 *Prix Scientifique du Quebec*, presented every fourth year for achievement in the physical sciences and related fields by the Province's Department of Cultural Affairs.

Dr. Bachynski, won the $5000 prize for the advances he has contributed to the field of high energy plasmas and their interaction with electromagnetic waves.

Dr. Bachynski, who joined RCA Limited in 1955 shortly after obtaining his doctorate in physics from McGill University, has been the company's director of research since 1965. In that position he has led research into electromagnetic wave propagation as well as microwave, laser and plasma physics. He also supervises RCA Ltd's communications and space technology laboratory, digital information technology laboratory, semiconductor laboratory, and its physical electronics laboratory.

Because of his broad experience in both laser and plasma physics, Dr. Bachynski has been an active member of the Quebec consortium which was recently awarded a contract by the federal government to determine a possible role for Canadian science in the international development of controlled thermo-nuclear energy for peaceful purposes.

Dr. Bachynski enjoys an international reputation as a scientist and has been awarded many distinctions including the David Sarnoff award for "Outstanding Individual Achievement". He is a Fellow of several learned societies, including the Royal Society of Canada and the American Physical Society and is a distinguished member of five others. He also serves on a number of university advisory boards and is a past president of the Canadian Association of Physicists.

Staff announcements

Chairman of the Board and Chief Executive Officer

Robert W. Sarnoff, Chairman of the Board and Chief Executive Officer, has announced the following organization changes in RCA Staff:

George C. Evanoff is appointed Vice President, Corporate Development, and will report to the Chairman of the Board and Chief Executive Officer. The Corporate Development function will be responsible for the formulation of the Company's long-range objectives and the development of business strategies to meet these objectives. It will include responsibility for Corporate Strategic Planning, Venture Planning, International Planning and other major activities having a significant impact on Corporate performance.

Charles C. Ellis as Senior Vice President, Finance will report to **Anthony L. Conrad,** President and Chief Operating Officer. The Finance organization, in addition to its present responsibilities, will assume the responsibility for Businesss Planning.

President and Chief Operating Officer

Anthony L. Conrad, President and Chief Operating Officer has announced the Marketing organization as follows: **Joseph W. Curran** as Staff Vice President, Marketing Services; **Edward J. Homer** as Director, Marketing Administration; **Robert W. Redecker** as Staff Vice President, Distributor Relations.

Corporate Development

George C. Evanoff, Vice President, Corporate Development has announced the organization of Corporate Development as follows: **Peter B. Jones** as Staff Vice President, Business Development; **Joseph V. Quigley** as Staff Vice President, Strategy Development; **Robert J. Eggert** as Staff Vice President, Economic and Industry Research; **Eugene J. Dailey** as Staff Vice President, International Planning; **Richard W. Sonnenfeldt** as Staff Vice President, New Business Programs; **Holmes Bailey** as Director, Consumer Information Systems Development; **Thomas T. Callahan** as Director, Systems Development; **George C. Evanoff** as Acting "SelectaVision"; **John F. Biewener** as Staff Vice President, "SelectaVision" Business Planning and Control; **Donald P. Dickson** as Staff Vice President, "SelectaVision" Programming Distribution; **Thomas J. McDermott** as Staff Vice President, "SelectaVision" Programming Development.

Engineering

Howard Rosenthal, Staff Vice President, has announced the organization of Engineering as follows: **Arnold S. Farber,** Staff Engineer; **Russel G. Groshans,** Staff Engineer; **Edwin M. Hinsdale,** Staff Engineer; **Doris E. Hutchison,** Administrator, Staff Services; **Harry Kleinberg,** Manager, Corporate Standards Engineering; **Eric M. Leyton,** Staff Engineer; **Arthur Sherman,** Staff Engineer; **Raymond E. Simonds,** Director, RCA Frequency Bureau; **William J. Underwood,** Manager, Engineering Professional Programs; and **Frank W. Widmann,** Staff Engineer.

Finance

Charles C. Ellis, Senior Vice President, Finance has appointed **Franz Edelman,** Director, Operations Research.

International

Eugene A. Sekulow, Vice President, International has reported that the following activities are transferred to the RCA Staff International organization: **Ralph E. Bates** as Staff Vice President, International Distributor Relations and Services; **William L. Newell** as Director, International Trade Policy Administration.

Manufacturing Services and Materials

George A. Fadler, Vice President, Manufacturing Services and Materials, has announced the appointment of **Robert T. Vaughan** as Staff Vice President, Manufacturing.

Robert T. Vaughan, Staff Vice President, Manufacturing has announced the Manufacturing organization as follows: **James L. Miller,** Director, Manufacturing Systems and Technology; **H. Robert Snow** continues as Director, Industrial and Manufacturing Engineering.

International

Eugene A. Sekulow, Vice President, International announced that the Board of Directors of *RCA Electronica Limitada* (Brazil) elected **Manuel DeArmas,** President.

Robert A. Schieber, Division Vice President, International has announced the appointment of **John M. Watkins** as General Manager, Consumer Electronics Division, *RCA Electronica Ltda.* (Brazil).

Electronic Components

Lucien DeBacker, Manager, Market Planning, Power Products has announced the new marketing organization as follows: **Ronald M. Bowes,** Manager, Market Planning, Regular Power Devices; **Edward D. Fleckenstein,** Manager, Market Planning, Large Power Devices; **Phillip H. Vokrot,** Manager, Market Planning, Lasers; **Lucien DeBacker,** Acting Manager, Market Planning, Microwave Devices. Reporting to Mr. DeBacker in his position as Acting Manager, Market Planning, Microwave Devices are: **Herbert Berkowitz,** Manager, Market Planning, Research and Devleopment; **Otto Johnk,** Manager, Market Planning, Solid State and Pencil Tubes; and **Joseph J. Snack,** Administrator, Product Coordination and Control.

David D. VanOrmer, Manager, Picture Tube Development, Engineering, Entertainment Tube Division, announced the organization of the Picture Tube Development Engineering Activity as follows: **Austin E. Hardy,** Manager, Chemical & Physical Laboratory; **Albert M. Morrell,** Manager, Design Laboratory & Engineering Standards; **Richard A. Nolan,** Manager, Pilot Development Engineering; and **William J. Schnelli,** Manager, Pilot Production Center.

Richard H. Hynicka, Plant Manager, Lancaster Color Picture Tube Plant, has announced the organization of the Lancaster Color Picture Tube Plant as follows: **William J. Harrington,** Manager, Planning & Controls; **Richard E. Myers,** Manager, Manufacturing; **Richard L. Spalding,** Manager, Quality and Reliability Assurance; **Yoneichi Uyeda,** Manager, Production Engineering; and **Willaim G. Weisser,** Manager, Production & Material Control.

Robert C. Pontz, Manager, Phototube & Solid State Opto-Electronics Operation, Industrial Tube Division, has announced the appointments of **Richard Glicksman** as Manager, Product Development (Solid-State Opto-Electronics Operation); **Andrew G. Zourides** as Manager, Manufacturing & Production Engineering (Solid-State Opto-Electronics Operation); and **Wayne E. Rohland,** Manager, Manufacturing — Phototube.

Charles W. Thierfelder, Divison Vice President, Manufacturing, Entertainment Tube Division, Electronic Components has announed the appointment of **John M. Fanale** as Director, Glass Operations at Circleville, Ohio.

Joseph H. Colgrove, Acting Manager, International Operations, Entertainment Tube Division, Electronic Components has announced the appointment of **Fred A. Daud** as Manager, International Coordination, Latin America; and **Robert H. Handler** as Manager, International Coordination, Far East.

Consumer and Solid State Electronics

Bernard V. Vonderschmitt, Vice President and General Manager, Solid State Division, has announced the appointment of **Richard J. Hall** as General Manager of the newly formed subsidiary company, *RCA Sendirian Berhad* (Malaysia).

Roy H. Pollack, Division Vice President and General Manager, Color and Black & White Television Division announced his organization as follows: **Harry Anderson,** Director, Manufacturing Operations; **David E. Daly,** Divison Vice President, Advanced Produce Planning; **Loren R. Kirkwood,** Director, Color TV Engineering; **William S. Lowry,** Division Vice President, Product Management, Color TV; **Tucker P. Madawick,** Division Vice President, Industrial Design; **Richard Mentzinger,** Director, Quality and Reliability.

RCA Records

Howard R. Hawkins, Executive Vice President RCA Corporation announced the appointment of **Kenneth D. Glancy** as President and Chief Executive Officer of RCA Records.

In his new position, Mr. Glancy succeeds **Rocco M. Laginestra** who will continue in an executive capacity within the RCA Corporate Staff.

Industrial Relations

James J. Brant, Staff Vice President, Industrial Relations, International and New Business has announced the appointment of **Frank A. Tylius** as Director, International Industrial Relations Administration.

Electromagnetic and Aviation Systems Division

Frederick H. Krantz, Divison Vice President and General Manager has announced the appointment of **Robert M. Hinkel** as Plant Manager for Electromagnetic and Aviation Systems Division, Van Nuys, Calif.

Awards

Aerospace Systems Division

Peter Nesbeda received a Technical Excellence Award for September in recognition of his work on the Wells Fargo software design and data acquisition subsystem.

Lionel Arlan received a Technical Excellence Engineering Award for October for his role and exemplary professional performance in the design fabrication and test of a specialized surveillance television camera.

The Electronic Quality Assurance Test Equipment (EQUATE) Team of **R.M. Beigel, B.A. Bendel, R.J. Bosselaers, G.A. Bowles, L.K. Dickman, J.E. Fay, E.M. Fisher, N. Utletsky, J.C. Haggis, A.J. Krisciunas, J.F. McGrann, R.W. O'Neill, E.W. Richter, F.A. Schwedner, E.M. Sutphin,** and **A.F. Vallance** received the technical excellence team award for September 1973 for development and implementation of a new concept in Automatic Test Equipment. This new approach to ATE was designed and developed for the Army Electronics Command under contract number DAAB05-71-C-2641.

Communications Systems Division

D.J. Parker, Chief Engineer of Government Communication Systems recently announced the selection of **Dick Noto** of the Advanced Technology Laboratories for a Technical Excellence Award. This extra-divisional award recognizes Mr. Noto's extraordinary contributions to GCS capability in design automation for large-scale integrated-circuit arrays.

Nergaard receives Kelly Award

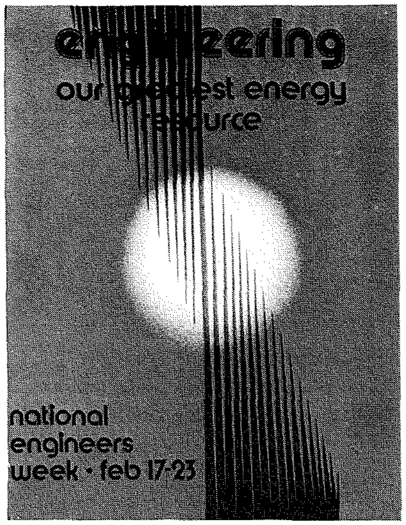
Dr. Leon S. Nergaard, who was Director of the Microwave Research Laboratory at RCA Laboratories before his retirement in 1971, will receive the Mervin J. Kelly Award "for outstanding contributions and leadership in the introduction of very high frequencies for telecommunications." The Kelly Award is presented each year by the IEEE for outstanding contribution in the field of telecommunication.

Dr. Nergaard has worked in research for RCA since 1933; he still serves as a consultant. He received the BSEE from the University of Minnesota and the MSEE from Union College in Schenectady, N.Y. He returned to the University of Minnesota and earned the PhD in Physics. A fellow of the American Physical Society and IEEE, Dr. Nergaard is also a member of the American Association for the Advancement of Science. He is a past chairman of the IEEE Professional Group on Electron Devices.

MSRD engineers cited for packaging

John R. Van Horn, Leader, Engineering Support, and **James A. Hill,** Packing Designer, of the Radiation Equipment Engineering Activity, Missile and Surface Radar Divison have received special recognition from the Society of Packaging and Handling Engineers.

Van Horn and Hill received third prize in the Military Packaging Category for their entry in the 28th Annual Packaging and Handling Competition. They designed a container which makes it possible to use only one container where previously several distinct fixtures would have been needed. Their entry combines the features of a test fixture, handling device and shipping container for a high performance uhf antenna being designed by MSRD for use on the Viking Spacecraft that will be launched for a soft landing on Mars during the 1970's.



Engineers week (Feb 17-23) focuses on energy

Recent concern over many aspects of the energy situation is reflected in the theme of National Engineers Week for 1974: "Engineering — Our Greatest Energy Resource."

The week, February 17-23, will emphasize the development and application of creative technology in developing solutions to energy problems.

Licensed engineers

When you receive a professional license, send your name, PE number (and state in which registered), RCA division, location, and telephone number to: *RCA Engineer*, Bldg. 204-2, RCA, Cherry Hill, N.J. As new inputs are received they will be published.

Communications Systems Division

E.J. Nossen, CSD, Camden, N.J.; PE-20601, New Jersey.

Electronic Components announces new line of black-and-white television cameras

RCA is entering the rapidly expanding $100 million closed-circuit video equipment market with a line of black-and-white closed-circuit television (CCTV) cameras.

The announcement, made recently in Hollywood, California, by Electronic Components during its annual sales meeting, pointed out that the new RCA cameras were primarily designed for use in the industrial surveillance, audio/visual, specialty and general CCTV equipment market.

"RCA's intent is to provide the closed-circuit tv equipment user with a camera that has the most desired features, is easy to operate, competitively priced, reliable and backed by service at the manufacturing location," according to Victor C. Houk, Manager of Video Equipment Marketing, RCA Electronic Components. "Initial sales emphasis will be in the domestic market," he noted.

Three new RCA cameras were demonstrated at the 1973 National Audio Visual Association Show in the Miami Beach Convention Center, which was open through January 8th.

Called 'RCA-Americans", the new cameras are RCA designed and will be manufactured at the RCA Electronic Components plant in Lancaster, Pa. They will also utilize RCA vidicon camera tubes.

The major distribution channel for RCA closed-circuit video equipment will be through CCTV dealer/installers. Sales specialists will be assigned throughout the United States to provide technical support to customers.

All of the new RCA cameras feature integrated circuit construction, automatic light compensation (ALC), 2/3-inch RCA vidicon camera tubes, easy to use controls and reliable performance.

The RCA TC1000 camera is a general purpose industrial surveillance camera for use in a wide variety of security applications in prisons, airports, freight docks, stores, banks, apartments, industrial plants and unmanned locations. In addition, the TC1000 can also be used in crowd control, merchandising, and in industrial applications to view remote operations, process control, and meter observations

The TC1000 is extremely flexible with all external interfaces—camera mounts, lens, output, controls—designed for use with a broad variety of systems and with the greatest all-around economy. Technical features include:

- Controlled video response that cuts down smears and confusing white outlines;
- A 10:1 gray scale which permits identification of faces even against difficult, bright backgrounds;
- 550 lines resolution;
- f/l.6 lens; and
- 8,000:1 automatic light compensation for high sensitivity with usable pictures in dim interiors or bright sunlight.

Most adjustments are set at the factory and the camera requires a minimum of attention from the user.

The RCA TC1005 and TC1010 cameras were designed for the more sophisticated education, training, tv production, signature verification, traffic and production control applications which require greater overall performance. The TC1010 has all of the features of the TC1005, plus a 2/3-inch silicon target vidicon camera tube. Both the TC1005 and TC1010 contain automatically adjusting circuits and 40,000:1 automatic light compensation. Resolution is 650 lines and 430 lines, respectively.

The RCA TC1055 and TC1056 cameras, called the "Twins", were primarily designed for use with small video tape recorders, but can also be used in studio systems. They contain a minimum of controls with many automatic circuits, built-in microphone, 3:1 zoom lens, optical split-image viewfinder, pistol-grip handle for ease of use, and high sensitivity with 8,000:1 automatic light compensation.

The TC1055 is self driven from a separate power supply included with the camera. The TC1056 takes drives and power from the video tape recorder. Adapters for use with various recorders now on the market are available with the camera.

Optional user prices on these cameras are: TC1000, $260 each; TC1005, approximately $450 each; TC1055 and TC1056, $375 each. The price of the TC1010 has not been established as yet. The TC1000, TC1055 and TC1056 will be available for delivery starting in March, 1974 and the TC1005 and TC1010 about mid-year.

Additional information on the RCA-Americans tv cameras is available from RCA Closed-Circuit Video Equipment Marketing, New Holland Pike, Lancaster, Pa. 17604.

RCA Review, December 1973
Volume 34, Number 4

Contents

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